ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Drift-Corrected D.C. Amplifier

Uni-Control Wide-Range Oscillator

Waveguide Characteristics

National Radio Exhibition

Three shillings
and sixpence

OCTOBER 1957 Vol 34 *new series* No 10

For continuous use at 130°C

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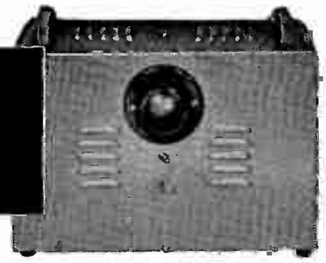
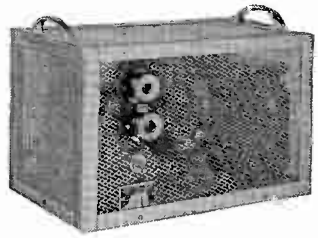
A-C AUTOMATIC VOLTAGE REGULATORS
39 BASIC TYPES IN 6 DESIGN SERIES



BMVR - 1725

WIDEST RANGE IN THE WORLD?

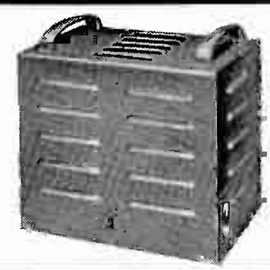
So far as we are aware, our range of A.C. Automatic Voltage Stabilisers is the largest in the World. We have a very wide range of standard models, single-phase patterns ranging from 200 VA to about 30 kVA (3-phase types up to about 90 kVA). There are 39 basic types, in six distinct design series, and all are available in standard form or as tropicalised instruments. We feel that on this account there can be few, if any requirements covering Stabilisers that we are not in a position to meet economically, efficiently and promptly.



BAVR 1000 & BAVR 1000 E

Here are very brief details of the six main series, in handy tabular form: cut this ad. out and use it as a Buying Guide; but please remember that if you do not see *exactly* what you require a written enquiry will probably reveal that we have a "special" to suit, or that the answer is under development. New stabilisers are regularly being added to our range. Several are at the very advanced development stage now—and we do design "specials". One such "special" (AM type 10D/20161) is illustrated (Illustrations not to scale). Nearly 100 have been supplied to Murphy Radio Ltd. for incorporation in equipment supplied by them to the Air Ministry for use on a chain of Radar Marker Beacons. 45 in slightly differing form are currently being made by us for the Air Ministry for another Radar Chain.

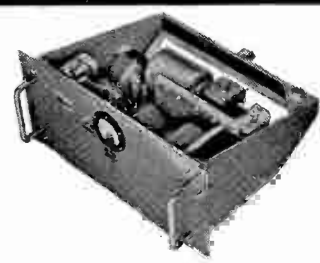
For complete data request our new 32-page illustrated Catalogue–Technical Manual, C.L.L. Form S-574.



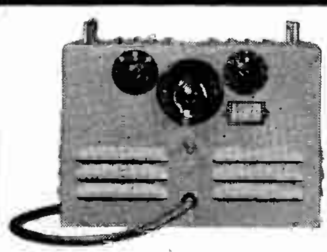
BMVR - 7000 - Series & TCVR - 7000 - Series

DESIGN SERIES	ASR	ATC	BAVR	BAVR-E	BMVR	TCVR
Input Voltage "Swing"	— 10% to + 5%	—20% to + 10%	— 10% to + 5%	— 10% to + 5%	Depends on power: typical is from — 19% to +8.5%	
Output Voltage Stability	± 2½%	± 5%	± 0.15%	± 0.15%	Usually ±0.5%	Usually ±0.5%
Change due to load (0-100%)	NEGLIGIBLE		±2.0%	±0.3%	NIL	NIL
Harmonics Generated	NIL	NIL	YES	YES	NIL	NIL
Response Speed	PRACTICALLY INSTANTANEOUS AVERAGING				I V/Sec.	40V/Sec.
	2-3 CYCLES		I CYCLE			
Power Ratings	1150VA 2300VA	575VA 1150VA	200VA 500VA 1000VA	200VA 500VA 1000VA	1600VA to 30kVA (18 models)	1600VA to 12kVA (11 models)
Basic Prices *	£24 to £34	£24 to £34	£50 to £79	£59 to £88	£75 to £237	£91 to £144

* *From May 1st 1956, subject to 7½% increase.*



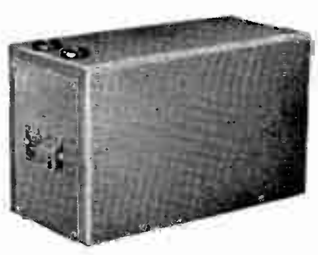
BMVR - 2750 - S58 (AM Ref. 10D 20161)



BMVR - 2750 VV & TCVR - 2750 VV

Claude Lyons Ltd.

STABILISER DIVISION



ASR - 1150 & ATC - 575.

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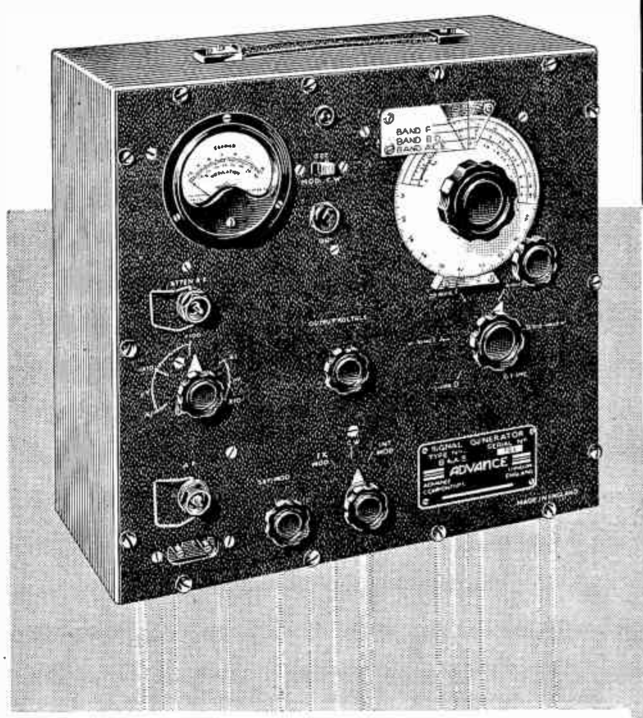

MODEL A 100 kc/s—80 Mc/s in six bands

MODEL B 30 kc/s—30 Mc/s in six bands

Calibration accuracy of both models is ±1%

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Full technical details in Folder R38

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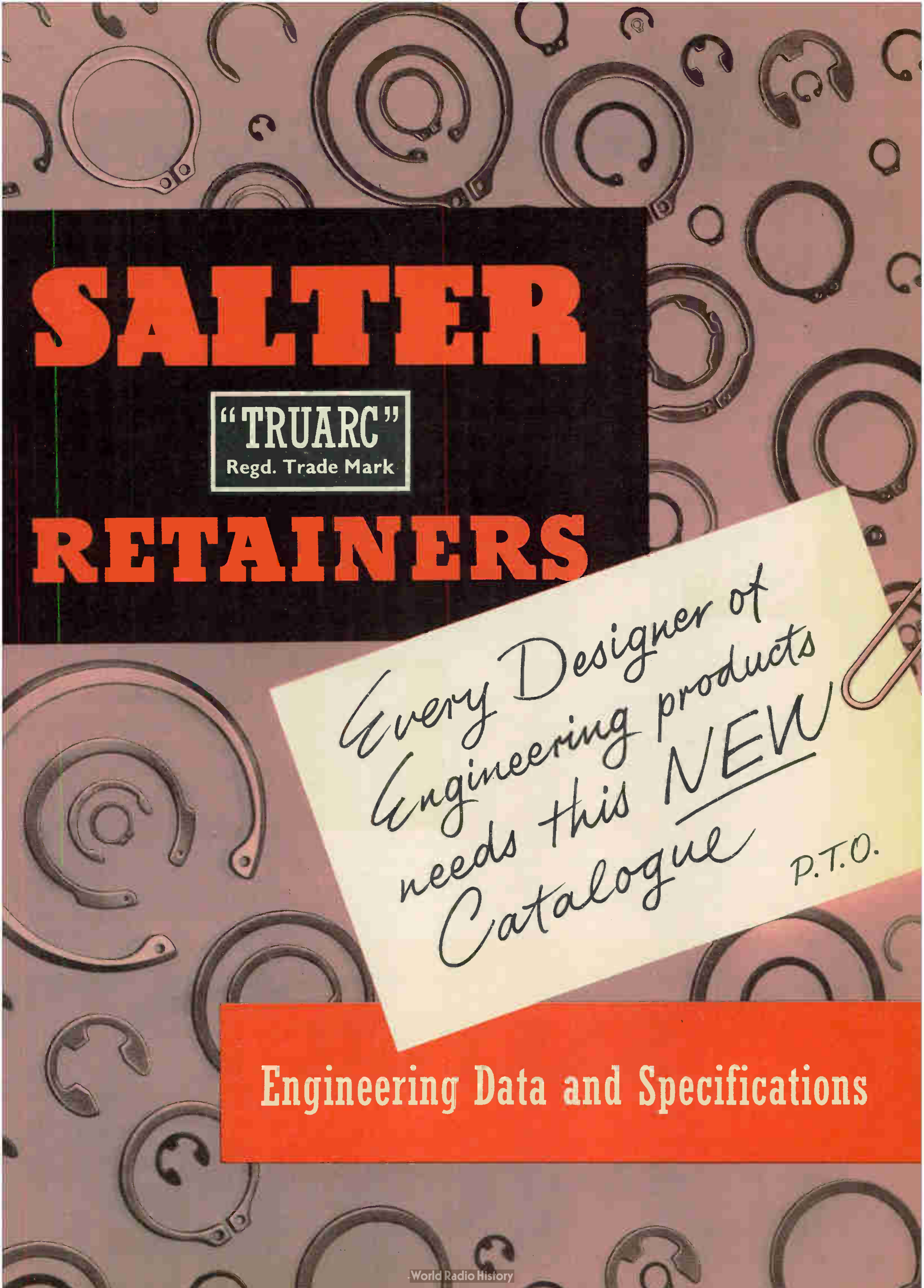
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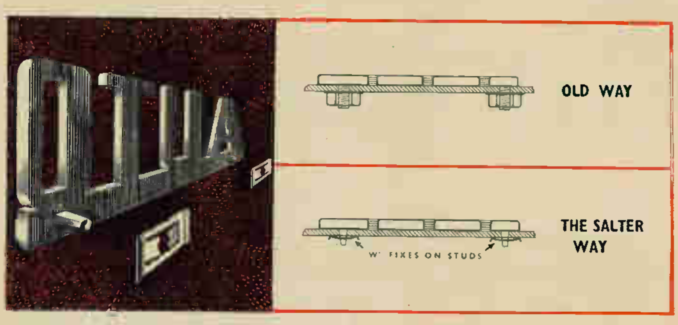

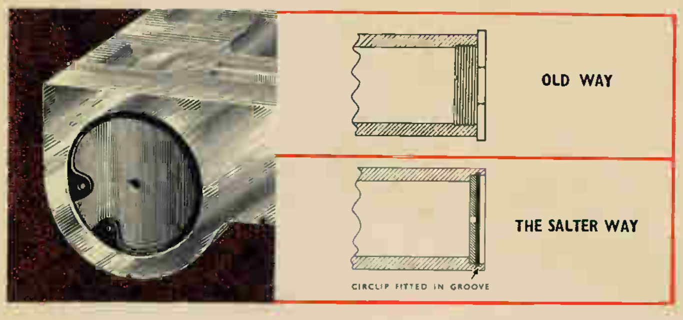

reduce assembly time—

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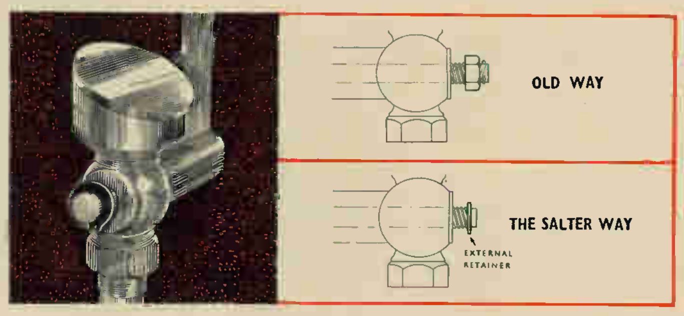

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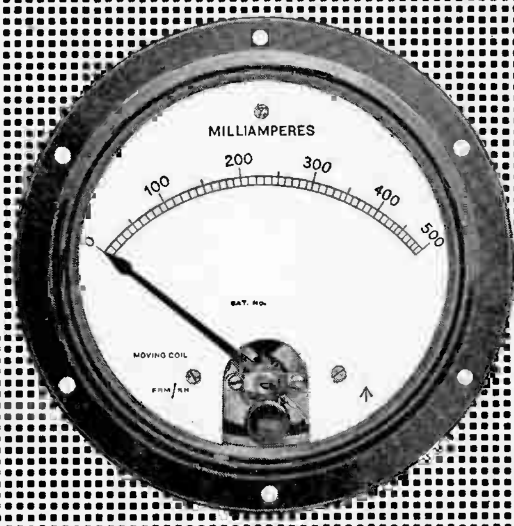

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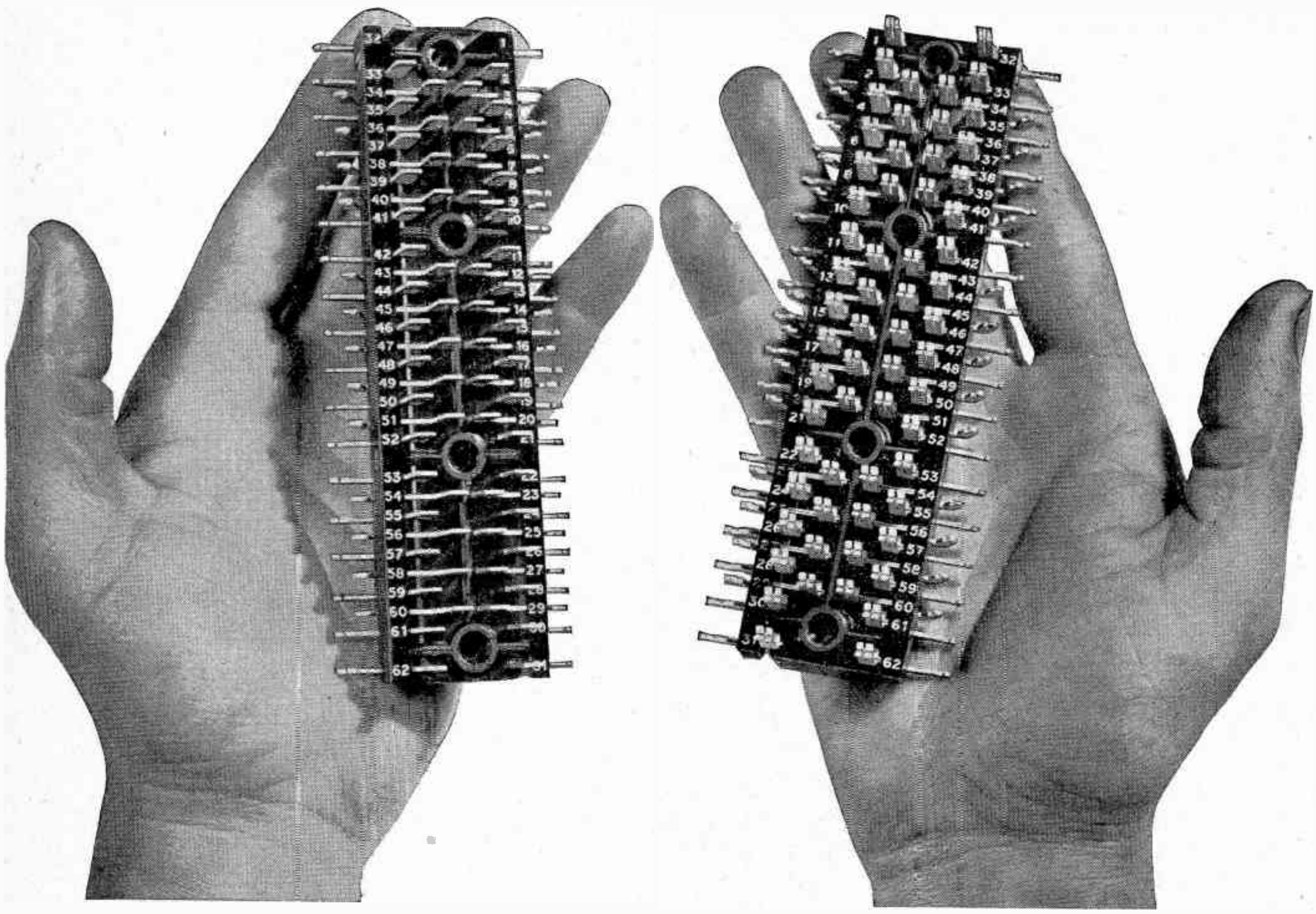

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Withdrawal pressure:	13 lbs.

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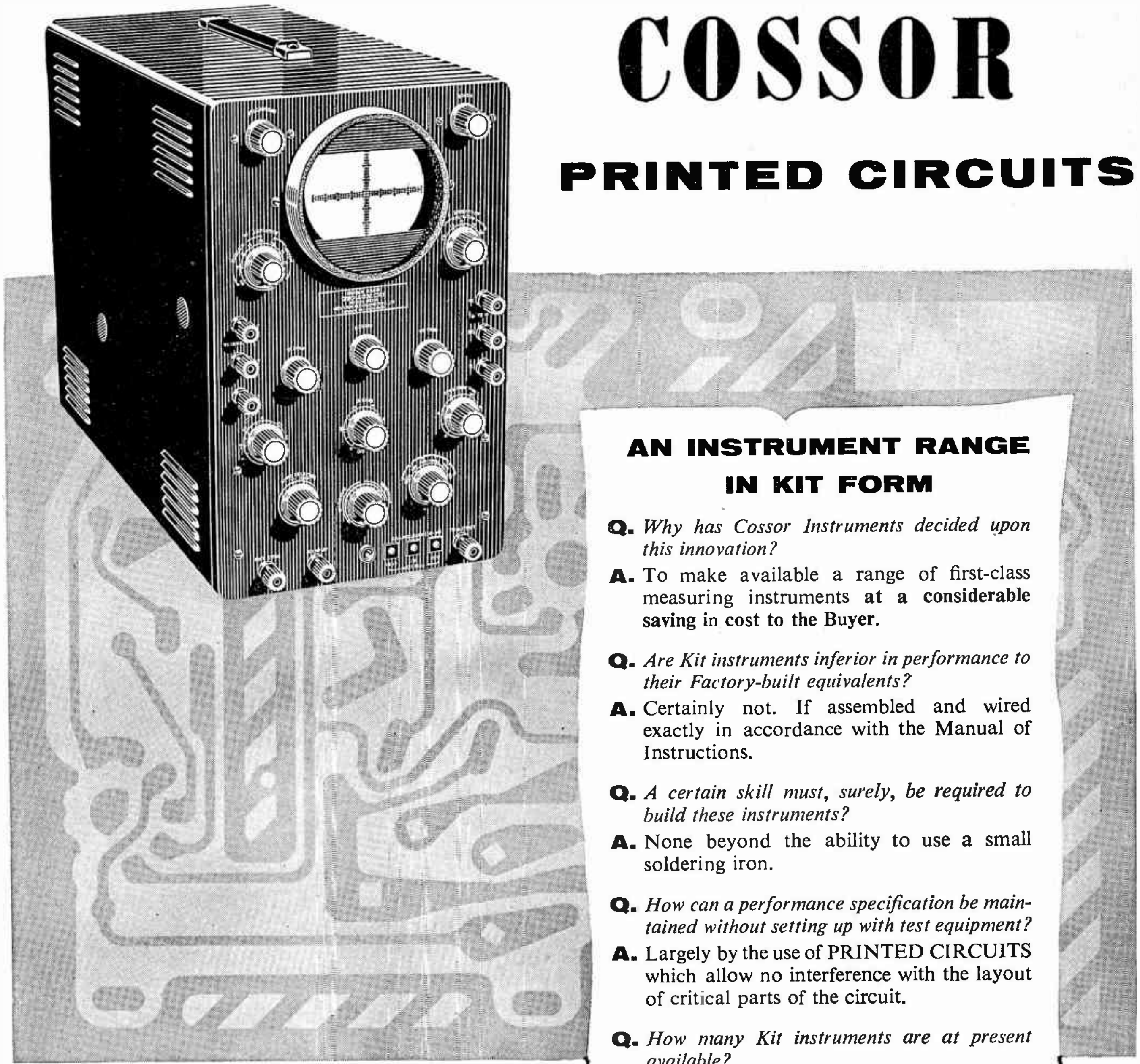

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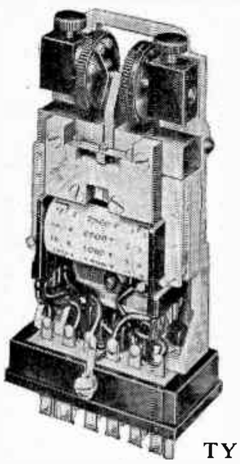
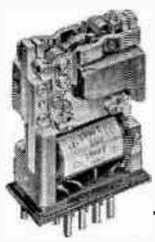


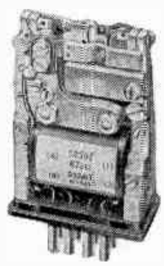
TYPE 3

TYPE 4

TYPE 5



TYPE 51



TYPE 6

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the **CARPENTER** Polarized Relay

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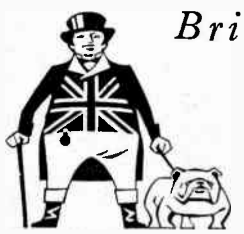

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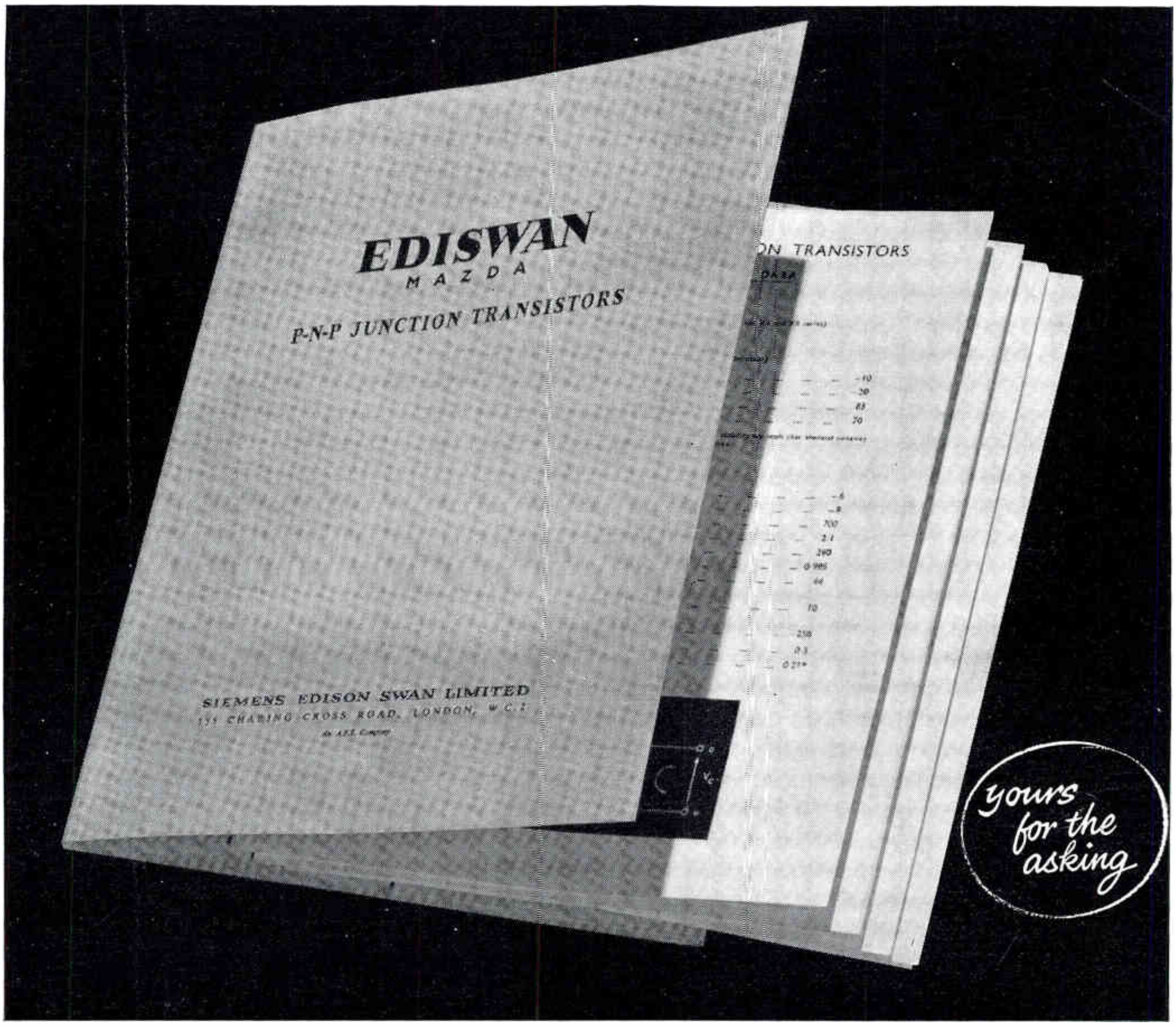

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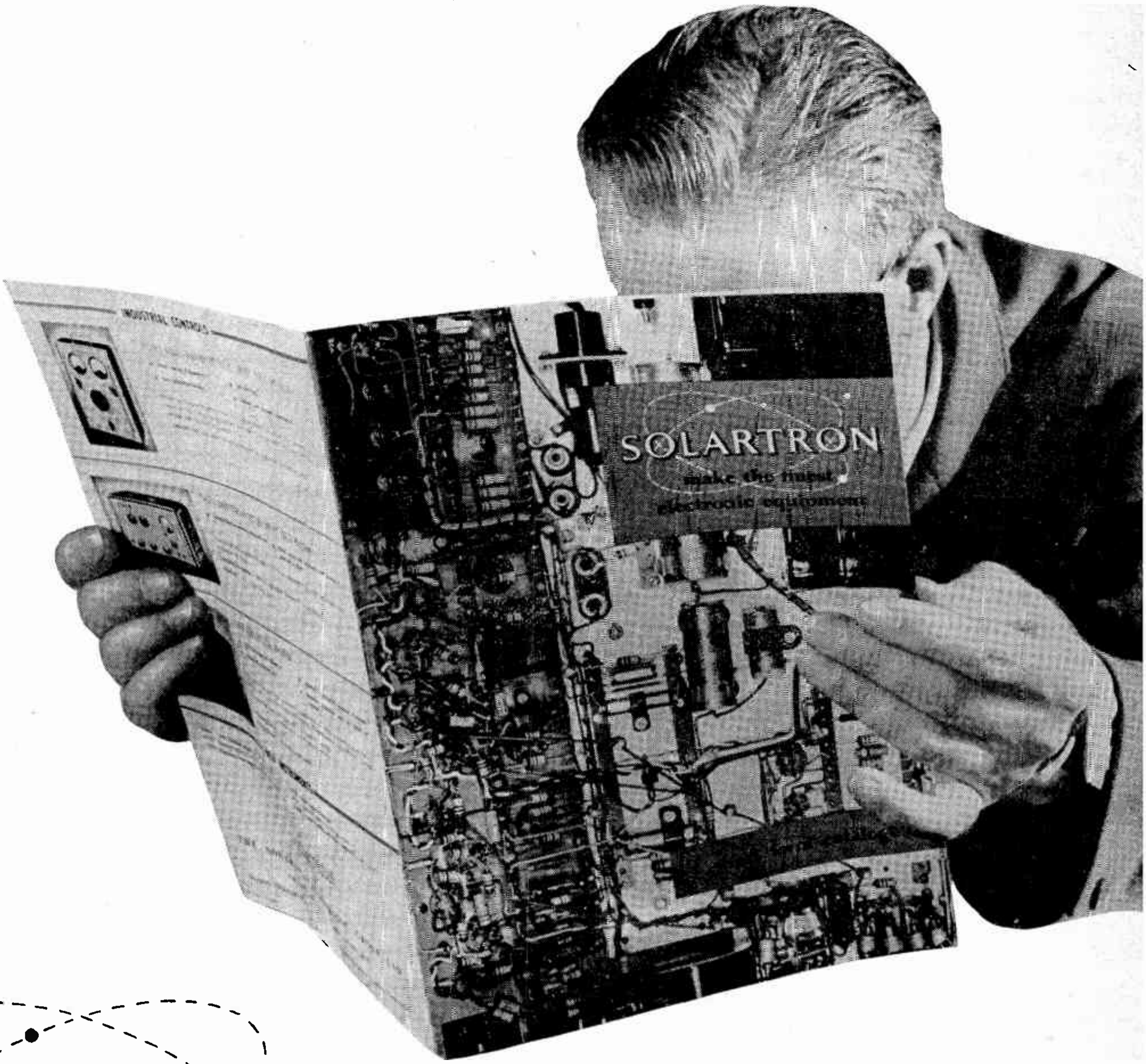

SOLARTRON's short form catalogue

ought to be in *your* desk drawer for easy reference. It is purposely brief and compact. More detailed literature is available on specific subjects . . . and don't forget that Solartron's specialist instrument engineers are available to visit and offer skilled advice. Your own time is now more than ever precious — why spend it on problems we have already thoroughly covered for you.

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TYPE 761

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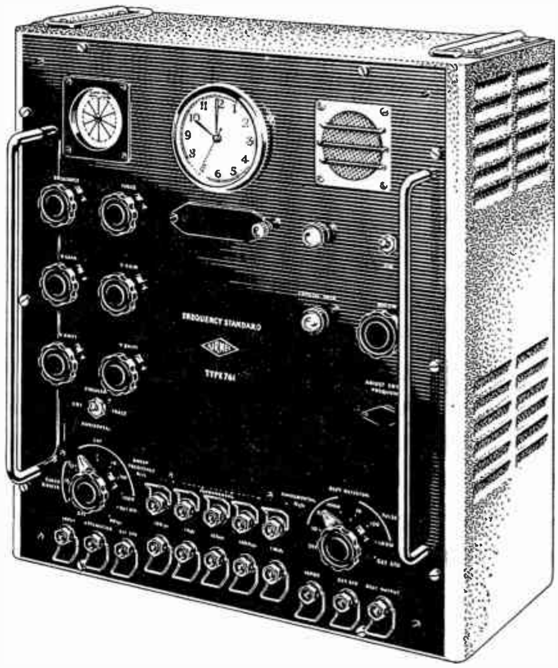

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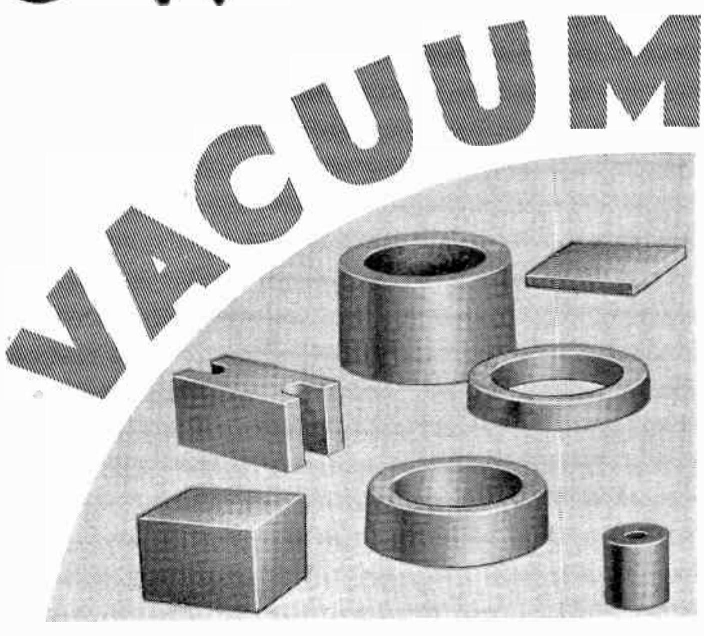

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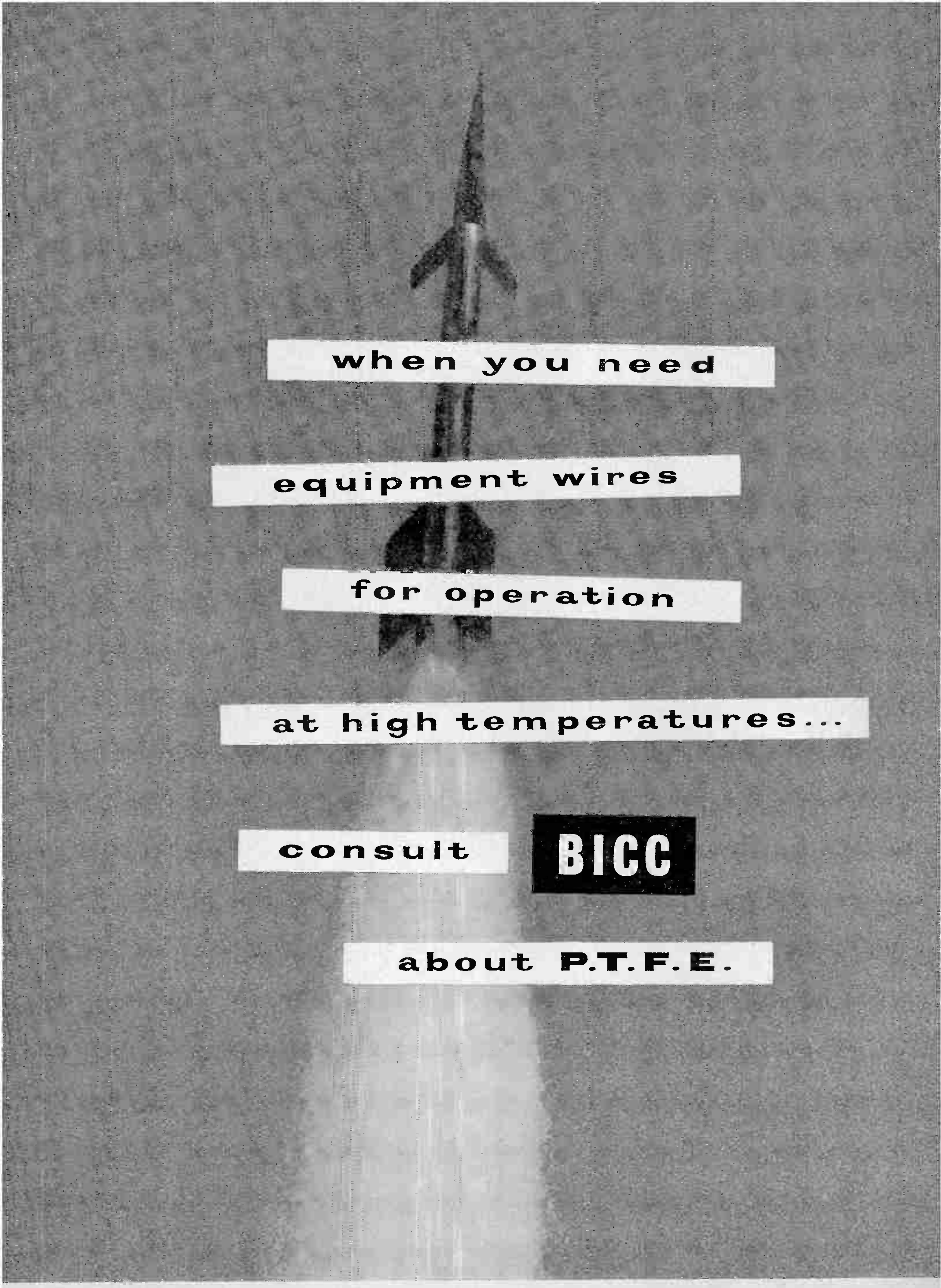

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E·M·I

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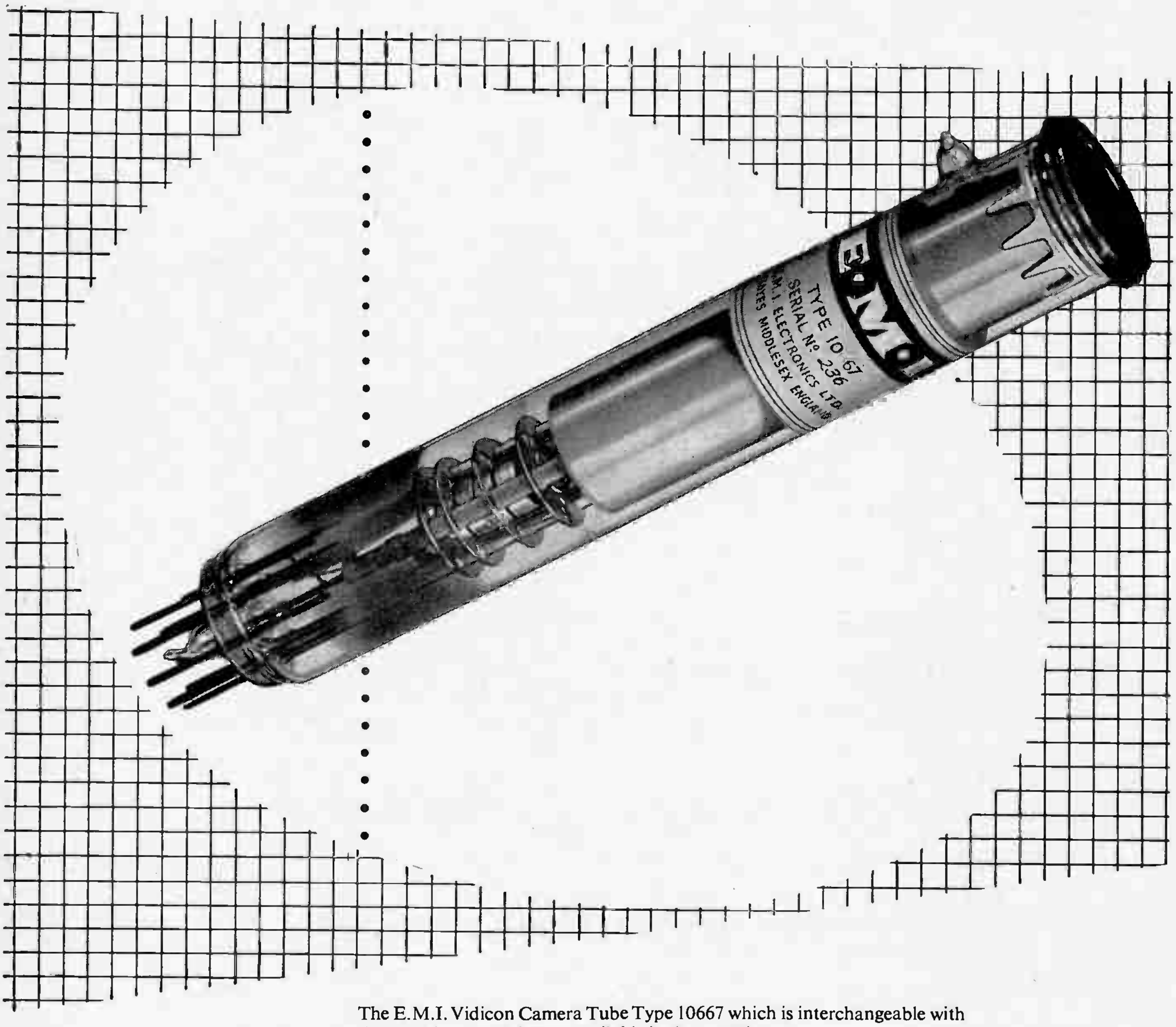

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ELECTRONIC &
RADIO ENGINEER

incorporating WIRELESS ENGINEER

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VOLUME 34 NUMBER 10

OCTOBER 1957

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

References

IT is customary in technical writing to refer to the literature of the subject. This is done partly in order that due credit may be given to earlier workers but mainly to help the reader by indicating where he can find a fuller treatment of certain aspects of the subject.

We have noticed lately, however, that some authors are starting to use references in a way which we consider improper. They use them so that the reader is forced to read earlier papers before he can understand what they write. Sometimes the whole paper rests upon the assumption that the reader is familiar with a previous one; sometimes an essential step in the argument is passed over by a reference.

We regard this as laziness on the part of the writer. It is his job to write so that he can be understood without outside reference by the kind of people for whom he is writing. It is unsafe to assume that they are all familiar with everything that has previously been published on the subject. Of course, it is unnecessary to repeat all previous arguments in full, but he should give a sufficient outline of them to make his paper self-contained and intelligible by itself. His references will then show those who want to go more deeply into the subject where they can find more information.

When a reference is used to avoid including information on a paper which should be there, there are two kinds which are especially useless. One is to the 'unpublished report', the other is to an obscure conference the proceedings of which are exceedingly hard to come by!

We recently met a case of this sort. We found an interesting paper which might be an important one. It contained certain statements and, in support of them, the author quoted a previous paper of his own which he had presented at a symposium. Although he gave the name and date of this there was no indication that his paper had ever been published. The statements that he made, although superficially not unreasonable, were of a kind that a little thought showed to be highly improbable. They shouted aloud for proof or, at the least, a good deal of supporting evidence. Although we have better facilities than many, we have so far failed to obtain a copy of his earlier paper.

Engineers are busy people and cannot afford to waste their time tracking down obscure references. Articles which depend for their proper understanding upon reference to other articles are likely to be left unread.

Drift-Corrected D.C. Amplifier

USE OF STABILIZING AUXILIARY CHANNEL

By M. H. McFadden, B.Sc.*

SUMMARY. *The need is explained for drift correction in d.c. amplifiers, especially those used in real-time analogue computers, and a continuously drift-corrected amplifier is described which is suitable for use in either a repetitive or real-time computer.*

For many applications in low-frequency measuring instruments, servo systems and particularly in 'real-time' analogue computers the use of directly-coupled amplifiers is essential. However, such amplifiers are liable to produce random voltages at their outputs when the input voltage is zero, owing to changes occurring within the amplifier. Change of cathode temperature in the valves contributes most of this zero error but changes in component values and h.t. voltages also add to the trouble.

The zero-error voltage is usually referred to the input grid of the amplifier so that various amplifiers can be fairly compared regardless of their gains. Also, as will be shown later, the inaccuracy caused by the zero-error voltage in a computer feedback amplifier is determined almost entirely by the feedback component values and the grid-referred zero-error. The gain has little effect provided it is high. Grid-referred zero-error is, of course, the output zero-error divided by the amplifier gain and can be represented by a generator of this e.m.f. connected in series with the input grid. The amplifier itself is then considered to be free from drift.

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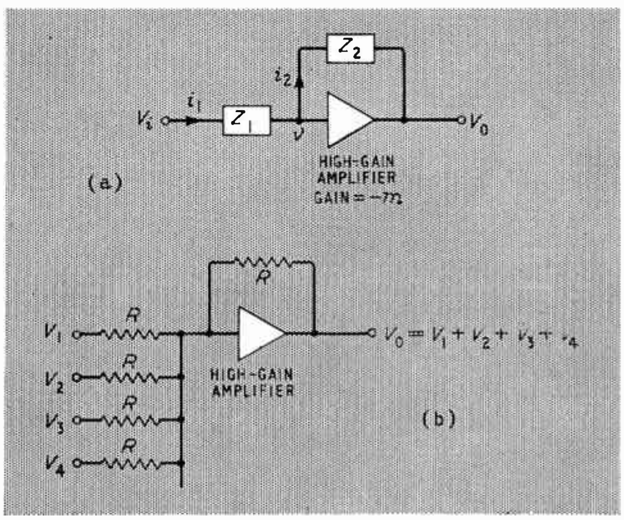
Fig. 1. *Feedback circuits used in analogue computers*



The computing units used in an analogue computer are of the type shown in Fig. 1 (a), and the nature of the impedance elements Z_1 and Z_2 determines the mathematical operation performed by the unit. Since the two most common units are the adder and integrator the effects of zero-error will be discussed in these two cases only. In general, if the amplifier gain is $-m$ and if no current flows into the input grid, the currents i_1 and i_2 are equal. The amplifier is also assumed to be drift-free and the prime after the symbols denotes that Laplace transforms are used. Under these conditions:

$$\frac{V'_i - v'}{Z'_1} = \frac{v' - V'_0}{Z'_2}$$

But $v' = \dfrac{-V'_0}{m}$

$$\therefore \frac{V'_i + (V'_0/m)}{Z'_1} = -\left[\frac{(V'_0/m) + V'_0}{Z'_2}\right]$$

$$V'_0\left[\frac{1}{mZ'_1} + \frac{1}{mZ'_2} + \frac{1}{Z'_2}\right] = -\frac{V'_i}{Z'_1}$$

or
$$\frac{V'_0}{V'_i} = \frac{mZ'_2}{Z'_1(1+m) + Z'_2} \quad \ldots \quad \ldots \quad (1)$$

Thus, if m is very large

$$\frac{V'_0}{V'} \approx \frac{Z'_2}{Z'_1} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

and independent of amplifier gain. If Z_2 and Z_1 are pure resistances R_2 and R_1, the unit becomes an adder with a gain of $- R_2/R_1$. Fig. 1 (b) shows an adder with several input voltages and, if all resistors including the feedback resistor are equal, the output voltage is approximately the sum of all the input voltages.

When used for integration Z_1 is a pure resistance R_1, and Z_2 a pure capacitance C. The ratio V'_0/V'_i (or transfer function) is then $1/pCR_1$ so that the output voltage is $1/CR_1$ times the integral of the input voltage.

The effects of zero-error in an adder may be followed by reference to Fig. 2 (a), which represents the zero-error as a generator in series with the input of a drift-free amplifier. In this case

$$v = V_0 \cdot \frac{R_1}{R_1 + R_2} \text{ and } V_0 = -m(v + V_d)$$

substituting for V_1

$$V_0 = -m\left(V_0 \cdot \frac{R_1}{R_1 + R_2} + V_d\right)$$

$$V_0\left[1 + \frac{mR_1}{R_1 + R_2}\right] = -mV_d$$

$$\therefore V_0 = -mV_d \bigg/ \left[1 + \frac{mR_1}{R_1 + R_2}\right] \quad \ldots \quad (3)$$

If $mR_1/(R_2 + R_1)$ is much larger than unity,

$$V_0 \approx \frac{V_d(R_1 + R_2)}{R_1} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

and so, as stated previously, is dependent only on the grid-referred zero-error voltage and the values of the feedback components.

In the case of the integrator shown in Fig. 2 (b),

$$v' = \frac{V'_0 pCR_1}{1 + pCR_1}$$

But since $m(v' - V'_d) = V'_0$ then $v' = \dfrac{V'_0 + mV'_d}{m}$

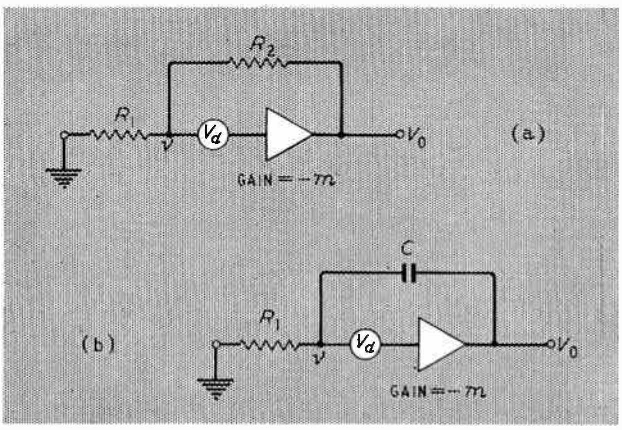
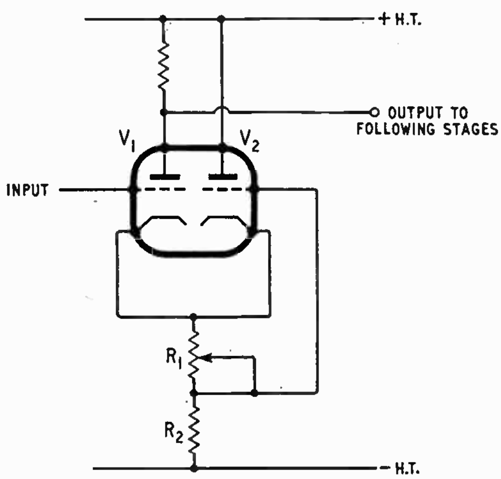


Fig. 2. *Effect of drift in (a) an adder and (b) an integrator*

Fig. 3. *The Miller heater-voltage compensation circuit*



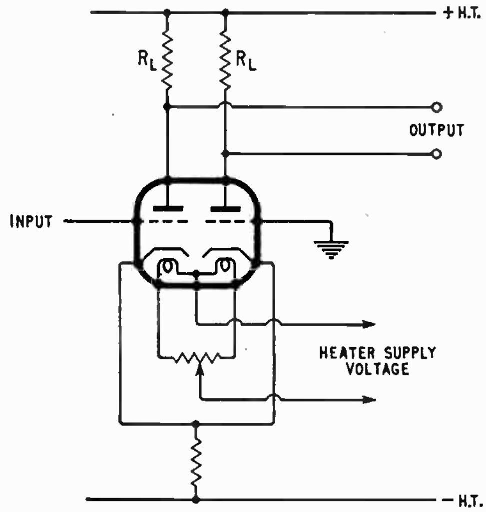


Fig. 4. *Differential balancing of cathode temperatures*

substituting this expression for v'

$$\frac{V'_0 + mV'_d}{m} = \frac{V'_0 pCR_1}{1 + pCR_1}$$

$$\therefore V'_0\left[\frac{pCR_1}{1 + pCR_1} - \frac{1}{m}\right] = V'_d$$

If m is large $V'_0 = V'_d\ (1 + 1/pCR_1)$ and if V_d is assumed to be constant in value (with a Laplace transform V_d/p)

$$V'_0 = \frac{V_d}{p}\left[1 + \frac{1}{pCR_1}\right]$$

or $$V_0 = V_d + \frac{V_d t}{CR_1} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

The output voltage therefore increases linearly with time at a rate V_d/CR_1 volts per second, starting from an initial value V_d when the voltage across the integrating capacitor is zero.

The Need for Drift Correction

In computers used in real-time problems the lowest integrator time-constant likely to be encountered is $CR_1 = 0{\cdot}1$ second and the highest adder gain (R_2/R_1) is normally less than 100. It is obvious from equation (5) that the time-dependent drift voltage at the output of an integrator places a limit on the maximum computing time if a reasonable accuracy is to be maintained. In real-time problems computing periods of hundreds of seconds are common and in aircraft simulation, where it is desirable to observe the behaviour of the aircraft in simple manœuvres, 1,000 seconds may be required. Using an amplifier whose peak linear output is 50 volts, the grid-referred error voltage requires to be less than 50 microvolts if the error in an integrator with time constant 0·1 second is not to exceed 1% of full scale (i.e., 0·5 volt) in 1,000 seconds. [See equation (5).] It is necessary therefore to aim at a grid-referred error voltage of less than this value in the design of amplifiers suitable for applications of this kind.

Many circuit arrangements have been employed to reduce zero-error in d.c. amplifiers. Since heater-

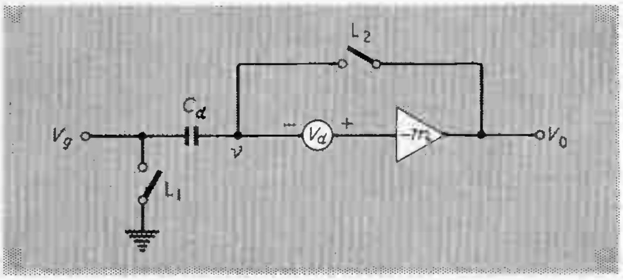


Fig. 5. *A drift-correction circuit for repetitive operation*

voltage changes in the input stages contribute a large proportion of the drift, a number of heater-voltage compensation circuits have been designed. The popular Miller compensation circuit is shown in Fig. 3 and it can be shown[1] that if R_1 is adjusted so that V_2 operates under conditions such that its mutual conductance is $1/R_2$ compensation will be optimum. Another circuit is shown in Fig. 4 where the anode currents in the two valves are equalized by a differential adjustment of their heater voltages. It is claimed[2] that this equalizes the thermal characteristics of the cathodes of the two valves, resulting in a reduction of drift when the heater voltage changes.

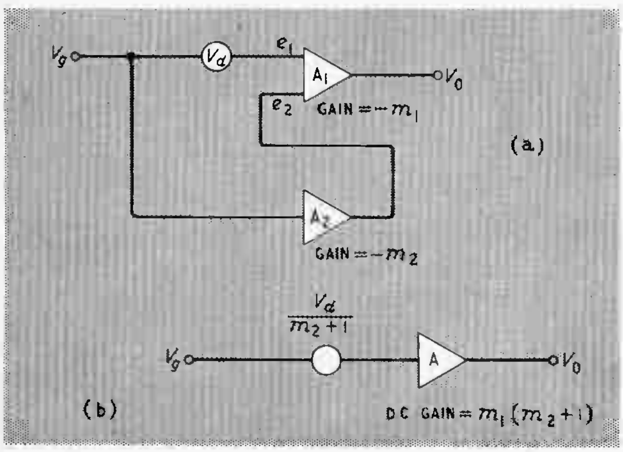
It is difficult to obtain grid-referred error voltages of less than 0·5 mV over long periods with circuits of this sort and some form of external drift correction is required. One system which is worthy of mention, although it is not suitable for long computing periods, is shown in Fig. 5. The circuit is extremely good in repetitive computers where the solution is repeated continuously and displayed on a cathode-ray tube. The relays L_1 and L_2 are open during the computing period and are closed for a time between each computing period. When the relays are closed $v = V_0$ and, if they remain closed sufficiently long, the capacitor C_d will charge to the voltage v. If the amplifier gain is $-m$, and if grid current is ignored:

$$V_0 = -m(v - V_d)$$
$$= -m(V_0 - V_d)$$
$$\therefore V_0 = \frac{m}{m+1}(V_d)$$

and if m is large $V_0 = v \approx V_d$.

At the start of the next computing period when L_1 and L_2 reopen, the capacitor is connected in series with the input of the amplifier and the p.d. across it cancels the zero-error voltage. Each computing period therefore starts free from zero-error and this condition will remain as long as V_d remains constant and C_d retains its charge. Because of grid current, capacitor leakage and the fact that V_d does not remain constant, the drift correction is effective only for periods of about 60 seconds. For longer periods the error voltage requires to be continuously corrected and a circuit which meets the requirements of a real-time computer is shown in Fig. 6 (a) and will be described in detail.

Fig. 6. *Basic circuit of continuous drift-correction system; the equivalent circuit of a continuously drift-corrected amplifier at d.c. is shown in (b)*



Basic Cascade Circuit

The main amplifier is designed for fast repetitive computation and therefore has a comparatively wide bandwidth to minimize errors resulting from phase shift.[3] The auxiliary amplifier A_2 is a modulated carrier type and, to describe the operation of the system, is assumed to be free from zero-error. This auxiliary amplifier operates over a very restricted frequency range.

The main amplifier has two differential inputs, hence the output voltage caused by a voltage e_1 at one input and a voltage e_2 at the other is $-m(e_1 - e_2)$.

Since $e_1 = V_g - V_d$ and $e_2 = -m_2 V_g$

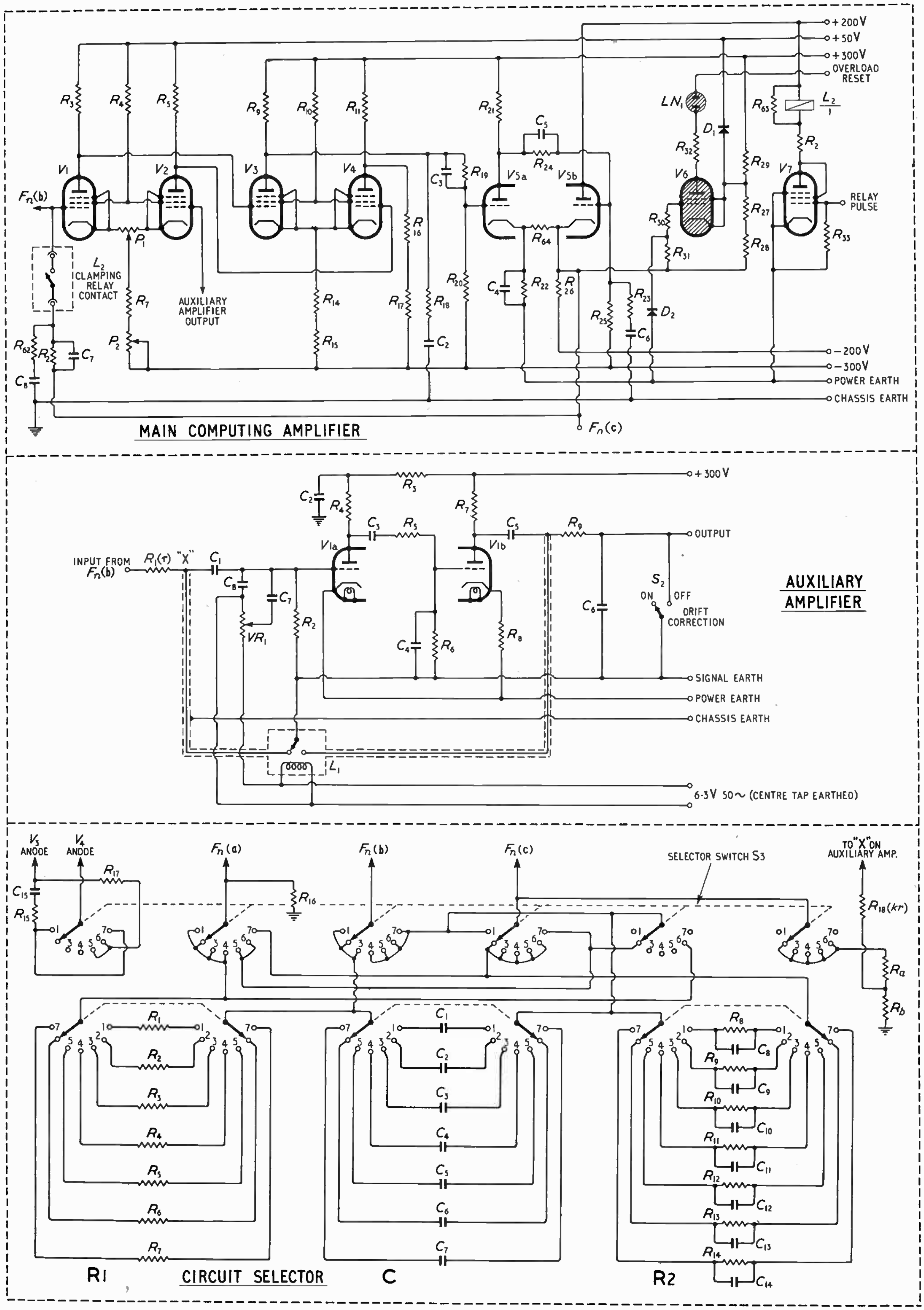
$$V_0 = -m_1(V_g - V_d) + m_2 V_g$$
$$= -V_g(m_1 + m_1 m_2) + m_1 V_d$$

The complete system has therefore a gain of $-m_1(1 + m_2)$ at zero frequency and an output zero error of $m_1 V_d$ which, when referred to the input of the system, is $V_d/(m_2 + 1)$. An equivalent circuit at zero frequency is shown in Fig. 6 (b).

In designing the circuit, one main point to note is that variations in zero-error voltage occur slowly because, as already mentioned, they are due mainly to temperature variations of the valve cathodes and components in the amplifier where the thermal time-constants are quite large. This means that the auxiliary amplifier A_2 can operate over a relatively narrow frequency range and this is a significant factor in its design. The method adopted is to chop the input signal V_g into a series of amplitude-modulated square waves, and a.c. coupling is then used in the auxiliary amplifier to eliminate drift. The output of the amplifier is demodulated to reproduce the amplified signal voltage. An electromagnetic relay is used for modulation and demodulation and is driven from the 50-c/s mains supply.

A detailed circuit embodying this system is shown in Fig. 7. The main amplifier which uses direct coupling

Fig. 7. *Right. Circuit diagram of continuously drift-corrected amplifier*



+200V
+50V
+300V
OVERLOAD RESET
RELAY PULSE
L2 CLAMPING RELAY CONTACT
AUXILIARY AMPLIFIER OUTPUT
−200V
−300V
POWER EARTH
CHASSIS EARTH
MAIN COMPUTING AMPLIFIER
+300V
OUTPUT
INPUT FROM Fn(b)
ON
OFF
DRIFT CORRECTION
AUXILIARY AMPLIFIER
SIGNAL EARTH
POWER EARTH
CHASSIS EARTH
6·3V 50~ (CENTRE TAP EARTHED)
V3 ANODE
V4 ANODE
SELECTOR SWITCH S3
TO "X" ON AUXILIARY AMP.
R1
CIRCUIT SELECTOR
C
R2

throughout has two cathode-coupled electrometer valves V_1 and V_2 in the first stage. The second stage is a cathode-coupled pair of high-gain pentodes, the third stage is a single triode and the output stage a cathode-follower. An overload circuit which indicates when the output voltage exceeds $\pm$ 50 volts consists of a thyratron V_6, a neon lamp LN_1 and clamping diodes D_1 and D_2. A clamping relay, whose contacts are connected to the output and input of the amplifier and used for discharging integration capacitors at the end of a computing period, is energized by V_7. The d.c.-voltage amplification of the main amplifier is approximately 10,000, dropping by 3 dB at a frequency of 2 kc/s. The input grid current is less than 2×10^{-11} ampere and the grid-referred zero-error does not exceed 0·8 mV over a period of one hour.

The auxiliary amplifier consists of two stages using the two sections of a double triode in cascade. The components R_5 and C_4 are incorporated to prevent high-frequency oscillation which may occur during the period when the armature of the chopper relay is in transit between the two contacts. The inter-contact capacitance couples the output to input during this period. Grid-current biasing is employed in the first stage of the amplifier, using a high-value resistor for R_2. This method of biasing permits direct earthing of the cathode of V_{1a}, which is desirable to minimize a.c. hum at the input due to heater-cathode capacitance.

The h.t. supply to V_{1a} is smoothed by R_3 and C_2 as a further aid to noise reduction. The capacitance of C_5 requires to be carefully chosen in relation to the output resistance of V_{1b}. Theoretical considerations and experimental results indicate that the time-constant R_0C_3 should be smaller than $1/2\pi f_c$ where f_c is the carrier frequency (50 c/s) and R_0 is the output resistance. Higher values result in C_5 being insufficiently charged during the charging period (0·01 sec) when one side of C_5 is earthed via the relay contact. This reduces the effective gain of the amplifier. When C_5 is too low in value the charge on C_5 available each cycle for charging C_6 is unduly low and again the effective gain is reduced.

In practice, the auxiliary amplifier inevitably has a small zero-error caused by noise picked up at the input or generated within itself. Most of the noise is at 50 c/s and is therefore in synchronism with the relay operation so that the same instantaneous value is sampled by the chopper relay on each cycle and gives rise to a d.c. zero-error at the output of the auxiliary amplifier. The use of a centre-tapped heater and chopper relay supply, in addition to reducing a.c. hum, also permits the use of the hum-balancing circuit C_7, C_8 and VR_1.

Tests on the amplifier system show that, if the output voltage is adjusted in this way, over a period of an hour the grid-referred zero-error is never more than 25 microvolts. (See Fig. 9.) The grid-referred zero-error of the auxiliary amplifier alone does not exceed 5 microvolts over the same period.

The switching rate of the relay armature in the auxiliary amplifier sets a limit on the maximum signal frequency which can be handled. At input frequencies approaching the switching frequency, complex phase errors occur which could cause instability when the feedback network is connected. To prevent this, the

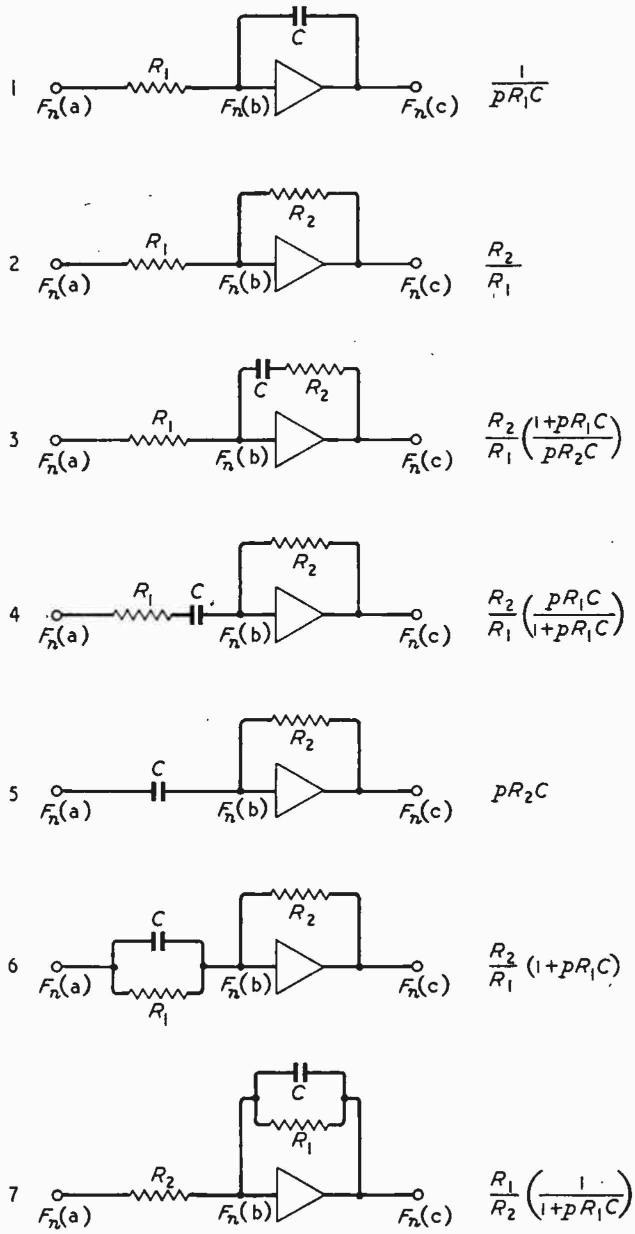
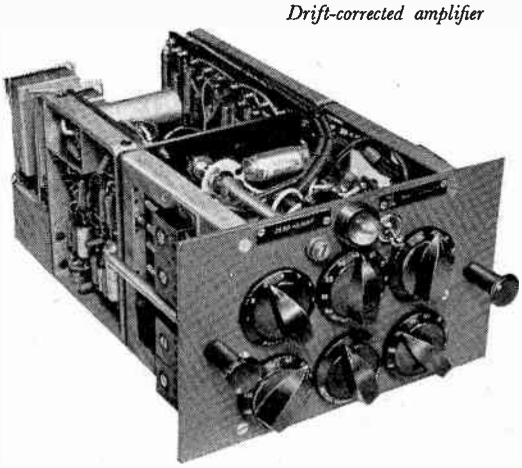


Fig. 8. *Feedback arrangements selected by the circuit selector shown in Fig. 7*

Drift-corrected amplifier



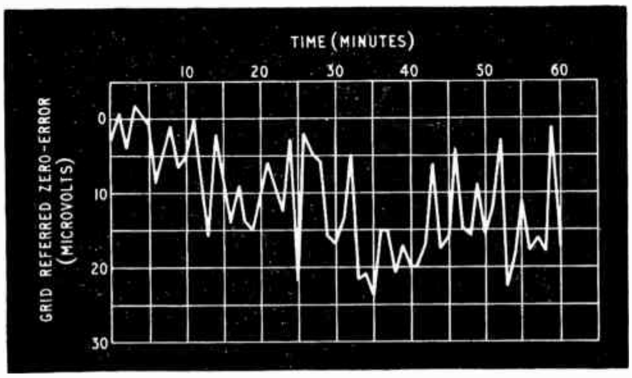


Fig. 9. *Zero-error of the drift-corrected amplifier over a period of one hour*

time constant R_9C_6 of the output filter circuit is made so large that it almost completely determines the frequency response of the amplifier.

The gain of the auxiliary amplifier at an angular frequency ω is therefore very closely $-m_2/(1+j\omega T)$ where m_2 is the zero-frequency gain and T the time-constant of the filter circuit and the gain of the overall system is $-m_1\,[1+m_2/(1+j\omega T)]$ where m_1 is the gain of the main amplifier.

Phase and Gain Characteristics

The zero-frequency gain of the auxiliary amplifier shown in Fig. 7 is 1,000 and the time-constant of the output filter is 20 sec. The gain and phase response of the complete drift-corrected system are as shown in Fig. 10. At zero frequency the overall gain is 10^7, falling as frequency increases to 10^4, the gain of the main amplifier. During the change-over the phase change of the system reaches a peak, falling off to near zero again as the change-over reaches completion. When feedback networks are connected across the amplifier the phase error is, of course, very much reduced but is still of some significance. Consider first an adder with feedback components R_1 and R_2 [as shown in Fig. 2 (a)] and amplifier gain modulus m and phase angle θ (i.e., $me^{-j\theta}$ assuming that θ is a phase lag).

$$\frac{V_0}{V_i} = \frac{me^{-j\theta}R_2}{R_2+(me^{-j\theta}+1)R_1} \quad \text{[From Equation (1)]}$$

$$= \frac{R_2}{\dfrac{R_2e^{j\theta}}{m}+R_1+\dfrac{R_1e^{j\theta}}{m}}$$

$$= \frac{R_2}{\dfrac{R_1+R_2}{m}(\cos\theta+j\sin\theta)+R_1}$$

Phase angle ϕ with feedback

$$= \tan^{-1}\frac{\dfrac{R_1+R_2}{m}\sin\theta}{\dfrac{R_1+R_2}{m}\cos\theta+R_1}$$

In practice $\dfrac{R_1+R_2}{m}\cos\theta$ is much smaller than R_1

$$\therefore \phi = \tan^{-1}\frac{R_1+R_2}{mR_1}\sin\theta$$

and, since ϕ is a small angle,

$$\phi = \frac{R_1+R_2}{mR_1}\sin\theta.$$

Applying this formula to the curves shown in Fig. 10 the phase response of an adder with $R_2/R_1 = 100$ is plotted in Fig. 11 and it will be noticed that the maximum phase change occurring within the range of solution frequencies (i.e., approximately 0–30 c/s) is 0·55° occurring at 10 c/s. Now it can be shown[3] that, in the solution of a simple second-order differential equation with no damping, phase error in the computing loop will cause 1% spurious damping after N cycles where $N = 0{\cdot}181/\phi$ and ϕ is the phase error in degrees. If the adder under discussion is used in such a computing loop a 1% amplitude error will occur after only

$$N = \frac{0{\cdot}181}{0{\cdot}55} = 0{\cdot}33 \text{ cycles.}$$

The cascade coupling of the main and auxiliary amplifiers is therefore not suitable for use in an adder, particularly in one intended for repetitive computation where solution frequencies of 10 c/s are common.

In the case of an integrator with feedback components R_1 and C [as shown in Fig. 2 (b)]:

$$\frac{V_0}{V_i} = \frac{me^{-j\theta}\dfrac{1}{j\omega C}}{\dfrac{1}{j\omega C}+(me^{-j\theta}+1)R_1}$$

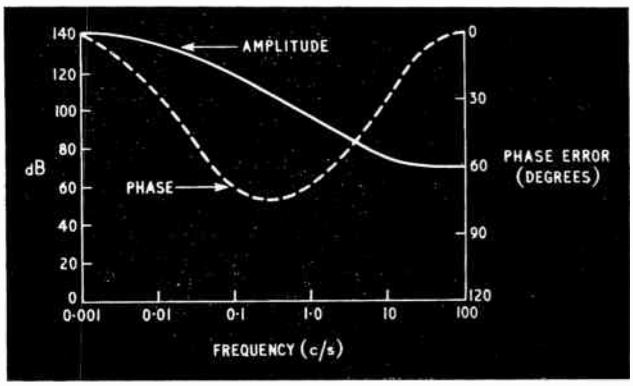
$$= \frac{\dfrac{1}{j\omega C}}{\dfrac{1}{j\omega Cm}(\cos\theta+j\sin\theta)+R_1+\dfrac{R_1}{m}(\cos\theta+j\sin\theta).}$$

The 90° phase change indicated by the numerator is of course inherent in the process of integration. Any departure from this (shown in the denominator) is a phase error.

The integrator phase error ϕ is therefore

$$= \tan^{-1}\frac{\dfrac{-\cos\theta}{\omega Cm}+\dfrac{R_1\sin\theta}{m}}{\dfrac{\sin\theta}{\omega Cm}+R_1+\dfrac{R_1\cos\theta}{m}}$$

Fig. 10. *Gain and phase characteristics of the drift-corrected amplifier (cascade connection)*



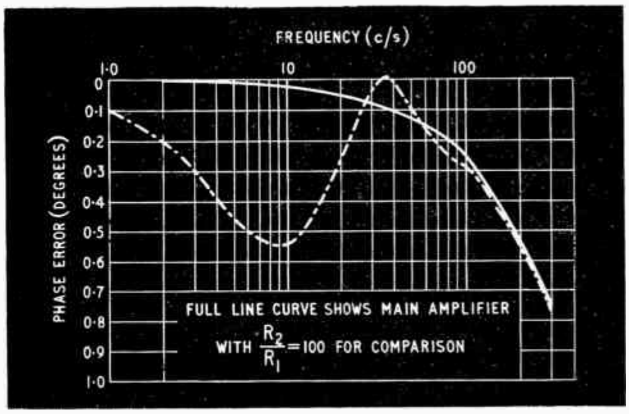


Fig. 11. *Phase response of a continuous drift-correction system connected as an adder with $R_2/R_1 = 100$*

$\frac{\sin\theta}{\omega Cm}$ and $\frac{R_1 \cos\theta}{m}$ are negligibly small since m is in excess of 10^4 and, as the smallest value of C is 0·1 μF, the maximum phase error occurs at $\omega \approx 50$.

$$\therefore \phi = \tan^{-1} - \frac{1}{m}\left(\frac{\cos\theta}{\omega CR_1} + \sin\theta\right)$$

and again, because m is larger than 10^4, and the maximum value of θ less than 90° the integrator phase error is less than 10^{-4} radians and quite insignificant.

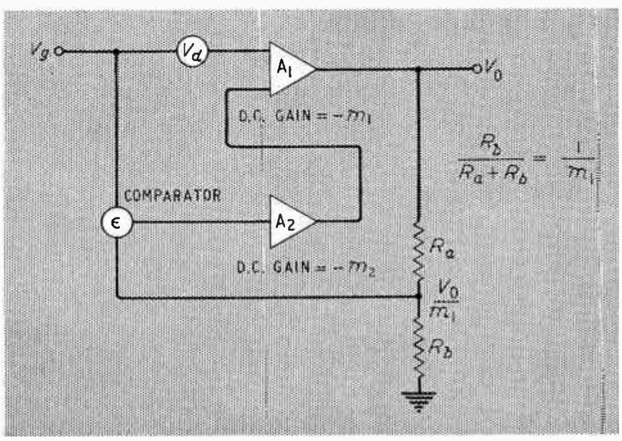
Alternative Circuit

To avoid an excessively large phase error in the adder the alternative connection of the auxiliary amplifier shown in Fig. 12 may be used. The d.c. gain of the main amplifier is m_1 and the components R_a and R_b are chosen so that the output from the potential divider is V_0/m_1. This voltage is compared with the input voltage V_g and the difference is amplified in the auxiliary amplifier and fed back into the main amplifier to reduce the difference.

If V_d is the grid-referred error voltage of the main amplifier and the auxiliary amplifier is assumed to be drift-free:

$$V_0 = -m_1(V_g - V_d) + m_2(V_g + V_0/m_1)$$

Fig. 12. *Basic circuit showing the alternative connection of the auxiliary amplifier*



$$V_0(1 + m_2) = V_g(-m_1 - m_1 m_2) + V_d m_1$$
$$V_0 = -m_1 V_g + V_d m_1/(1 + m_2)$$

This connection of the auxiliary amplifier is therefore as effective as the cascade connection in reducing zero-error but the gain at zero frequency is that of the main amplifier only. If the potential-divider components at the output of the amplifier are properly chosen the auxiliary amplifier does not handle any part of the input signal, which means that the frequency response of the system is the same as the main amplifier. The system is therefore well suited for use as an adder where the gain of 10^4 will only result in a 1% error when $R_2/R_1 = 100$. The cascade connection is obviously preferable for use in integrators because, as reference to equation (1) will show, the amplifier gain should increase as signal frequency is reduced to achieve accurate integration.

The Circuit Selector shown in Fig. 7, which selects the various feedback arrangements shown in Fig. 8, also changes the auxiliary amplifier connection to cascade when position 1 (integration) or position 7 (phase lag) is chosen.*

In practice the gain m_1 changes, due to ageing of valves, etc., and, because the potential divider is then no longer correct, a part of the input signal will pass through the auxiliary amplifier giving rise to a slight phase error. Measurements show, however, that when m_1 changes by 10% the phase error does not exceed 0·05°.

In the cascade arrangement it is important that the insulation resistance of the integration capacitors should be very high, otherwise it is not possible to achieve the low-frequency performance which should result from the high amplifier gain. If the leakage resistance across the capacitor is R_L the gain of the amplifier is, in effect, reduced to R_L/R_1 if R_L/R_1 is much less than 10^7.

It should also be noted that the disagreement between the drift-correction factor of 32 (i.e., main amplifier zero-error divided by the zero-error of the complete system) obtained in practice and the theoretical figure of $m_2 + 1 = 1000$ is due to the zero-error introduced by the auxiliary amplifier and also to the input grid current of the main amplifier. It is possible to reduce the error caused by grid current by using a.c. coupling at the input of the main amplifier. This is quite permissible since the direct input to the main amplifier contributes very little to the d.c. gain.

Acknowledgments

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* *These and other features of the amplifier are covered by British and foreign patents.*

Uni-Control Wide-Range Oscillator

By S. N. Das, M.Sc.*

SUMMARY. *The article describes a wide-range oscillator using a single tank circuit between the grid and anode of a triode. A variable* LC *circuit, containing a variable inductor formed by two parallel rectangular strips bent in the form of a semicircle and two variable capacitors having semicircular rotors and stators, is connected to the grid and anode of the triode through a pair of transmission lines. At the high-frequency end the lines connecting the grid and anode of the valve are used as a transmission line, the* LC *circuit being automatically isolated. The transition from the* LC *mode of operation to the transmission-line mode of operation is smooth, because the strips forming the transmission line are continued on the same circle as the strips forming the tuning inductance. With the tank circuit described here, continuous oscillation has been obtained from* 25 *to* 580 *Mc/s. The frequency range of oscillation can be easily extended at both the high- and low-frequency ends by suitably modifying the circuit parameters.*

A radio-frequency oscillator with relatively constant output, operating over a wide frequency range, has many applications. An ordinary *LC* circuit (with capacitance variation) gives a tuning range of about 3 : 1 and, for increasing the range, a second control is required.

The range of operation is considerably increased by using butterfly circuits[1] in which both *L* and *C* are variable. The low-frequency range of the butterfly circuit is limited by the size of the circuit elements and the upper frequency is limited by the presence of residual capacitance and inductance. In another type of wide-band oscillator the frequency range of operation has been increased by using a variable *LC* circuit with a conical helix[2] as tuning inductor.

The paper describes a resonator working over a frequency range of about 23 : 1. The tank circuit is a combination of a variable *LC* circuit and a variable short-circuited transmission line.

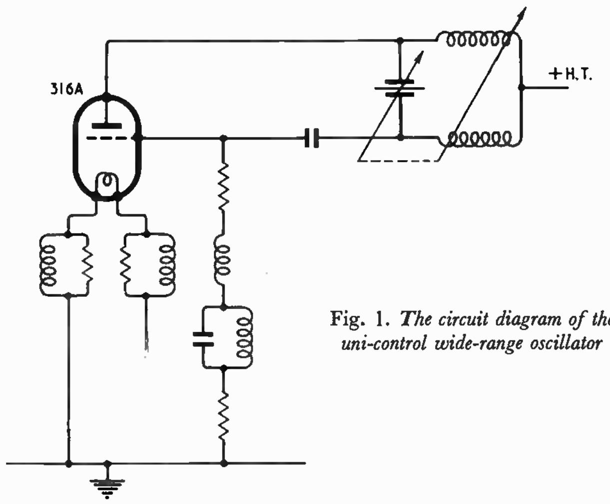

Fig. 1. *The circuit diagram of the uni-control wide-range oscillator*

Circuit Design

The circuit diagram of the oscillator is shown in Fig. 1. The oscillator valve is a 316A triode. The filament voltage is applied through chokes shunted by resistances so as to provide a high impedance between the filament and ground over the entire range of operating frequencies. The h.t. is connected to one end of the tank circuit and the grid resistor is connected directly to the grid point of the valve. A choke and a tuned circuit are connected in series with the grid resistor to obtain continuous oscillation throughout the range.

Tank Circuit

The tank circuit is a combination of variable *LC* (both variable) and short-circuited transmission line. At the low-frequency end the variable *LC* circuit is operative and the transmission line merely acts as the connecting leads to the grid and anode. At the high-frequency end the short-circuited transmission line is operative. The disposition of the transmission line is such that the transition from the *LC* mode of operation to transmission-line mode of operation is smooth. The *LC* circuit is automatically isolated electrically when the circuit operates in the transmission-line mode.

The variable capacitor is of the split-stator type, the two stator sections being placed one above the other and electrically insulated. Each stator consists of a number of plates of which one plate (Fig. 2) has been specially designed to function as the inductance arm and also as one line of the parallel transmission line. The other stator and rotor plates are shown in Fig. 3. When the stators are mounted one above the other the two semicircular inductor arms (Fig. 2) are parallel and one above the other. To obtain variation of inductance an insulator arm is rigidly fixed to the rotor shaft and a metal plate is mounted at the end of the arm so that it lightly presses against the inductance strips. As the metallic strip short-circuits the inductance arms at the point of contact, the tuning inductance is that due to the loop BD formed by the upper strip BA and a similar lower strip connected together at the points D by the short-circuiting strip. As the rotor shaft is rotated the

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short-circuiting metal plate moves along the inductance strips and thus a smooth variation of inductance is obtained. The tank circuit is shown in Fig. 4.

When the rotors are fully inside the stators the short-circuiting strip is at A and thus the maximum values of inductance and capacitance are obtained. As the shaft is rotated in a clockwise direction the rotors move out of the stators and the inductance is reduced due to the rotation of the short-circuiting strip along the inductor arms. Rotation of the shaft thus reduces both the capacitance and inductance. As the angular width of both the rotors and stators is less than 180° and as the inductance arms are semicircular, minimum capacitance is attained before the inductance is reduced to the minimum value. Further rotation of the shaft reduces the inductance and when the short-circuiting strip is at B (Fig. 2) the *LC* circuit is entirely short-circuited. At this point the grid and anode look into a short-circuited transmission line formed by the strip CB and another such strip below it.

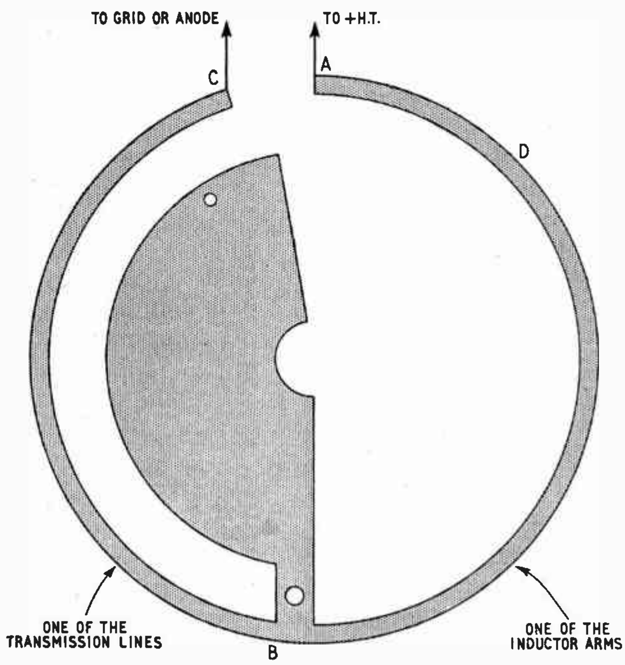
Continued rotation of the shaft in the same direction will decrease the length of the short-circuited transmission line and thus increase the frequency.

The end point A of the inductance arm is connected to the h.t. At the end C of the transmission line the anode of the valve is directly connected to the resonator. The grid is connected to the lower transmission line through a capacitor formed by two small brass plates fixed together with mica dielectric between them. The transmission line is a tapered one, and the distance between the legs is reduced from 1·25 in. at B to 0·39 in. at the valve end C.

Experimental Results

The frequency range of the oscillator depends on the

Fig. 2. *One of the stator plates, containing one of the inductor arms and one of the transmission lines*



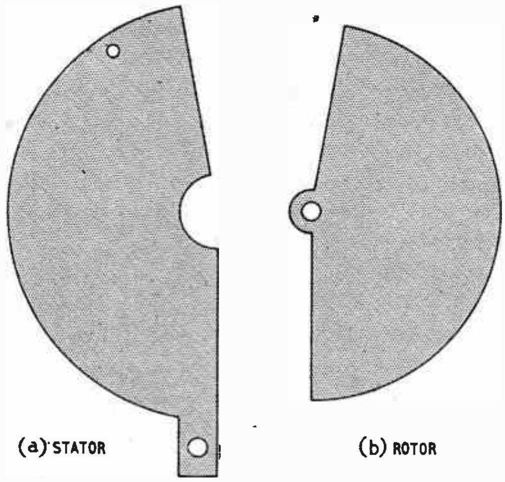


Fig. 3. *Stator and rotor plates*

size of the resonator. The dimensions of the resonator used in this investigation were as follows:

Outer radius of the inductor plate	3·75 in.
Width of the inductor arm	0·25 in.
Radius of the stator plates	2·75 in.
Radius of the rotor plates	2·5 in.
Number of the stator and rotor plates on either side	5

The frequency variation with the rotation of the rotor shaft is shown in Fig. 5. The grid-current variations are shown in Fig. 6.

Theoretical Analysis

In this section an attempt will be made to calculate the frequency range of the oscillator theoretically. The frequency range of such an oscillator can be calculated theoretically from the known dimensions of the tank circuit.

At the high-frequency end the tank circuit consists of a short-circuited transmission line, comprising the anode and grid leads together with the strips connecting the leads to the inductor strips. For lower frequencies the tank circuit consists of a parallel-plate capacitor and an inductance. The length of the transmission line is known for any position of the movable arm and, as such, the calculation of frequency for this mode of operation is simple. For the *LC* mode of operation *L* and *C* are to be calculated separately. The capacitance can be calculated from the geometry of the stator and rotor plates. The inductance is calculated from the formula for the inductance of two parallel rectangular strips.

Transmission-Line Mode of Operation ($\theta = 171°$ *to* $326°$)

The angle of rotation is calculated from the initial position when the rotors are fully inside the stators and the short-circuiting metallic strip is at A (Fig. 2).

For any setting of the rotor (Fig. 7) the length of the transmission line is given by

$$l' = \frac{\pi r}{180}(342° - \theta) + X - Y \qquad \ldots \quad \ldots (1)$$

Where l' = length of transmission line in metres.
r = outer radius of the inductor plate in metres.
X = length of the valve pin and the connecting lead in metres.
Y = width of the short-circuiting metallic strip in metres.
θ = rotor position in degrees.

In the high-frequency range the frequency of oscillation[3] is given by

$$f = \frac{2760}{[Zl'C']^{\frac{1}{2}}} \quad \ldots \quad (2)$$

Where f = frequency, Mc/s.

$$C' \text{ (pF)} = C_{ga} + \frac{C_{gk}\,C_{ak}}{C_{gk} + C_{ak}}$$

C_{ga} = capacitance between grid and anode in pF.
C_{gk} = capacitance between grid and cathode in pF.
C_{ak} = capacitance between anode and cathode in pF.
Z = characteristic impedance of the line in ohms.

LC Mode of Operation ($\theta = 0°$ *to* $171°$)

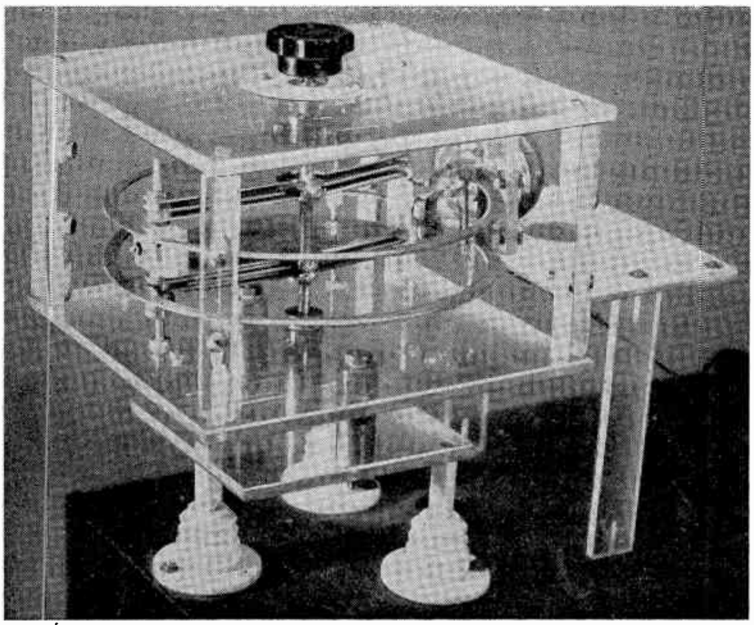
At the lower frequency the dimensions of the inductor strip and the connecting leads are small fractions of the wavelength. As such, the circuit can be treated as consisting of lumped constants.

The capacitance[4] is given by

$$C_0 = 0{\cdot}2244\frac{KA}{2d}\,(n - 1) \quad \ldots \quad (3)$$

Where C_0 = capacitance, pF.
K = dielectric constant.
A = smallest area between the rotor and stator in square inches.

Fig. 4. *The tank circuit used in Fig. 1*



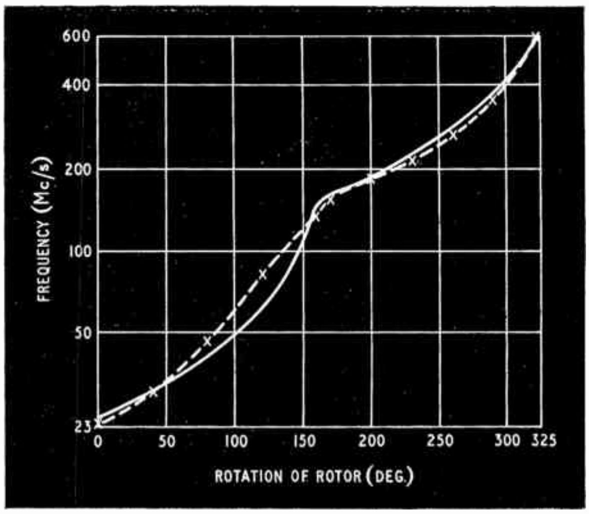


Fig. 5. *Variation of frequency with the position of the rotor. The dotted curve is theoretically obtained and the experimental curve is shown in solid line*

d = separation of plate surfaces in inches.
n = number of plates.

In the above expression the fringing of flux lines at the edge of the plates has been neglected and so the value obtained from the relation is less than the actual capacitance. The factor $\frac{1}{2}$ takes account of the series connection of the two capacitors.

For any position of the rotor (Fig. 7) the length of the inductance strip is given by

$$l = \frac{\pi r}{180}(187° - \theta) - Y \quad \ldots \quad (4)$$

Where l = length of the rectangular strip in metres.
r = outer radius of the inductor strip in metres.
θ = position of rotor in degrees.
Y = width of the short-circuiting strip in metres.

The low-frequency inductance[5] of two parallel rectangular bars is given by

$$L_0 = 0{\cdot}4l\left[2{\cdot}303\log_{10}\frac{D}{b+c} + 1{\cdot}5 - \frac{D}{l} + 0{\cdot}2235\frac{b+c}{l}\right] \quad \ldots \quad (5)$$

Where L_0 = inductance, μH.
b = width of strip in metres.
c = breadth of strip in metres.
D = centre spacing in metres.

When the circuit elements behave as lumped constants the frequency of oscillation is

$$f = \frac{1}{2\pi\sqrt{L_0 C_0}} \quad \ldots \quad (6)$$

where C_0 and L_0 are given by equations (3) and (5) respectively. Two typical calculations are given below for two modes of operation of the experimental oscillator tested.

(a) Transmission-line mode of operation ($\theta = 320°$).

Z (characteristic impedance) $= 312$ ohms.
$l' = 0{\cdot}0366$ m.
$C' = 2{\cdot}0$ pF.
Substituting these values in Equ. (2) $f = 590$ Mc/s.

(b) LC mode of operation ($\theta = 0°$; i.e., rotors are fully inside the stators and the short-circuiting metallic strip is at A).
$L_0 = 0{\cdot}316\ \mu$H.
$C_0 = 128$ pF.
Substituting these values in Equ. (6) $f = 25$ Mc/s.

In Fig. 5 the variation of frequency with rotor position as obtained experimentally is compared with that calculated theoretically. The agreement between the two is quite satisfactory.

It may be mentioned that in the LC mode of operation when both L and C are low the measured frequency of oscillation differs appreciably from the calculated value. Again, the lower the value of L and C the larger the difference. This is due to the following three causes. First, the error in the calculated value of the capacitance at low values of C which arises because the 'fringing effect' is neglected. Second, at higher frequencies the effective capacitance is larger than the calculated value owing to the residual inductance of the capacitor. Third, the effective impedance of the LC circuit between the grid and anode is different from that of the LC circuit at the point B owing to the transmission line, and this is more noticeable at higher frequencies when the length of the transmission line becomes an appreciable fraction of the operating wavelength. The third factor has been taken into account in the calculated frequencies shown in the curve.

Conclusion

The experimental and theoretical investigations described above show that a variable LC circuit can be used in conjunction with a transmission line as the resonator of a wide-range oscillator for efficient working over a large frequency range. The frequency of this type of oscillator varies smoothly with rotor movement both at low and at high frequencies. The hump obtained at the region of transition can be smoothed out by shaping the capacitor plates. The frequency range of such an oscillator can be extended on the high- and low-frequency sides. At the high-frequency end the frequency range of the oscillator can be increased by the following methods: by decreasing the characteristic impedance of the transmission line at the valve end, and by using a valve (e.g., 703A) with lower interelectrode capacitances. The low-frequency end can be extended by increasing both the inductance and capacitance. For the same spacing between the plates the capacitance can be increased by increasing the number and diameter of the capacitor plates. A convenient way of increasing the inductance is to increase the diameter of the inductance strips and the spacing between the two arms.

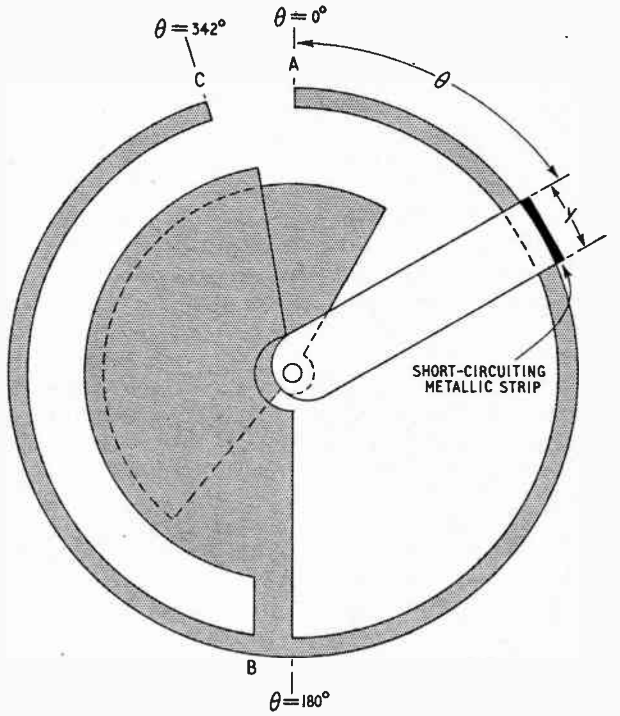
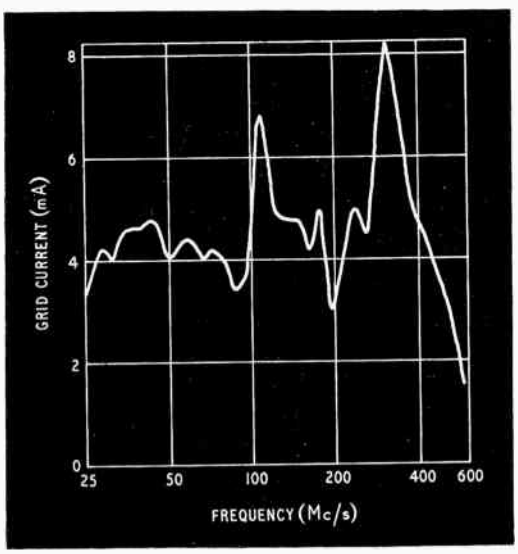


Fig. 7. *The diagram showing the length of the transmission line and the inductor arm for any rotor position θ*

Fig. 6. *Variation of the grid current through the fixed resistor* (8000 Ω) *over the frequency range of the oscillator*



Acknowledgment

The above investigations were carried out in the Electrical Engineering Department, Jadavpur University. The author wishes to express his sincere gratitude to Professor J. S. Chatterjee for suggesting the problem and for discussions during the progress of the work and to Professor H. C. Guha for his interest in the work.

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SON ET LUMIÈRE

This does *not* foreshadow a graceful compliment to Greenwich, and I am afraid you will have to look elsewhere for a revelation of the wondrous work by which Diane de Poitiers booms nightly in rich purple baritone from every resonant cavity in the Chateau of Chenonceux. I think we should have understood one another just as well if I had headed the article simply "Sound and Light". But as I shall be dealing with a case in which the two are often confused in the theory, I hope I may be forgiven for using the fashionable term for confusion, synthesis, amalgamation, or whatever you like to call it when they are mixed together practically and thrown at you by the kWh.

Recalling that the June "Fringe" had ended with a note on nuclear induction, in which a microwave cause produces an audio-frequency effect, I had thought that perhaps I might discuss the radar applications of the Doppler principle from the same point of view, filling out the analogy with such technical details as were available. But I see that the subject has been very capably treated from the technical point of view by Mr. G. E. Beck, of Marconi's, in *Wireless World* for May 1957. The title of his article is "Airborne Doppler Navigation. Radio Application of Well-known Sound Effect". It is the second half of this title that I want to examine in some detail; for the Doppler effect in sound cannot rightfully be applied by analogy to electromagnetic waves, and the effect itself, even if well-known, is probably not generally understood.

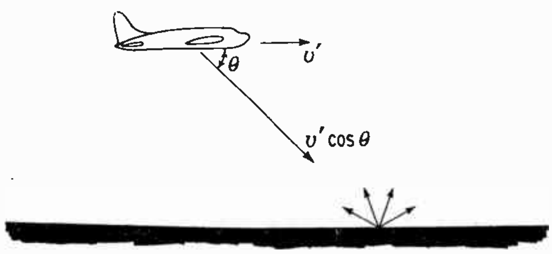

Fig. 1. *From G. E. Beck's article in "Wireless World", May* 1957, *illustrating the symbols used in the text*

There is a long-standing precedent for a little mild confusion, since it appears that Doppler himself tried to deduce stellar velocities from the *colour* of stars, naturally without success. And the major works on sound and acoustics dismiss the topic rather briefly, as accounting for the change in pitch of a locomotive whistle and therefore a matter of only passing interest. The first really full discussion I found of the optical case is in G. Joos's "Theoretical Physics"; and there is a good account also in R. W. Ditchburn's book on "Light", with references to recent experimental work. It is no use thinking in terms of an engine's wolf-whistle when you are dealing with light and radio waves; for you are brought back to fundamentals with a jolt—the soundness of the Michelson-Morley experiment, the Lorentz transformation, and the special theory of relativity.

Mr. Beck's article gives an example of what I mean. For, referring to Fig. 1, which represents an aircraft with ground-speed v' projecting radio waves of frequency f at an angle θ to the ground, and receiving the reflected radiation, the frequency difference f_D between the received and emitted waves is worked out to be approximately

$$f_D = \frac{2f}{c} v' \cos \theta,$$

where c is the velocity of light. And, indeed, this is what it really is. But in the first line of the derivation he has the emitted waves approaching the reflector with a speed greater than c; at $(c + v)$, where v stands for $v' \cos \theta$, in fact. Now, this step is basically unsound even in the acoustic case for, once the source has committed a wave to the medium, the subsequent speed of propagation is the same whether the source follows through or not. Yet the result is acceptable; and the trouble with the Doppler effect is that at least with first-approximation formulae, you can get the right answer in the wrong way. To be fair about this, Beck's argument follows very closely that of Alexander Wood's "Acoustics", except for one essential point; Wood never states or implies that the waves travel with any velocity but their own.

At this point, I hope, you prick up your ears—and antennae. Their *own* velocity ? Can there be such a velocity ? I mean, of course, their velocity relative to the medium in which they are propagated. Now, in the case of sound waves, if we choose to measure velocities relative to the source or to the observer we can presumably do so; and there is no restriction on the values thus obtained for the relative velocity. But the usual Doppler-effect sound formula applies to motion of source or observer relative to the medium. In the case of electromagnetic waves we cannot measure velocities of source or observer relative to the medium, nor can we take either one separately as the origin of a fixed reference system; we can only measure their velocity relative to one another. And, however great this relative velocity may be, it can have no effect whatever on the value of c.

The Acoustic Case

For sound waves, we know what the medium is, and can refer the motion of the waves, the source, and the observer all to the air. If there happens to be a wind, then this will not alter the velocity of sound relative to the air, but the 'ground speed' of source and observer must be translated to 'air speed' values, which are the u_s and u_o of this section. The only constant quantities are the true frequency of the waves emitted by the source (f), and their velocity in the air (here called c for ease in comparing formulae later). The diagrams represent wave-fronts separated at wave-length intervals; that is, at time intervals of $1/f$ sec. In Fig. 2, when $u_s = 0$, successive waves spreading out from S form a symmetrical pattern; and the 'true' wavelength λ is given by $c = f\lambda$.

Now let the source alone move, with velocity u_s in a

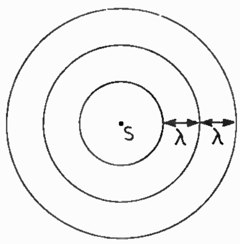

Fig. 2. *Pattern of waves emitted with frequency f from a source S fixed with respect to the medium. The wavelength λ, which is c/f, is the same in all directions*

direction which, at the instant considered, makes an angle θ with the 'line of sight' SO (Fig. 3). Successive waves spread out from centres which are the instantaneous positions of the source at the moment of emission; and their separation (which is the observed wavelength) is now different in different directions. Since in $1/f$ sec, the source has moved a distance $(u_s \cos\theta)/f$ in the direction SO, the wave-length in this direction is $\lambda_1 = \left(\lambda - \frac{u_s}{f}\cos\theta\right)$ or, since $c = f\lambda$, $\lambda_1 = \lambda\left(1 - \frac{u_s}{c}\cos\theta\right)$. But the waves are still travelling with the same velocity c, so that the observed frequency f_1 is c/λ_1, whence

$$f_1 = f\left(1 - \frac{u_s}{c}\cos\theta\right)^{-1} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

It is true that, to the observer, *both* f and λ are changed; this is because c is constant, and the motion of the source has in the first instance altered λ.

Next, suppose that the source is fixed and the observer moves towards the source with velocity u_o, at an angle θ to the line of sight. In each second, $\frac{u_o}{c}\cos\theta$ waves have reached him which would not have done so if he had waited for them at rest. And remembering that $c = f\lambda$, the observed frequency is now

$$f_2 = f\left(1 + \frac{u_o}{c}\cos\theta\right) \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

The physical result is different, for the first-instance effect is a change in frequency. And *the numerical result for a given relative velocity of source and observer is different in the two cases.* For example, take $\theta = 0$, and $f = 550$ c/s. If $u_o = 0$ and $u_s = 100$ ft/sec, as $c = 1\,100$ ft/sec, then $f_1 = 605$ c/s; while if $u_s = 0$ and $u_o = 100$ ft/sec, then $f_2 = 600$ c/s. We said at the start that u_s and u_o were measured relative to the air; it is clearly not sufficient simply to use the velocity of source and observer *relative to one another*, as the same relative velocity gives two quite different results. The whole validity of this argument rests on our ability to measure u_s and u_o relative to the medium which carries the waves.

The next point to notice is that, while the difference between 550 and 605 is a first-order effect, corresponding to a rise in pitch of about a tone, the difference between 600 and 605 is a second-order effect, a little less than what the musicians call a 'comma'. The smaller the value of u/c the less important do second-order effects become; and, indeed, if u/c is small enough, we can write the right-hand side of either (1) or (2) simply as $f\left(1 + \frac{u}{c}\cos\theta\right)$, where u stands for u_s or u_o at will, or $u\cos\theta$ stands for the relative velocity along the line of sight. It looks perhaps as if there has been a lot of fuss about nothing after all; but there are cases in which the second-order effect is important.

The 'echo' formula of Mr. Beck's article is obtained by combining (1) and (2), writing $u_s = v'$ and $u_o = v'$, when the observed frequency is

$$\begin{aligned} f_3 &= f_1\left(1 + \frac{v'}{c}\cos\theta\right) \\ &= f\left(1 + \frac{v'}{c}\cos\theta\right)\left(1 - \frac{v'}{c}\cos\theta\right)^{-1} \\ &= f\left(1 + \frac{2v'}{c}\cos\theta\right) \text{ approximately,} \end{aligned}$$

whence $f_D = (f_3 - f) = \frac{2f}{c}v'\cos\theta.$

It is translated to the radar case by taking c as the velocity of light.

Before this stage in the discussion is reached, the books on sound rather lose interest. A. B. Wood invites the reader to have fun with the formulae if he wants to. Rayleigh sets a good example in this respect by toying with the idea of giving a piece of music sufficient start, pursuing it at $u_o = 2c$, and then hearing it all back-wards—cruelly widening the already vast domain of pleasures that are denied to the tone-deaf. This kind of exercise is all very well, but there are sterner tasks ahead.

Formulae and Frequency

Although wavelengths were used in deriving formula (1), the final formulae of both (1) and (2) apply equally

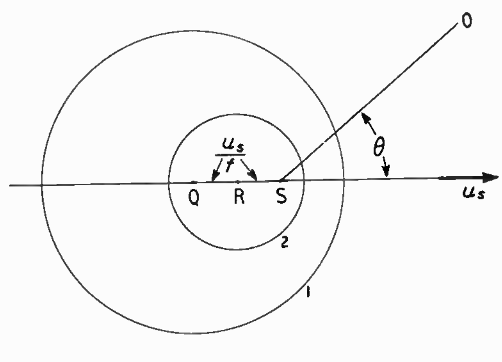

Fig. 3. *Pattern of waves emitted with frequency f from a source moving with velocity u_s relative to the medium. Here S is about to emit one; wave 1, centred symmetrically on Q, was emitted when S was at Q; wave 2, similarly centred on R, was emitted when S was at R. The distances SR = RQ = u_s/f. The wavelength is shortened by u_s/f ahead of S, and lengthened by u_s/f in its rear; while in the direction SO the wavelength is shortened by $(u_s\cos\theta)/f$*

well to the *frequency* of *pulses*. This point is taken up by Mr. Beck, and is an argument for using continuous rather than pulsed radiation for navigational purposes. In fact, the formula (2) for *pulses* antedates the explicit statement of Doppler's principle by about 150 years. For, the pulses being the light-signals indicating eclipses of a satellite of Jupiter, and u_0 being the orbital speed of the earth, this is the way in which Romer first obtained a value for c, by observing the change in the *frequency* of the eclipses as θ varied during the year. These observations, given in I. B. Cohen's monograph on Romer, were made in 1671 !

The Michelson-Morley Experiment

The argument of the last section but one can be looked at in another way. If it were possible, working with great enough source and observer velocities, to determine optical Doppler shifts with sufficient accuracy, the existence or otherwise of the 'fixed ether' could be substantiated. If the second-order difference between the results of equations (1) and (2) were indeed detected, then this would shake the very foundations of the theory of relativity. Acceptance of relativity, then, forecasts that if it could ever be done, then this kind of experiment, however carefully performed, would give a negative result. The positive experimental basis for this belief is a negative result obtained from an experiment of a different type, the attempt to observe displacement of optical interference fringes between two light beams which had traversed paths of equal lengths in directions at right angles to one another. Motion of the earth in its orbit through a fixed ether carrying the light-waves would cause a difference in the times taken for the two paths, and a consequent shift in the fringe pattern depending on the orientation of the apparatus. In the original experiment of Michelson and Morley (about 1880) no fringe shift was detected.

Two suggestions made in the paper of Michelson and Morley do not seem to have been carried further. The first was to repeat the experiment at a very high altitude, and the second was to look for a second-order effect in the frequency of eclipses of Jupiter's satellites.

The question as to whether with more refined methods of observation any shift at all could be found has been taken up at intervals more recently. That expected for a fixed ether with the apparatus of Michelson and Morley was of the order of a whole fringe; G. Joos, in 1930, showed that any effect there might be was less than a thousandth of this, while on the other hand substantial shifts (up to about 8 per cent of that expected) were reported by D. C. Miller. The experiments of Dr. L. Essen at the National Physical Laboratory (*Nature*, 7th May, 1955) using microwaves in a cavity resonator failed to detect any significant change in the resonant frequency with orientation, confirming the original conclusion and suggesting that Miller's result was too high by at least a factor of 10. The point that faces us then is, that the Michelson-Morley experiment has stood the test of very stringent investigation; all attempts to detect an ether, or motion relative to it, have failed. And there is no doubt that the acoustic-analogy treatment of the Doppler effect in optics is therefore unsound.

The Lorentz Transformation

Consider two 'observers' 0 and 0′, each aware of a given event. Let the velocity of 0′ relative to 0 along the line 00′ be u, and take 00′ as the common x-axis for co-ordinate systems referred to 0 and 0′ in turn. The co-ordinates of the event in space and time, referred to 0, are x, y, z, and t. For the same event, referred to 0′, they are x', y', z', and t'. Then, if c is the velocity of light, and writing the ratio u/c as β,

$$x' = (x - ut)\ (1 - \beta^2)^{-\frac{1}{2}} \quad \ldots \quad \text{(a)},$$

and

$$t' = (t - ux/c^2)\ (1 - \beta^2)^{-\frac{1}{2}} \quad \ldots \quad \text{(b)}.$$

These formulae are called the Lorentz transformation. The result of the Michelson-Morley experiment; the special theory of relativity; and the invariable constancy of the velocity of light, are all involved in these expressions. The full derivation is given both by Joos and by Ditchburn. From (b) the well-known Einstein time-dilatation results; this has been verified directly by observations on the decay of π-mesons in flight, for it is found that the faster they travel the longer they survive in the observer's time-scale.

The Radial Doppler Effect

Using the Lorentz transformation we can, without knowing any more about the propagation of the light than that it travels in all directions with a constant and

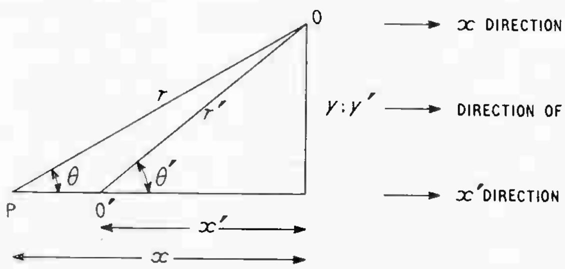

Fig. 4. *Illustrating the Doppler effect. The observer O is at rest in the x, y, t system, and the source O′ at rest in the x′, y′, t′ system. The velocity of O′ relative to O is u, in the common x-direction. Here r′ and θ′ are the polar co-ordinates of O when O′ is origin, and r and θ the corresponding values in O's system, so that the apparent position of the source, as seen from O, is at P*

unalterable velocity c, and without asking which of the two we ought to imagine to be fixed, obtain an expression for the observed frequency in terms of the *relative velocity* u of source and observer, which is *not* a first-order approximation. Here 0′ is taken as the source, and 0 the observer.

First, in a frame of reference fixed with respect to 0′, let it send out a spherical wave of frequency f', represented by

$$S' = \frac{A'}{r'} \exp\left[2\pi j f'\,(t' - r'/c)\right]$$

Everything in the source's frame of reference is denoted by a prime.

The observer 0 is at rest in his own frame of reference, in which his co-ordinates are x and y, the x-direction being that of the relative motion of the two systems. His co-ordinates in the source's frame of reference are x' and

y'. He will receive a wave of frequency f, where

$$S = \frac{A}{r} \exp\,[2\pi jf\,(t - r/c)].$$

This will appear to him to be travelling at an angle θ to the x-axis, while from the source's point of view the angle is θ' (Fig. 4). In the 0′ frame, $r' = x' \cos\theta + y' \sin\theta'$, while in the 0 frame, $r = x\cos\theta + y\sin\theta$.

Then, substituting for r and r', in the two frames of reference,

$$S' = \frac{A'}{r'} \exp\,[2\pi jf'\,\{t' - (x'\cos\theta + y'\sin\theta')/c\}]$$

and

$$S = \frac{A}{r} \exp\,[2\pi jf\,\{t - (x\cos\theta + y\sin\theta)/c\}]$$

Now, using the Lorentz transformation, we can write in the expression for S',

$$x' = (x - ut)\,(1 - \beta^2)^{-\frac{1}{2}}$$
$$y' = y$$

and $t' = (t - ux/c^2)(1 - \beta^2)^{-\frac{1}{2}}$. To apply these results, the relative motion need not be along 00′, as it was earlier; it has only to be in the x-direction, which was previously taken to coincide with 00′ but no longer does so.

Also, the expressions for S and S', which describe the same wave, must in reality be the same whatever co-ordinate system they are written down in. The condition for this is that, after substitution has been made for x' and t', the x and y and t indices are separately equal in the two expressions. Writing down only the exponents,

$$2\pi jf'\left[(t - ux/c^2)\,(1 - \beta^2)^{-\frac{1}{2}} - \left(\frac{x - ut}{c}\right)\cos\theta'\,(1 - \beta^2)^{-\frac{1}{2}} - \frac{y\sin\theta'}{c}\right] = 2\pi jf\left[t - \frac{x\cos\theta}{c} - \frac{y\sin\theta}{c}\right]$$

whence

$$f = f'\left(1 + \frac{u\cos\theta'}{c}\right)(1 - \beta^2)^{-\frac{1}{2}} \quad \ldots \quad \ldots \quad (3)$$

In equation (3), then, f' is the frequency emitted by the source; f is the frequency received by the observer. We do not speak in terms of 'true' and 'apparent' frequencies, for these words have no meaning; or, rather, both f and f' are 'true' although they are different.

To a first approximation, formula (3) gives the same result as (1) and (2). For, if $u \ll c$, then $\beta^2 \ll 1$, and $(1 - \beta^2)^{-\frac{1}{2}} \approx 1$. Thus

$f = f'\left(1 + \frac{u}{c}\cos\theta'\right)$, which, with the necessary change of symbols is, in fact, the same as the approximate form into which both (1) and (2) were fitted. But now we are quite certain what u represents. It is the relative velocity of approach of source and observer. It has been rather a long job to establish this; but there is no other way of doing it satisfactorily.

The Transverse Doppler Effect

The expression for θ obtained from the exponent equation is

$$\cos\theta = \frac{\cos\theta' + u/c}{1 + u/c\,\cos\theta'}$$

If $\theta = 90°$, then $\cos\theta' = -u/c$, and from equation (3),

$$f = f'\,(1 - \beta^2)^{-\frac{1}{2}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

There is thus a Doppler-effect difference between the observed and the emitted frequencies due to relative motion *at right angles to the line of sight*. Now, in the acoustic case, if θ is 90°, then the two frequencies are equal, and that is all there is to it. The existence of the transverse Doppler effect should show directly that the relativity treatment is in fact correct in the optical case, and the acoustic analogy wrong. In Joos' book, the possibility of looking for this effect is mentioned; Ditchburn gives a reference to the work done in 1941 by Ives and Stilwell, in which they verified formula (4) using as moving sources a carefully collimated and homogeneous molecular beam. I have not so far been able to consult their paper myself, and have no idea of the accuracy these workers obtained; but to do such an experiment at all is a tremendous feat. And the result, as a *direct* confirmation of the special theory of relativity, is important.

To conclude, then, you can take what algebraic liberties you like, to a first approximation, with a first-approximation formula once you have got it. But, while on the way there, it is important to realize when you are compromising in the matter of fundamentals, even though in nearly every optical or radio application (the rather involved matter of the astronomer's red-shift is one exception) u is so much less than c that the approximate formula suffices. The popular conception of the Doppler effect, and its hackneyed illustration, are summed up concisely in the words of Orsino: "That's train again. It hath a dying fall". I hope that, without giving you excess of it, I have at least shown that there is a good deal more to it than many people believe.

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National Radio Exhibition

EARLS COURT 28th AUGUST—7th SEPTEMBER 1957

From the point of view of the public the National Radio Exhibition is *the* radio show. The B.B.C. and I.T.V. stage shows designed to attract the public and to give it some little insight into the working of a television studio. The exhibits, too, are displayed with the aim of attracting the public.

The exhibition is concerned almost entirely with domestic apparatus, and television holds pride of place, as it has done for a good many years now. This year, however, there was a much greater emphasis upon apparatus for the high-quality reproduction of sound, or high fidelity, as it is more generally called. The increasing interest in the gramophone record was reflected, too, in the large number of record players on view.

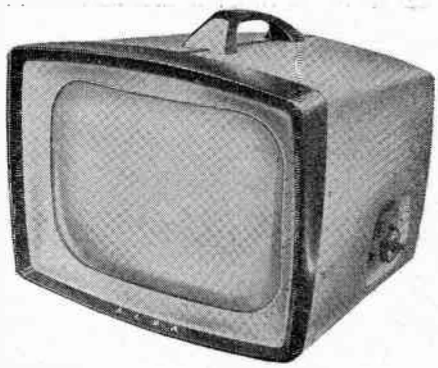
The only obvious development in television receivers is the trend towards larger tubes. The 21-in. tube is now a commonplace and the 17-in. appears to be the popular one. Except for portable sets the 9-in. tube is obsolete and the 12-in. very nearly so ; even the 14-in., which only a few years ago seemed large, appears to be on the way out. There are still quite a lot of them and, actually, a 14-in. picture is large enough for many circumstances, but there are more 17-in. Some makers, indeed, have already abandoned it and market only sets with 17-in. and 21-in. tubes.

These tubes invariably have rectangular faces and for the 21-in. the general practice is to employ a 90° deflection angle. This considerably shortens the tube, as compared with the older 70° type, and so enables a shallower cabinet to be used. This not only reduces the cost but results in an article of more acceptable dimensions. The greater angle, of course, makes scanning more difficult and more carefully designed deflector coils are needed to secure satisfactory focusing.

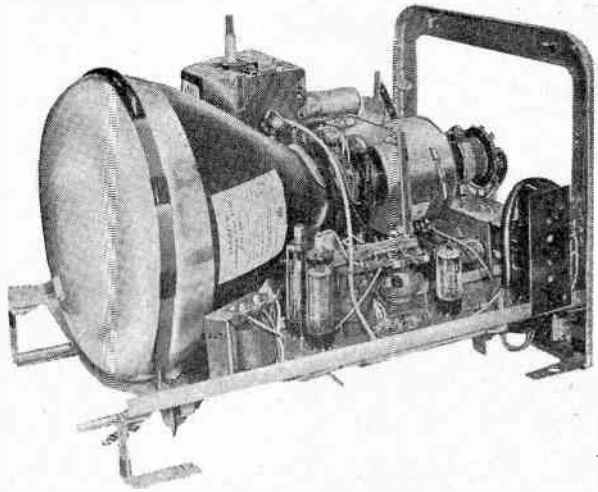
This 90° deflection angle is also being adopted for 17-in. tubes but here it is not yet general, most sets still retaining 70° types.

The cult of the portable or transportable television receiver is spreading, many more examples being found this year than last. The smallest has a 9-in. tube and in the Spencer-West version the weight has been got down to 17 lb. Most sets of this type, however, have 12-in. or 14-in. tubes. In most cases the circuitry is much the same as in standard sets and they differ mainly in three things, a rather smaller picture, the provision of a rod aerial on the case, and the simple container. Instead of cabinet work, the case is imitation leather or frankly metal.

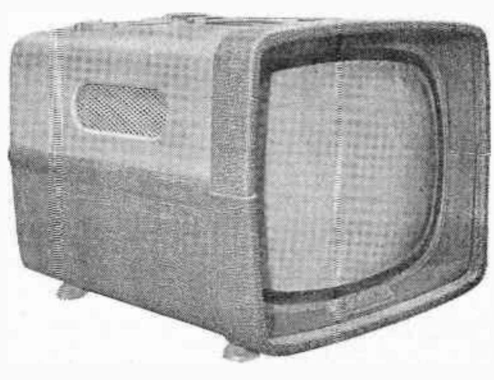
An examination of the circuits employed in this year's television receivers reveals only minor changes. 'Front-ends' are virtually standard with a cascode r.f. stage



Alba T909 transportable with 14-in. tube and rod aerial; the weight is 27 lb.



Chassis of Spencer-West portable set with 9-in. tube



H.M.V. transportable television receiver model 1864

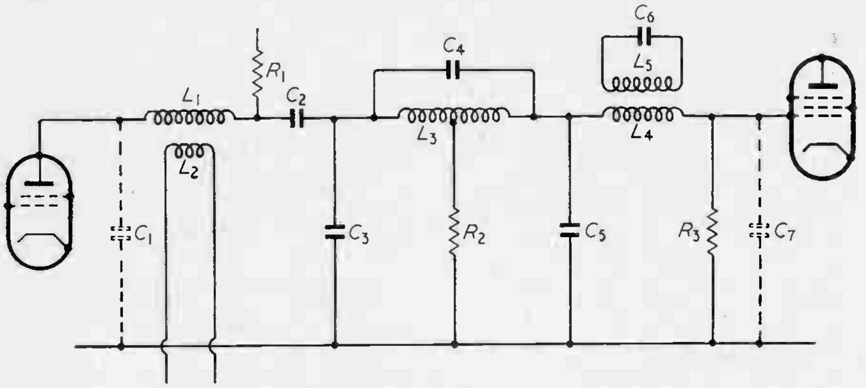
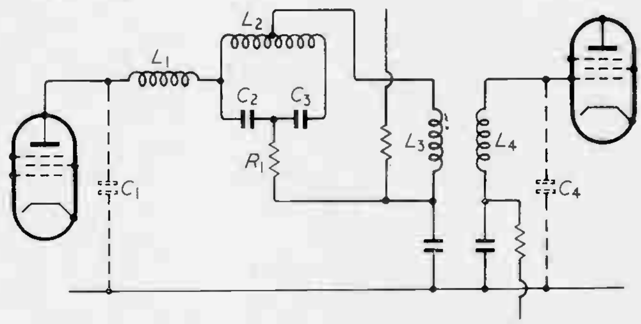


Fig. 1. *Intervalve coupling used in Pye television receiver including a bridge-T sound rejector* $L_3C_4R_2$

Below: Fig. 2. *Ferguson 'adjacent-sound' rejector circuit* $L_2C_2C_3$ *in i.f. amplifier*



embodying a double-triode valve and a triode-pentode frequency-changer. There are generally three tuned circuits at signal frequency, a coupled pair being used between the two stages. With the oscillator, this makes four coils which need changing for channel selection and the general practice is to do this by means of a turret. Some makers, however, notably Pye and Invicta, use the incremental-inductance method with switches and Bush retain the permeability tuner. Whatever method is adopted the result is the same from the user point of view for he is invariably provided with a 12-position station selector with a clicker mechanism. It is the general practice, too, to fit a fine tuner operating on the oscillator circuit ; however, Murphy have now abandoned this for their current sets for they claim to have achieved adequate oscillator stability for it to be dispensed with.

The only decided trend in circuit development this year is one which some may think trivial but which may, nevertheless, have the effect of decidedly improving performance. This trend is one towards the use of bridged-T networks in the vision i.f. amplifier to help in sound-channel rejection. It is not the use of the bridged-T network for this purpose that is new, of course, for there have been one or two examples to be found in most previous years. What is new is the number of people who now include it and the variety of forms which it takes.

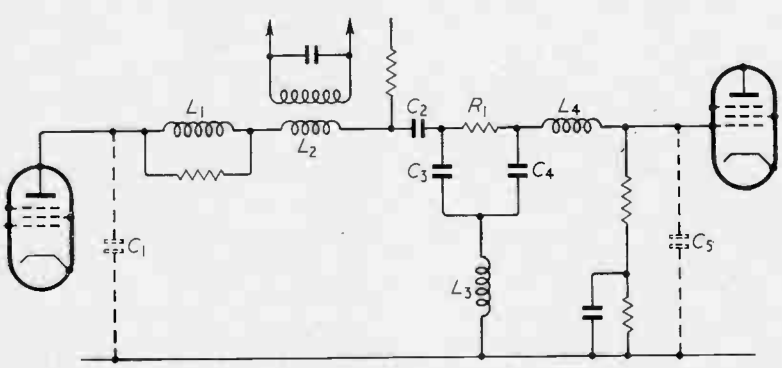
An example which at first sight appears particularly complicated is used in the Pye sets. It is shown in Fig. 1 and the bridged-T network comprises $L_3C_4R_2$. There is a double transmission path through it, by C_4 on the one hand, and by L_3R_2 on the other. As a result, the circuit is analagous to a balanced bridge and at one frequency gives no transmission. It behaves much as a parallel tuned circuit of infinite Q.

This T network is used as the coupling element between two resonant circuits $L_1C_1C_2C_3$ and $L_4C_5C_7$. With R_2 removed and C_4 short-circuited, the circuit would be that of a pair of tuned circuits with 'bottom-end' capacitance coupling by C_3 and C_5 in parallel. A further rejector of normal type L_5C_6 is coupled to the secondary, while L_2 forms part of another tuned circuit coupled to the primary and utilized to feed the sound channel.

The sound pick-out circuit L_2 is tuned for maximum signal in the sound channel with an input at 38·15 Mc/s. The trap L_5C_6 is tuned for minimum output from the vision channel with an input at 38·65 Mc/s, while L_3C_4 is adjusted for minimum output at 38·15 Mc/s. The coupling elements are tuned for maximum output, L_1 at 37 Mc/s and L_4 at 35·5 Mc/s.

Ferguson adopt a somewhat different arrangement which is sketched in Fig. 2. The trap is $L_2C_2C_3$ with R_1 for a resistance balance. In this case the trap is tuned to 33·15 Mc/s to reject the sound signal of the adjacent channel. The 38·15-Mc/s trap has the different form

Fig. 3. *Modified circuit used by Ferguson for 'own-sound' rejection*



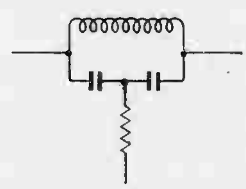


Fig. 4. *Kolster-Brandes bridged-T rejector*

shown in Fig. 3. The circuit is that of a coupled pair $L_1L_2C_1C_2$ and L_4C_5 in which L_2 is part of the sound pick-out circuit. The trap comprises $L_3C_3C_4$ with R_1 for a resistance balance, and the trap forms the coupling element between the tuned circuits.

Kolster-Brandes have always rather favoured the bridged-T trap and retain it in the simple form of Fig. 4 as a 'top-end' coupling element between a pair of tuned circuits.

Still another form is adopted by Stella, Fig. 5, who have a coupled pair of circuits L_1C_1 and L_2C_4, the coupling being by the trap $L_3C_2C_3$. The resistance balance is obtained by R_1 between the 'hot' ends of the coupled circuits.

In most cases the bridged-T traps are supplemented by one or more traps of the ordinary kind and some manufacturers still rely only on this type.

It is a characteristic of the bridged-T trap that its bandwidth at high attenuation is very narrow indeed and so it would appear to demand very high oscillator stability and accurate tuning if the sound signal is to be kept within its bandwidth. This is why probably more use has not been made of it in the past.

As it is now used, however, its function is more to give a very sharp cut-off at the edge of the pass-band than to provide a narrow notch in the response curve. The property of zero transmission at one frequency enables the L/C ratio to be chosen for a minimum effect in the pass-band while, because of the resistance balance, very high attenuation just outside the pass-band can be obtained. Still further from the pass-band the response is prevented from rising unduly by other trap circuits of more normal type. The bridged-T circuit, in conjunction with ordinary traps and the natural selectivity of the intervalve couplings, enables a sharp transition to be secured between the pass and attenuating regions.

Apart from the trap circuits, i.f. amplifiers are substantially unaltered. It is the normal practice to use two wideband stages of which the first is common to both vision and sound channels, followed by one further stage in each separate channel.

This tendency towards a common i.f. stage is new, for until recently it was the general practice to separate the sound and vision signals immediately after the frequency-changer. Some sets, notably fringe-area models, have more i.f. stages. Thus, Invicta have two vision and two sound-channel stages in addition to a common stage in the model 138, while Pye have one set with three stages in each channel but, in that, the channels are separated in the mixer output.

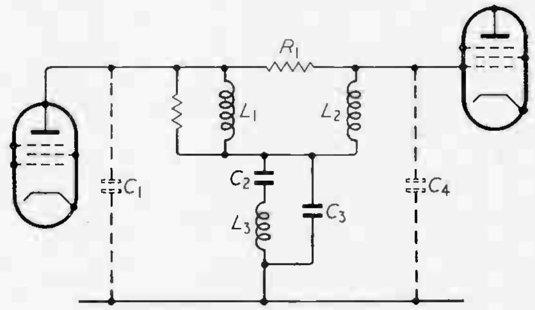
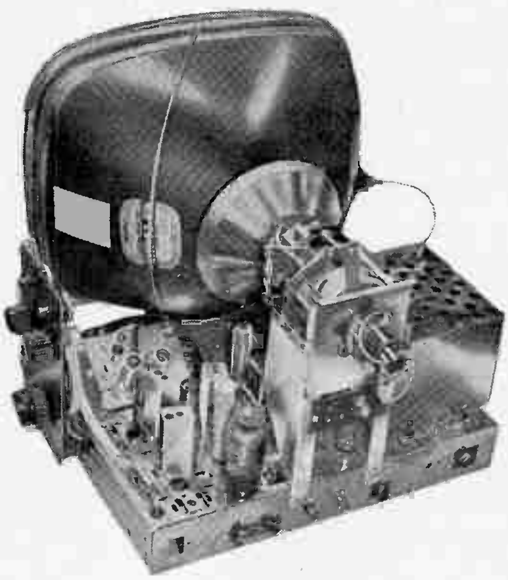


Fig. 5. *In the Stella rejector the resistance R1 offsets the coil losses of the trap circuit*

Ecko TCG336 receiver with 17-in. tube, v.h.f. radio and record player

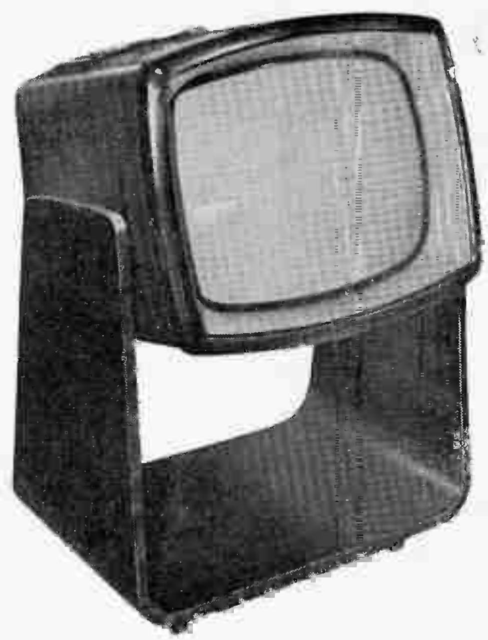


Chassis of McMichael receiver with 17-in. tube

Diode detectors in both vision and sound channels are universal and further diodes are normal for ignition-interference suppression. The video stage is a pentode but some makers add a triode cathode-follower ; a triode-pentode valve is then invariably used for the two stages. The cathode-follower not only provides a low output impedance but, by relieving the video amplifier of the tube capacitance, enables it to give more gain.

The video signal is normally fed directly to the cathode of the c.r. tube and via a capacitor to the grid of a pentode sync separator, d.c. restoration being achieved by grid current. The mean grid current of this valve is now quite often used to develop an a.g.c. voltage for the vision receiver. The a.g.c. system so obtained is no more complex than, and in its details is very similar to, that of any ordinary sound receiver. Unlike the latter, however, the control voltage is not dependent only on the strength of the signal. The television signal does not have a constant mean level of carrier, only a constant black level. As a result, the mean grid current of the sync separator, and hence the a.g.c. voltage, depend upon the mean picture brightness as well as upon the signal strength. An a.g.c. system of this type, therefore, tends to keep the mean brightness constant irrespective of whether the changes which it corrects are brought about by signal variations or by changes of brightness in the transmitting studio.

In practice, the system does work quite well, presumably because the mean brightness of television pictures does not vary very much. Some designers, however, prefer something theoretically more perfect and incorporate gated a.g.c. In this the video signal is sampled during its intervals at black level, usually during the back porch, and the samples are integrated to give a control voltage.



Murphy V310C with 17-in. tube; the receiver is pivoted on its stand so that the viewing angle can be adjusted

Until a few years ago a.g.c. was not much used in British television and when it was introduced the mean-level type was the most popular, but a good many varieties of the gated forms were used. It now appears that gated a.g.c. is much less widely used; being now confined mainly to fringe-area models. Mean-level a.g.c. is relatively common. One of its major functions is to keep the tube input more or less constant when switching between B.B.C. and I.T.V. channels. It is not necessary to use a.g.c. for this, however, and Murphy do not in their model 310; instead, there are two adjustable cathode-bias resistors for the r.f. stage which are connected in circuit by a switch linked to the turret tuner. They can thus be pre-set to equalize the receiver outputs on the two channels. This is a simple scheme which has much to commend it.

Timebase circuits remain substantially unchanged in their paper form but a good deal of development work on them has actually taken place to obtain increased output. As already mentioned, cathode-ray tubes with with a total deflection angle of 90° are coming into use, and more scanning volt-amperes are consequently needed than for the 70° type. To a large extent this is obtained by improved component design, particularly of the line-scan transformer. However, the increase needed is being minimized by improvements in deflector coils. The net result is that circuits are hardly changed and provide the extra output with little or no increase of input power.

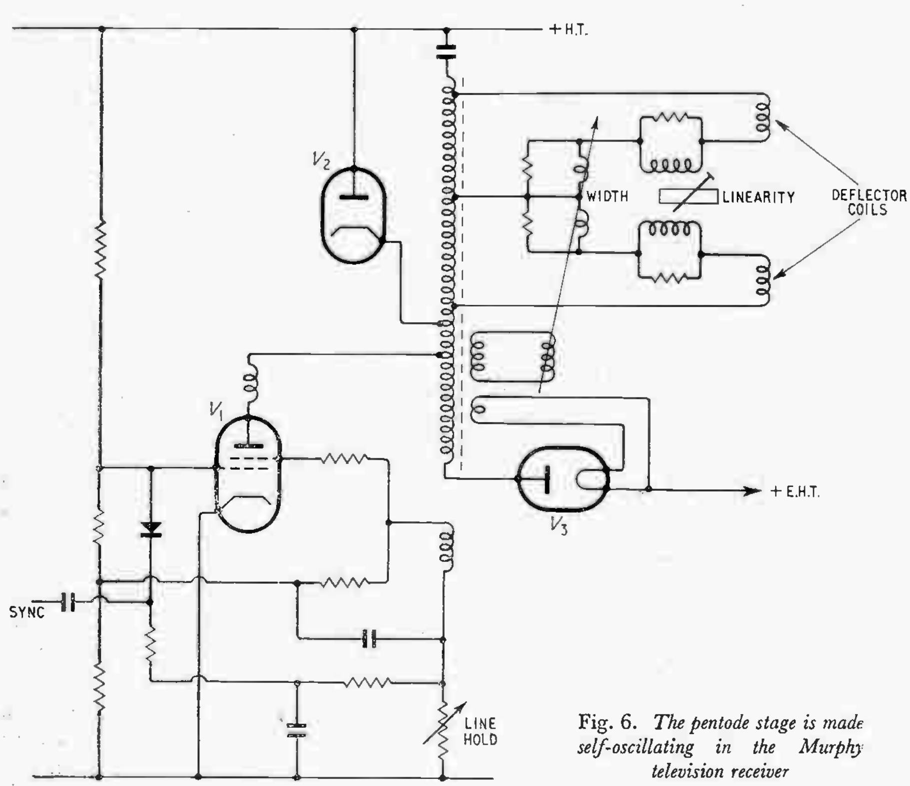
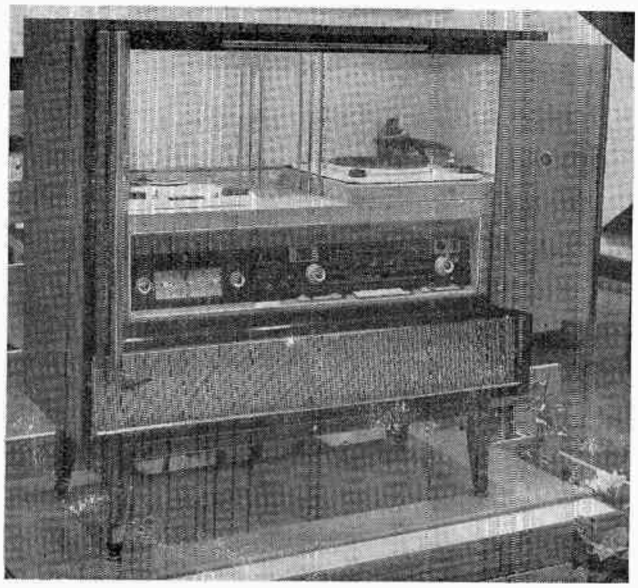


Fig. 6. *The pentode stage is made self-oscillating in the Murphy television receiver*

The general practice is to use a pentode auto-transformer coupled to the deflector coils with an energy-recovery diode and a further diode to provide e.h.t. from the line flyback. Linearity control is usually obtained with the aid of a ferrite-cored coil, the core being partially saturated by a small permanent magnet. The pentode is driven at its grid by a blocking oscillator or other saw-tooth generator.



Ferguson model 403 *RG with automatic record changer and tape deck*

There are exceptions, however, and an interesting one is the Murphy in which the pentode stage is made self-oscillatory by coupling between the anode and the screen-grid. The 'single-valve' timebase has always had its adherents but has never come into widespread use. The Murphy circuit, shown in Fig. 6, is somewhat different from most of the earlier 'single-valve' circuits. The anode circuit of V_1 is conventional and differs in no essential from other timebases of the driven type; V_2 is the energy-recovery diode and V_3 the e.h.t. rectifier. An extra winding on the transformer is included in the screen-grid circuit. During the scan the voltage across this winding is substantially constant and with a polarity such that the screen-grid is driven positive to the cathode. During flyback there is a relatively large negative pulse. The voltage waveform is, of course, the same as that on the anode but reversed in polarity and reduced in amplitude. A potential divider across the h.t. supply applies a small positive bias to both control and screen grids, but the main positive voltage comes from the transformer winding. This voltage on this winding is also integrated and applied to the control grid.

It is interesting to see that, this year, designers are tending to revert to direct locking of the line timebase by the differentiated sync pulses. The trend towards flywheel sync appears to have ceased and, generally speaking, it is now employed only in fringe-area models. No uniformity in frame-pulse separation methods is yet evident and designers still differ widely about the best way of achieving frame locking.

Line and frame flyback suppression pulses are now commonly applied to the c.r. tube. They are derived through simple *RC* networks from suitable points in the timebases. Only one difference from previous years has been noted; the pulses are now more commonly applied to the first anode of the c.r. tube than to the control grid. Spot wobble remains a feature of Ekco sets.

The television set is, of course, particularly easy to adapt for v.h.f. sound reception since the sound i.f. bandwidth is of the right order of magnitude for f.m. Quite a lot of sets exist which provide this additional facility. A few extra coils are needed in the turret tuner and switching must be provided to change the sound detector from the usual simple diode of television sound to a ratio detector. Provision is usually made also for switching the heaters of the purely television valves out of circuit.

Sound receivers for the v.h.f. band only are still something of a rarity. Band II reception is usually provided as an adjunct of television, as already mentioned, or of the medium and long wave a.m. receiver. In these the basic receiver is of conventional type with a triode-heptode frequency-changer, one i.f. stage at around 465 kc/s, detector and audio amplifier. For Band II the oscillator is switched off and the heptode acts as an extra i.f. stage; the i.f. transformers are switched to 10·7 Mc/s and the detector is changed to a ratio detector. The whole is preceded by a double-triode one half of which acts as an r.f. amplifier and the other as a self-oscillating mixer.

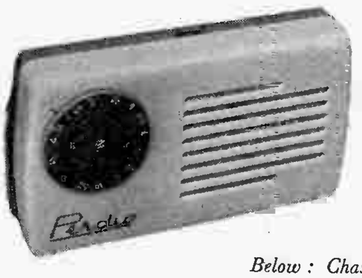
A few purely v.h.f. receivers do exist, however. Ekco have models of this type, and so have Dynatron. The latter is also available in chassis form as a 'tuner' for high-fidelity equipment. It is the FM2LV and has a tuning range of 88–100 Mc/s with switch-operated pre-tuned circuits; a.f.c. is incorporated with a control range of $\pm$ 300 kc/s. A Foster-Seeley discriminator and limiter are included.

Vidor portable covering the v.h.f./f.m. band as well as medium and long waves

The "Hi-Fi" cult has led many firms to produce power amplifiers of high grade with outputs of 5 W upwards and a wide range of speaker systems. We use the word systems advisedly for so many of them are now a cabinet housing combinations of loudspeakers to cover different portions of the audio spectum. The cabinet too is rarely just a cabinet. Internally, it is what might be described as an audio-frequency resonator system designed to accept the sound from the loudspeaker and convey it to the external air in such a way that resonances are minimized and the response maintained smoothly to quite low frequencies.

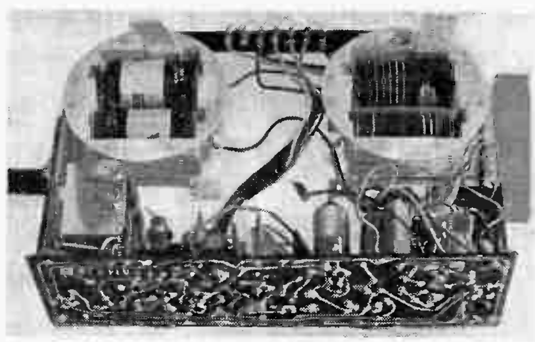
In most cases Hi-Fi equipment is available in the form of units, such as loudspeakers, loudspeaker cabinets, amplifiers, pre-amplifiers, mixers, tuners, so that the enthusiast can select just what he needs. The units are not always bare chassis, however; they are often available

Bush *RP*21 *record player with* 4-*speed motor and automatic changer*



Left: *Perdio transistor portable receiver*

Below: *Chassis of Pam transistor portable showing printed circuit and battery compartments*



in nicely finished cases and it is possible to build up quite a presentable equipment with a minimum of labour.

The portable gramophone has always been popular and the portable electric record player has been increasingly so in recent years. Now a glance round the exhibition reveals that there are almost as many types as there are radio sets. Although physically small, they often incorporate four-speed motors and sometimes auto-changers! Mostly mains operated there are quite a few battery types using transistor amplifiers.

Portable receivers, too, are to be found in great numbers and here the transistor is making great headway. One set, which was claimed to be the smallest at the exhibition is the Perdio which measures only $5\frac{3}{4}$ in. by $3\frac{1}{4}$ in. by 1 in. and weighs 13 oz. This is a truly miniaturized set and it is not the general practice, even when using transistors, to adopt miniaturization. The sets are smaller than valve sets but more because transistors are smaller than valves and need a smaller volume of battery than for any other reason. Sets of the portable type are generally smaller than in previous years and this is to a large extent due to the quite widespread use of printed circuits. These are now rapidly coming into general use in all types of equipment, but there are still more sets which do not use them than do. Last year Pye adopted printed circuits almost entirely in a television receiver, obviously with success since they have retained them this year. Few other people have gone as far as this, but a great many are using them for sub-assemblies and a set in which the major part comprises ordinary wiring may include, say, an amplifier with a printed circuit.

The exhibits at Earls Court are of domestic apparatus. In addition to their technical side, therefore, they must be considered as pieces of furniture. A television set or a broadcast receiver is used in a living room and is required to harmonize with a furnishing scheme. Quite a few sets are offered with alternative cabinet styles, but by no means all. There is a vogue this year for the so-called Continental styling having a near-black case with lots of brass work as ornamentation. The conventional polished wood cabinets are to be found on most stands, however, and have the great merit of fitting unobtrusively into most furnishing schemes except perhaps for the extreme in 'contemporary'.

MEETINGS

I.E.E.

16*th October*. "Some Radio Aids for High-Speed Aircraft", by J. S. McPetrie, Ph.D., D.Sc. (Chairman's address, Radio and Telecommunication Section).

28*th October*. "Domestic High-Fidelity Reproduction", by J. Moir.

5*th November*. "The Design of the Control Unit of an Electronic Digital Computer", by M. V. Wilkes, M.A., Ph.D., F.R.S., W. Renwick, M.A., B.Sc. and D. J. Wheeler, Ph.D., "A Decimal Adder using a Stored Addition Table", by M. A. MacLean, M.Sc. and D. Aspinall, B.Sc. "An Accurate Electroluminescent Graphical Output Unit for a Digital Computer", by T. Kilburn, M.A., Ph.D., D.Sc., G. R. Hoffman, Ph.D., B.Sc. and R. E. Hayes, M.Sc.

These meetings will commence at 5.30 *at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.*2.

Brit. I.R.E.

30*th October*. "Scatter Propagation", by M. Telford, B.Sc., at 6.30 at the London School of Hygiene and Tropical Medicine, Keppell Street, Gower Street, London, W.C.1.

The Television Society

17*th October*. Discussion on "Servicing Modern Television Receivers".

31*st October*. "Performance of Television Receiver Turret Tuners", by K. H. Smith.

These meetings will commence at 7 *o'clock at the Cinematograph Exhibitors Association*, 164 *Shaftesbury Avenue, London, W.C.*2.

Institute of Physics

15*th October*. "Dense Electron Beams", by B. Meltzer, B.Sc., Ph.D., to commence at 5.30 at the Institute of Physics, 47 Belgrave Square, London, S.W.1.

Physical Society

15*th October*. "Present State of Acoustic Theory as applied to Small Rooms", by J. Moir, to be held at 5.30 in the Physics Department, Imperial College of Science and Technology, S. Kensington, London, S.W.7.

Waveguide Characteristics

EXACT FORMULAE AND CURVES FOR THE ATTENUATION AND PHASE PROPAGATION COEFFICIENTS

By A. E. Karbowiak*

SUMMARY. *Application of the surface impedance approach to the analysis of the wave propagation in parallel plane waveguides, rectangular waveguides and circular waveguides is shown to lead to a general expression for the propagation coefficient,* $\gamma = (\alpha + j\beta)$; *viz.,* $\{\gamma_0^2 + 2\,M\,(j\,Z_s)\}^{\frac{1}{2}}$, *where* γ_0 *is the propagation coefficient of the perfect waveguide,* M *is a coefficient depending on frequency and waveguide geometry and* Z_s *is the surface impedance of the waveguide walls. This expression is shown to hold above and below as well as at the cut-off frequency.*

α *and* β *characteristics of various waveguides are also obtained and extreme cases (very high and very low frequencies) are discussed in detail.*

Lord Rayleigh[1] may truly be said to have pioneered the theory of waveguide propagation; his theory of perfect waveguides (waveguides whose walls are perfectly conducting), however, found little practical application mainly because of the lack of suitable equipment and apparatus.

In the last few years before the last war the interest in waveguides was revived by a number of investigators[2,3] and the simple theory of perfect waveguides extended in certain cases to metallic waveguides[4,5,6]. Since the war a great deal has been published about the theory of waveguides so that it is impossible even to cite all the references. However, as far as the attenuation and phase propagation of waveguides are concerned fundamentally little progress has been made.

The inadequacy of the conventional formulae is clearly evidenced, for example, by the facts that at the cut-off frequency the formulae give an infinite value for the attenuation coefficient (which must be finite), and that they are inapplicable below the cut-off frequency.

The need for a formula for the propagation coefficient of a waveguide below the cut-off arose for the first time in connection with precision cut-off attenuators. Suitable approximate formulae have been developed in a few isolated cases,[7,8] but here again the formulae are inaccurate in the vicinity of the cut-off frequency.

The need for an attenuation formula that would hold above and below, as well as at the cut-off, has not been satisfied, but recently Kerns and Hedberg[9] have obtained an expression for the attenuation as well as the phase-propagation coefficient of a rectangular waveguide that is valid above as well as below the cut-off frequency but the formula fails to give the correct answer in the immediate vicinity of the cut-off frequency.†

Very recently it has been pointed out that the surface impedance approach[10] should lead to the required formula and suitable expressions for the attenuation and phase-propagation coefficients of a rectangular waveguide have been obtained[11]. These formulae have been shown to hold above and below, as well as at the cut-off frequency.

The purpose of the present paper is to derive, using the surface impedance approach, suitable formulae for the attenuation and phase-propagation coefficients of rectangular and circular waveguides for all modes of operation. The formulae obtained are 'exact' in the sense that they hold throughout the frequency spectrum. The case of plain metallic waveguides is treated in some detail.

1. Fundamental Formulae

The Essence of the Surface Impedance Approach

The surface impedance approach has been the subject of a separate paper[10] and it is not intended here to go through the derivation of the various formulae and consequently frequent reference to that paper will be made. However, to illustrate the method, one simple problem of propagation in a planar waveguide will be worked out ab initio.

Briefly, a wave-mode $M_{m,n,\bullet}$ (the suffix $\bullet$ indicates that the mode is propagated axially in the positive z-direction) is postulated in a cylindrical waveguide. This mode is assumed to be, for a perfect waveguide, a normal mode of the region bounded by the waveguide wall S and all field components are assumed to be proportional to

$$e^{j\omega t - \gamma_0 z} \qquad \text{(1)}$$

$$\left.\begin{aligned} \text{where } \gamma_0 &= j\beta_0 = j\,\frac{2\pi}{\lambda_g} \\ k_0^2 &= h_0^2 - \gamma_0^2 \\ h_0 &= \text{cut-off coefficient} \\ k_0 &= \frac{\omega}{c} = \frac{2\pi}{\lambda_0} \end{aligned}\right\} \qquad \text{(2)}$$

* *Standard Telecommunication Laboratories Ltd*

† W. Schaffeld and H. Bayer, (*Arch. Elektr. Übertragung*, 1956, Vol. 10, p. 90) have recently obtained a more accurate expression for the attenuation coefficient of a circular waveguide

A physical waveguide is imperfect in that its walls, S, are not perfectly conducting. The walls of such a waveguide are imagined to have a surface impedance Z_s whose value depends on the physical nature of the guide wall and sometimes the mode of operation. The field components in such a waveguide are proportional to

$$e^{j\omega t - \gamma z} \qquad \text{(3)}$$

where

$$\left.\begin{aligned} \gamma &= \alpha + j\beta \\ k_0^2 &= h^2 - \gamma^2 \\ h &= h_0 + \delta h \\ \gamma &= \gamma_0 + \delta\gamma \\ \beta &= \beta_0 + \delta\beta \end{aligned}\right\} \qquad \text{(4)}$$

The quantity δh is a function of Z_s and the guide geometry, while the quantity $\delta\gamma = \alpha + j\delta\beta$ is calculable therefrom. Throughout, the quantity Z_s is assumed to be homogeneous and isotropic though it may be a function of frequency. These restrictions, however, are unnecessary but the formulae given hereafter may easily be generalized to take into account a possible anisotropic nature of Z_s. It is, however, essential that the inequality

$$|Z_s| \ll 1 \qquad \text{(5)}$$

be satisfied; this condition, fortunately, is fulfilled in the majority of practical cases.

The Planar Waveguide

Consider an $E_{0n\bullet}$ mode in a planar waveguide as shown in Fig. 1. Evidently the wave has only three field components (E_z, H_y and E_x) and the two relevant ones are given by[10]

$$\begin{aligned} E_z &= \sin hx \\ H_y &= -j\frac{k_0}{h}\cos hx \end{aligned} \qquad \text{(6)*}$$

If now the guide surface S has a surface impedance, Z_s, then by definition

$$-\frac{E_z}{H_y} = Z_s = \frac{h}{jk_0}\tan ha \qquad \text{(7)}$$

Since $h_0 = \frac{n\pi}{a}$, then by virtue of (4) we may approximate to $\tan ha$ $(= a\delta h)$ and we get

$$\delta h = \frac{1}{a}\frac{k_0}{h_0} j\, Z_s \qquad \text{(8)}$$

Consequently Equ. (4) yields

$$\gamma = \alpha + j\beta = \left\{ (h_0 + \delta h)^2 - k_0^2 \right\}^{\frac{1}{2}} \qquad \text{(9)}$$

$$= \left\{ h_0^2 - k_0^2 + \frac{2}{a} k_0 \,(j\, Z_s) \right\} \qquad \text{(9a)}$$

$$\approx \left\{ \gamma_0^2 + \frac{2}{a} k_0 \,(j\, Z_s) \right\}^{\frac{1}{2}} \qquad \text{(9b)}$$

since $|\delta h| \ll h_0$.

This formula for the propagation coefficient of the waveguide is true throughout the frequency spectrum and can be used provided $Z_s(\omega)$ is known as a function of frequency and provided that at all frequencies of interest the inequality (5) is satisfied.

* Note that here $k_y = 0$ and consequently $k_x = h$.

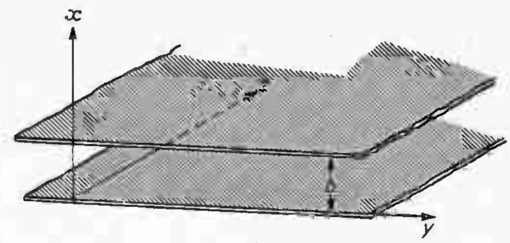


Fig. 1. *Planar waveguide*

If, as in practical cases, the bottom as well as the top guide surface is imperfect then evidently the quantity δh has to be doubled.

To take an important practical case, suppose that the guide is made of smooth metal plates (conductivity $= \sigma$ and thickness d) then, if the waveguide is operated at frequencies at which the skin depth is much smaller than the thickness of the metal plates (which is true in all practical cases) then it can be shown that (see Appendix)

$$Z_s = R_s + j\,X_s = Z_m/Z_0 \qquad \text{(10)}$$

where

$$Z_m = \sqrt{\omega\,\mu_m/\sigma}\;\underline{/45^\circ} \qquad \text{(11)}$$

Evidently at frequencies remote from cut-off (either above or below) we may approximate to (9b) a stage further, namely

$$\gamma = \alpha + j\beta = \gamma_0 + \frac{1}{a}\frac{k_0}{\gamma_0}(j\, Z_s) \qquad \text{(12)}$$

Above cut-off this reduces to

$$\left.\begin{aligned} \beta &= \beta_0 + \frac{1}{a}\frac{k_0}{\beta_0} X_s \\ & \qquad\qquad (\omega > \omega_c) \\ \alpha &= \frac{1}{a}\frac{k_0}{\beta_0} R_s \end{aligned}\right\} \qquad \text{(13)}$$

which for metallic waveguides leads [from Equ. (11) and (13)] to

$$\alpha = \frac{1}{a}\frac{k_0}{\beta_0}\sqrt{\omega\epsilon/2\sigma} \qquad \text{(14)}$$

Below cut-off (12) gives

$$\left.\begin{aligned} \beta &= \frac{1}{a}\frac{k_0}{\alpha_0} R_s \\ & \qquad\qquad (\omega < \omega_c) \\ \alpha &= \alpha_0 - \frac{1}{a}\frac{k_0}{\alpha_0} X_s \end{aligned}\right\} \qquad \text{(15)}$$

In the vicinity of the cut-off frequency the accurate expression (9b) must be used, which at the cut-off frequency gives

$$\gamma_c = \left[\frac{2}{a} k_0 j\, Z_s\right]^{\frac{1}{2}} \quad (\omega = \omega_c) \qquad \text{(16)}$$

This, of course, is finite.

The above formulae are accurate and can be used without discrimination, whatever the physical nature of the surface, as long as the inequality (5) is not violated.

If the waveguide of Fig. 1 supports an $H_{0m\bullet}$ mode then the quantity h can be shown to be given by[10]

$$j\frac{Z_s}{k_0} = \frac{\tan ha}{h} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (17)^*$$

Since $h = h_0 + \delta h$ then provided that

$$|Z_s|/k_0 \ll 1 \quad .. \quad .. \quad .. \quad .. \quad (18)$$

we may approximate as in connection with E-waves and we get

$$\delta h = \frac{1}{a}\frac{h_0}{k_0}(j\,Z_s) \quad .. \quad .. \quad .. \quad .. \quad (19)$$

Consequently we get

$$\left.\begin{aligned}\gamma = \alpha + j\beta &= \left\{(h_0 + \delta h)^2 - k_0^2\right\}^{\frac{1}{2}} \\ &\approx \left\{h_0^2 - k_0^2 + \frac{2}{a}\left(\frac{h_0}{k_0}\right)^2 k_0\, j\,Z_s\right\}^{\frac{1}{2}} \\ &= \left\{\gamma_0^2 + \frac{2}{a}\left(\frac{h_0}{k_0}\right)^2 k_0\, j\,Z_s\right\}^{\frac{1}{2}}\end{aligned}\right| \quad (20)$$

These formulae are valid provided the inequality (5) and the additional restriction (18) are satisfied. For sufficiently small values of $\omega\left(=\frac{k_0}{c}\right)$ the expressions (20) diverge because of the violation of the condition (18). Thus at very low frequencies the transcendental equation (17) must be used to evaluate h and therefore γ is given by

$$\gamma = \left\{h^2 - k_0^2\right\}^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad (21)$$

In practical applications, however, this procedure is unnecessary since, the inequality (18) and consequently formulae (20) hold throughout the frequency spectrum right down to below a fraction of 1 Mc/s.

The Circular Waveguide

In the case of circular waveguides (radius s) operated in any $E_{mn\bullet}$ mode the quantity δh can be shown to be

$$\delta h = \frac{1}{s}\frac{k_0}{h_0}(j\,Z_s) \quad .. \quad .. \quad .. \quad .. \quad (22)$$

A comparison of Equs. (8) and (22) shows that a circular waveguide has the same axial propagation coefficients as a planar waveguide having one imperfect wall and whose separation between the plates equals the radius of the circular waveguide. Hence Equ. (9) et seq. can be used to give corresponding quantities for circular waveguides provided the quantity a is replaced by s.

If the circular waveguide is excited in any H_{0n} mode then the quantity δh is given by[10]

$$\delta h = \frac{1}{s}\frac{h_0}{k_0}(j\,Z_s) \quad .. \quad .. \quad .. \quad .. \quad (23)$$

provided that the inequality (18) is satisfied. In this case Equ. (20) can be used to evaluate the phase propagation and attenuation coefficients. At very low frequencies (frequencies much smaller than 1 Mc/s) the inequality (18) will not be met and the cut-off coefficient, h, must be calculated from the following transcendental equation[10]

$$j\frac{Z_s}{k_0} = \frac{1}{h}\frac{J_1(hs)}{J_0(hs)} \quad .. \quad .. \quad .. \quad .. \quad (24)$$

* Note that in reference [10] there is an approximation in the equation following Equ. (23); viz., $\sin k_x x$ should read $\tan k_x x$.

The quantity γ is then obtained from Equ. (21). However, this is a situation that is unlikely to arise in any practical application.

In the case of higher order H-modes ($m \neq 0$), provided the inequality (18) is satisfied, the quantity δh is given by[10]

$$\delta h = \frac{1}{s}\left\{\frac{h_0}{k_0}\right\}\left\{1 + \left(\frac{\beta_0}{h_0}\right)^2\left(\frac{m}{h_0 s}\right)^2\right\} \left\{1 - \left(\frac{m}{h_0 s}\right)^2\right\}^{-1}(j\,Z_s) \quad (25)$$

The substitution of Equ. (25) into (9) gives

$$\gamma = \{\gamma_0^2 + \Delta\}^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (26)$$

where

$$\begin{aligned}\Delta &= 2k_0\left[\frac{1}{s}\left(\frac{h_0}{k_0}\right)^2 + \frac{1}{s}\left(\frac{m}{h_0 s}\right)^2\left\{1 - \left(\frac{m}{h_0 s}\right)^2\right\}^{-1}\right](j\,Z_s) \\ &= 2\,M_c\,(j\,Z_s) \quad .. \quad .. \quad (27)\end{aligned}$$

This expression is, again, accurate throughout the frequency spectrum provided that the inequalities (5) and (18) are satisfied.

Remote from the cut-off frequency we have

$$\gamma = \gamma_0 + \frac{M_c}{\gamma_0}(jZ_s) \quad .. \quad .. \quad .. \quad .. \quad (28)$$

so that above the cut-off we have

$$\left.\begin{aligned}\beta &= \beta_0 + \frac{M_c}{\beta_0}X_s \\ \alpha &= \frac{M_c}{\beta_0}R_s\end{aligned}\right\} (\omega > \omega_c) \quad .. \quad .. \quad (29)$$

and below cut-off,

$$\left.\begin{aligned}\beta &= \frac{M_c}{\alpha_0}R_s \\ \alpha &= \alpha_0 - \frac{M_c}{\alpha_0}X_s\end{aligned}\right\} (\omega < \omega_c) \quad .. \quad .. \quad (30)$$

while in the vicinity of the cut-off frequency the exact expression

$$\gamma = \left\{\gamma_0^2 + 2\,M_c\,(j\,Z_s)\right\}^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad (31)$$

must be used; this at the cut-off gives

$$\gamma_c = \left\{2\,M_c\,(j\,Z_s)\right\}^{\frac{1}{2}} \quad (\omega = \omega_c) \quad .. \quad .. \quad (32)$$

Rectangular Waveguide

In the case of a rectangular waveguide operated in the H_{0n}-mode the cut-off coefficient, h $(=\sqrt{k_x^2+k_y^2})$ is again obtained as a sum of $h_0 + \delta h$ (provided, of course, that inequalities (5) and (18) are satisfied). The quantity δh can be shown to be given by [10]

$$\delta h = \left\{\frac{1}{b}\frac{k_0}{h_0} + \frac{2}{a}\frac{h_0}{k_0}\right\}(j\,Z_s) \quad .. \quad .. \quad .. \quad (33)$$

and consequently γ is given by Equ. (26) with

$$\Delta = 2\,M_R\,(j\,Z_s) \quad .. \quad .. \quad .. \quad (34)$$

where

$$M_R = k_0\left\{\frac{2}{a}\left(\frac{h_0}{k_0}\right)^2 + \frac{1}{b}\right\} \quad .. \quad .. \quad .. \quad (35)$$

2. Numerical Examples

General

In the numerical examples, given below, we will concentrate our attention on plain metallic waveguides. It will be assumed that the inequalities (5) and (18) are satisfied and, in addition, that the thickness of the waveguide wall is much greater than the skin depth in the metal of which the waveguide is made.

It will be convenient to plot the curves of α and β against k_0, rather than ω and in this connection we note that

$$(j\,Z_s) = \sqrt{\frac{\omega\,\mu_m}{\sigma}}\,\sqrt{\frac{\epsilon_0}{\mu_0}}\,\underline{/\frac{3\pi}{4}} \quad \ldots \quad \ldots \quad (36)$$

If $\mu_m = \mu_0$ (non-magnetic conductors) then

$$(j\,Z_s) = k^{\frac{1}{2}}\,c^{\frac{1}{2}}\,\sqrt{\epsilon_0/\sigma}\,\underline{/\frac{3\pi}{4}}$$

$$= k^{\frac{1}{2}}\,Z\,\underline{/\frac{3\pi}{4}} \quad \ldots \quad \ldots \quad \ldots \quad (37)$$

where $Z = \sqrt{c\epsilon/\sigma}$

For a copper conductor the quantity Z is $6{\cdot}8 \times 10^{-6}$.

E-Wave in a Circular Waveguide

According to Section 1 using the substitution (37), the coefficient γ is given for any E-wave by

$$\gamma = \alpha + j\beta = \left\{h_0^2 - k_0^2 + 2k^{3/2}\left(\frac{Z}{s}\right)\underline{/\frac{3\pi}{4}}\right\}^{\frac{1}{2}} \quad \ldots \quad (38)$$

$$= \left\{h_0^2 - k_0^2 + 2hk^{3/2}\left(\frac{Z}{\nu_{mn}}\right)\underline{/\frac{3\pi}{4}}\right\}^{\frac{1}{2}} \quad \ldots \quad (38a)$$

The curves shown in Fig. 2 are the plot of the real

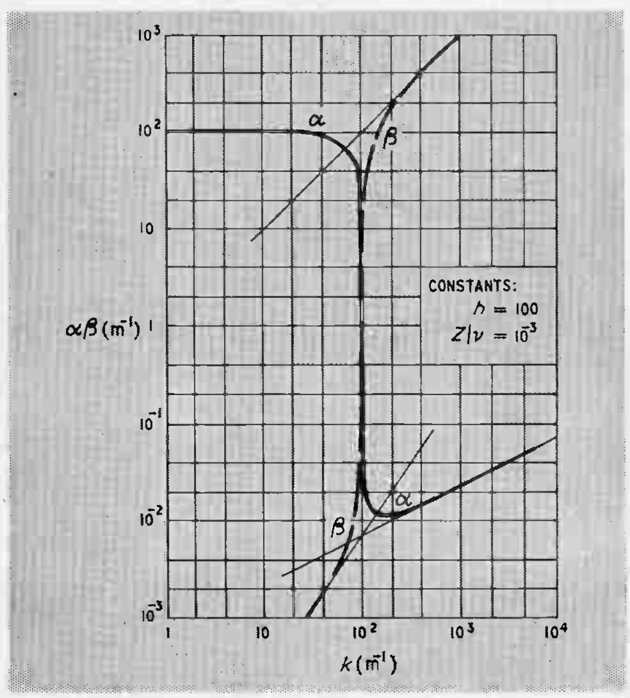


Fig. 2. *α and β coefficients of a circular waveguide carrying an E-mode*

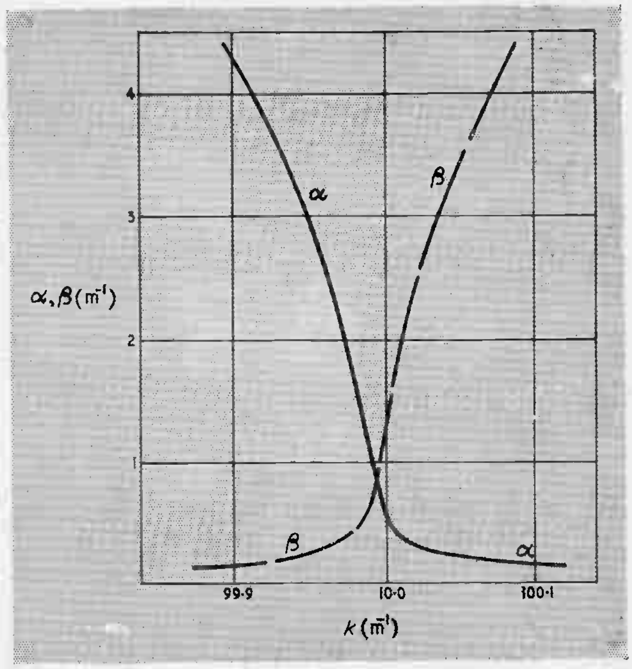


Fig. 3. *α and β coefficients in the vicinity of cut-off*

and imaginary parts of Equation (38) for $h = 100m^{-1}$ and $Z/s = 10^{-3}$ (i.e., $Z/\nu_{mn} = 10^{-5}$, since $hs = \nu_{mn}$ = the root of the appropriate Bessel function).

In the vicinity of the cut-off frequency (where $h^2 - k_0^2$ is a small number) to find α and β the exact expression (38) or (38a) must be used. At the cut-off this gives

$$\alpha = \left(\frac{2}{s}\right)^{\frac{1}{2}} k^{\frac{3}{4}}\, c^{\frac{1}{4}} \left\{\frac{\epsilon_0}{\sigma}\right\}^{\frac{1}{4}} \cos\frac{3\pi}{8} \quad \ldots \quad \ldots \quad \ldots \quad (39)$$

while β is given by the same expression with $\sin\frac{3\pi}{8}$ in place of $\cos\frac{3\pi}{8}$.

Remote from the cut-off frequency we may approximate to Equ. (38), viz.,

$$\gamma = \left\{h_0^2 - k_0^2\right\}^{\frac{1}{2}} + k^{3/2}\,(h_0^2 - k_0^2)^{-\frac{1}{2}}\,\frac{Z}{s}\,\underline{/\frac{3\pi}{4}} \quad \ldots \quad (40)$$

so that above cut-off

$$\beta \approx \sqrt{k_0^2 - h_0^2}$$

$$\alpha \approx \frac{k^{3/2}}{\sqrt{2}\,\sqrt{k_0^2 - h_0^2}}\,\frac{Z}{s} \quad \ldots \quad \ldots \quad \ldots \quad (41)$$

which are the well-known expressions.

However, below cut-off we have

$$\beta \approx \frac{k^{3/2}}{\sqrt{2}\,\sqrt{h_0^2 - k_0}}\,\frac{Z}{s}$$

$$\alpha \approx \sqrt{h_0^2 - k_0^2} \quad \ldots \quad \ldots \quad \ldots \quad (42)$$

The behaviour of α and β curves in the vicinity of the cut-off is shown on an enlarged scale in Fig. 3. It is to be noted that at the cut-off frequency β is larger than α by approximately a factor of two and α becomes equal to β at a frequency slightly below the cut-off. The general nature of the α and β curves when plotted on a logarithmic paper (Fig. 3) is peculiar to E-waves.

H_{on}-Wave in a Circular Waveguide

In the case of H_{0n}-waves, as explained in Section 1, provided the inequalities (5) and (18) are satisfied the quantity γ is given by (20), which with the substitution (37) gives

$$\gamma = \left\{ h_0^2 - k_0^2 + 2\, h_0^2\, k_0^{-\frac{1}{2}} \left(\frac{Z}{s}\right) \Big/ \frac{3\pi}{4} \right\}^{\frac{1}{2}} \quad .. \quad (43)$$

In the vicinity of the cut-off frequency the exact expression (43) must be used, and at the cut-off α and β assume the same values as in the case of E-waves [Equ. (39)].

Remote from the cut-off frequency we may approximate to Equ. (43), namely

$$\gamma = (h_0^2 - k_0^2)^{\frac{1}{2}} + \frac{h^2}{k_0^{\frac{1}{2}}\,(h_0^2 - k_0^2)^{\frac{1}{2}}} \left(\frac{Z}{s}\right) \Big/ \frac{3\pi}{4} \quad (44)$$

so that above cut-off we get the conventional formulae

$$\beta \approx \sqrt{k_0^2 - h_0^2}$$

$$\alpha \approx \frac{h^2}{\{2k\,(k_0^2 - h_0^2)\}^{\frac{1}{2}}} \left(\frac{Z}{s}\right) \quad .. \quad .. \quad .. \quad (45)$$

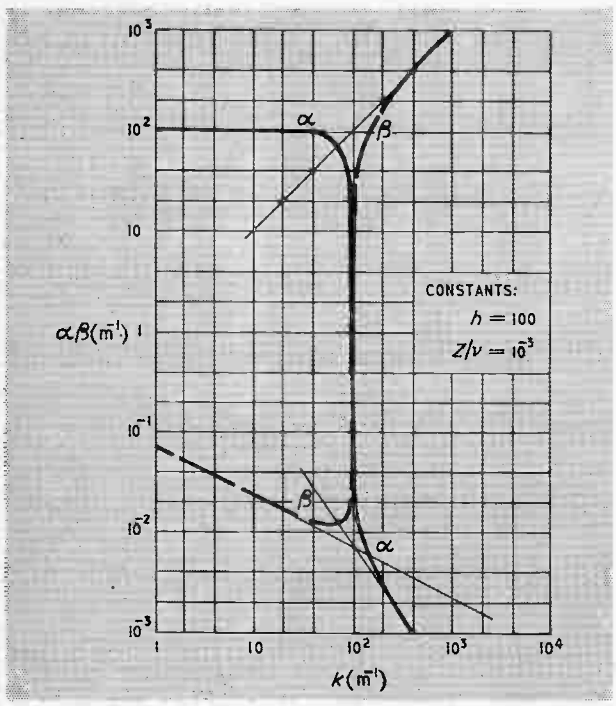
But below cut-off we have

$$\beta \approx \frac{h^2}{\{2k\,(h_0^2 - k_0^2)\}^{\frac{1}{2}}} \left(\frac{Z}{s}\right)$$

$$\alpha \approx \sqrt{h_0^2 - k_0^2} \quad .. \quad .. \quad .. \quad .. \quad (46)$$

The coefficients α and β are shown plotted in Fig. 4 as a function of k_0 for a waveguide transmitting an H_{0n}-mode and having $h = 100 m^{-1}$ and $Z/s = 10^{-3}$, as for E-waves. The behaviour of the α and β curves in the immediate vicinity of the cut-off is similar to that for an E-wave (Fig. 3); but the general nature of the α and β curves when plotted on a logarithmic paper is peculiar to the H_0-waves in a circular waveguide.

Fig. 4. *α and β coefficients of a circular waveguide carrying an H_0-mode*



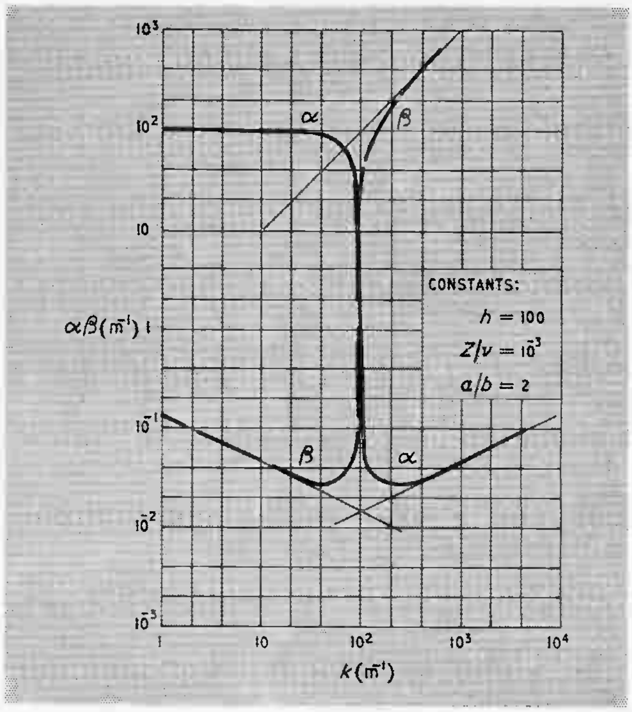


Fig. 5. *α and β coefficients of a rectangular waveguide carrying an H_0-mode*

An H_{0n}-Wave in a Rectangular Waveguide

In the case of a rectangular waveguide the waves of interest are the H_{0n}-modes for which we have (see Sec. 1).

$$\gamma = \left\{ h_0^2 - k_0^2 + 2\, M_R\, k^{\frac{1}{2}}\, Z \Big/ \frac{3\pi}{4} \right\}^{\frac{1}{2}} \quad .. \quad (47)$$

Substituting for M_R we get

$$\gamma = \left\{ h_0^2 - k_0^2 + 2\, Z \frac{k^{3/2}}{b} + \frac{2}{a}\, h_0^2\, k^{-\frac{1}{2}} \Big/ \frac{3\pi}{4} \right\}^{\frac{1}{2}} \quad (48)$$

The real (α) and imaginary (β) parts of γ are shown plotted as a function of k_0 in Fig. 5, for a waveguide whose $a/b = 2$, $h = 100$ and $Z/\nu_{mn} = 10^{-5}$, using Equ. (48) The nature of α and β curves when plotted against k_0 (Fig. 5) on logarithmic paper is peculiar to the H_{on}-modes in rectangular waveguides, but it is a feature of all H-modes that below the cut-off there is a region of negative velocity.

3. The Nature of α and β Curves

Restricted Frequency Range

The shape of the α and β curves is completely determined by, (i) the waveguide geometry and the mode type guided within it, (ii) the functional dependence of Z_s on frequency. In general, therefore, no two sets of α and β curves can be the same.

Notwithstanding, if the functional dependence of Z_s

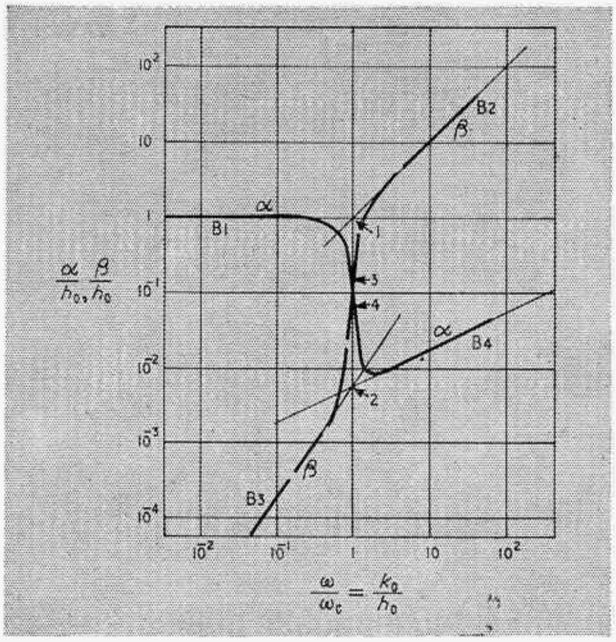


Fig. 6. *Normalized α and β curves for metallic circular waveguides operated in any E-mode*

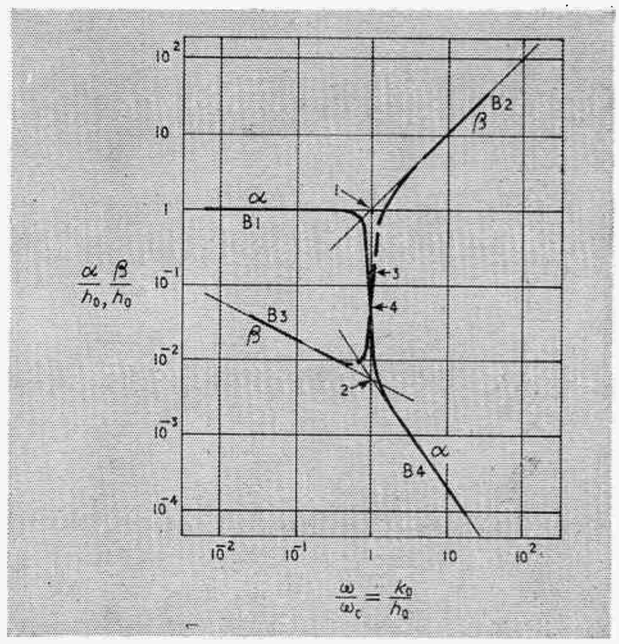


Fig. 7. *Normalized α and β curves for metallic circular waveguides operated in any H_0-mode*

on frequency is a simple one (as for example in the case of a smooth metal surface $Z_s \propto \sqrt{\omega}$) then, though the curves differ in detail, all E-modes (parallel plates or circular waveguides) have α and β curves of the same general shape (Fig. 6). In a similar manner it can be shown that all H_0-modes (whether guided by a parallel plate waveguide or a circular waveguide) have essentially similar α and β curves (Fig. 7). The third class of α and β curves is the largest one; it contains all higher order H_{mn}-modes ($m \neq 0$) in circular waveguides and all H-modes in rectangular waveguides (Fig. 8).

It will be observed that if α, β and k_0 are normalized with respect to the cut-off coefficient h_0 then in all cases the branches B.1 and B.2 intersect at a point (1, 1) and have slopes 0 and 1 respectively. The location of branches B.3 and B.4, however, depends on the mode carried in the waveguide as well as on the nature of the $Z_s(\omega)$ as a function of frequency. Yet, there are only three distinct sets of α and β curves. In particular, in the case of plain metal waveguides $Z_s \propto \omega^{\frac{1}{2}}$ and, therefore, branch 3 has a slope of 3/2 in the case of E-waves and $-\frac{1}{2}$ in all other cases; branch 4 on the other hand has a slope $-3/2$ in the case of H_0-waves supported in planar waveguides or circular waveguides and has a slope of $\frac{1}{2}$ in all other cases. If Z_s exhibits some other power dependence on ω then these slopes will change accordingly. In all cases the point of intersection of the branches B.3 and B.4 has an ordinate (normalized with respect to h_0) given by

$$\frac{\chi}{\nu_{mn}} R_s(h_0) \qquad \qquad (49)$$

where ν_{mn} is the characteristic number of the mode concerned, $R_s(h_0)$ is the value of the surface resistance at the cut-off and χ is a constant which is: (i) unity for all E-modes and H_0-modes in cylindrical and planar waveguides; (ii) the aspect ratio (a/b), in the case of rectangular waveguides; (iii) the quantity $[(\nu_{mn}/m)^2-1]^{-1}$ in the case of H_{mn} waves ($m \neq 0$) in circular waveguides.

The normalized values of α and β coefficients at the cut-off (points 3 and 4 respectively) are given by

$$\alpha_c = \left|\frac{2\eta}{\nu_{mn}} Z_s(h_0)\right|^{\frac{1}{2}} \cos(45° + \tfrac{1}{2}\phi_s)$$

$$\beta_c = \left|\frac{2\eta}{\nu_{mn}} Z_s(h_0)\right|^{\frac{1}{2}} \sin(45° + \tfrac{1}{2}\phi_s) \qquad (50)$$

where $Z_s(h_0) = |Z_s(h_0)| \angle\phi_s$ is the value of the surface impedance at the cut-off and η is a constant which is: (i) unity for all E-waves and H_0-waves in cylindrical and planar waveguides (Figs. 6 and 7); (ii) $2 + a/b$ for rectangular waveguides (Fig. 8); (iii) $[1-(m/\nu_{mn})^2]^{-1}$ for circular waveguides supporting any H_{mn} wave ($m \neq 0$) (Fig. 8).

Finally, we note that all H-modes are distinct from E-modes in that they exhibit a group velocity-characteristic, which is negative (slope of the β characteristic is negative) below a certain frequency in the stop-band.

Extreme Values of Frequency

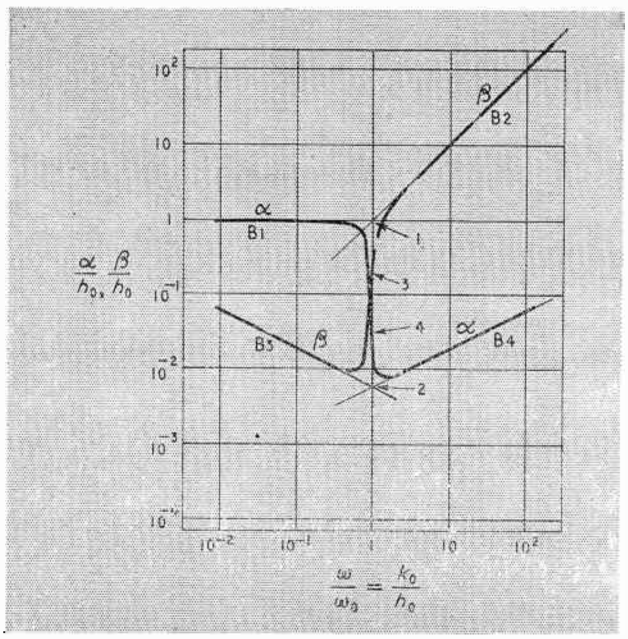
The theory given above is certainly accurate provided the inequalities (5) and (18) are satisfied. This undoubtedly is the case with commercially available metal waveguides over a frequency band at least one thousand times smaller than the cut-off to one thousand times higher than the cut-off.

However, the above theory cannot be applied indiscrimately at extremely high values of frequency, because Z_s is, in general, an increasing function of frequency so that for sufficiently large values of frequency the inequality (5) will no longer be satisfied. Thus, with plain metallic waveguides ($Z_s \propto \sqrt{\omega}$) at frequencies of the order of 10^8 Mc/s (frequencies corresponding to visible light) the inequality (5) fails to be satisfied. For practical applications, however, the range of validity of the formulae developed may be taken as infinite, as far as the high frequency end of the spectrum is concerned.

It is appropriate at this place to stress that before these extremely high frequencies are reached the shape of α and β curves will undergo a considerable change for yet another and more practical reason. Consider the simplest practical case: a metallic waveguide. Its surface impedance is given by Equ. (10) but with the proviso that the metal surface is homogeneous and perfectly smooth. However, commercially available waveguides have tolerances seldom better than about 0·001″ and even best ground finished metal surfaces have irregularities[12] whose depth is about 20 μ-inch with spacings of the order of 0·001″. Surface irregularities of that nature, though negligible in the centrimetric band and even sometimes in the millimetric band, become of primary importance in the submillimetric band. In fact at these frequencies the upper end of the β-characteristic (Branch B.2) will become broken up into a succession of pass- and stop-bands and then, of course, it becomes impossible to predict the behaviour of the waveguide except on a statistical basis. The matter is further complicated by the fact that at frequencies very much higher than the cut-off (say, $\omega = 100\ \omega_c$) the waveguide, in general, will be capable of supporting a multitude of different modes (say, in excess of 100): the consequential mode conversion losses are appreciable so that α and β characteristics are substantially modified.

In the low frequency end of the spectrum with all H-waves attention must be turned to the approximation (18). For commer ially available metallic waveguides this inequality is satisfied for frequencies even smaller

Fig. 8. *Normalized α and β curves for metallic waveguides either circular and operated in any H_{mn}-mode ($m \neq 0$) or rectangular and operated in any H_0-mode. The slopes of the four branches are: B.1, 0; B.2, 1; B.3, $-\frac{1}{2}$; B.4, $\frac{1}{2}$*



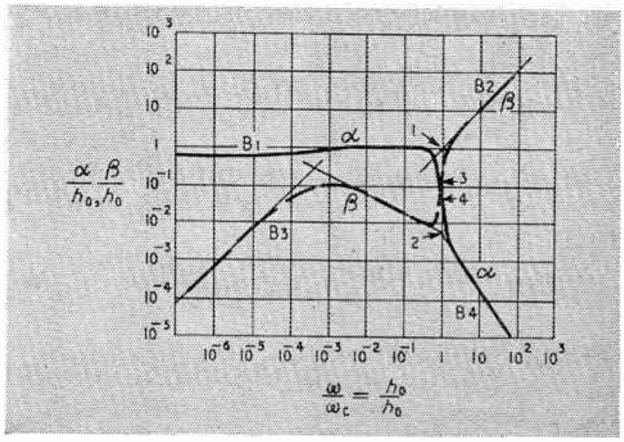


Fig. 9. *Normalized α and β curves for metallic circular waveguides operated in any H_0-mode, showing region much below cut-off*

than 1/100th of the cut-off but, at frequencies smaller than the cut-off by about 1,000, Equ. (19) becomes a poor approximation and for frequencies smaller than this the results cannot be relied upon. At these very low frequencies it is necessary in order to determine the value of the coefficient h to solve an appropriate transcendental equation without the approximation (18). Thus with rectangular waveguides Equ. (17) has to be solved while with circular waveguides Equ. (24) must be solved. In the later case it will be observed that as the frequency varies the coefficient h changes from

$$h = h_0 + \delta h \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (51)$$

at higher frequencies (say $\omega > \omega_c \times 10^{-3}$) to

$$h = h'_0 + \delta h' \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (52)$$

at lower frequencies (say $\omega < \omega_c \times 10^{-6}$) where h_0s is the nth root of J_1 and h'_0s is the nth root of the J_0 function and

$$\left.\begin{aligned} \delta h &= \frac{1}{s}\frac{h_0}{k_0}(jZ_s) \\ \text{and} \\ \delta h' &= -\frac{1}{s}\frac{k_0}{h_0}\frac{1}{jZ_s} \end{aligned}\right| \qquad \ldots \quad \ldots \quad (53)$$

At these extremely low values of frequency evidently

$$\gamma \propto h = h_0' + \delta h' \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (54)$$

Thus

$$\left.\begin{aligned} \alpha &= h_0' + \frac{1}{s}\frac{k_0}{h_0'}\frac{X_s}{|Z_s|^2} \\ \text{and} \\ \beta &= \frac{1}{s}\frac{k_0}{h_0'}\frac{R_s}{|Z_s|^2} \end{aligned}\right| \qquad \ldots \quad \ldots \quad (55)$$

With plain metallic waveguides, at these extremely low frequencies, the expression (11) will no longer be valid since the skin depth (see appendix) becomes comparable with the thickness of the waveguide wall. Under such circumstances it is necessary to use the exact expression for the impedance of a metal wall [see Equ. (61)]. At a sufficiently low frequency, however, the low-frequency approximation [see Equ. (64)]

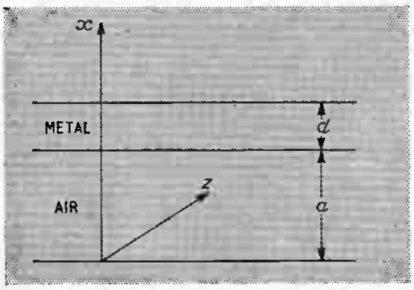


Fig. 10. *Parallel plate metal waveguide*

$$\left.\begin{aligned} R_s &= \frac{1}{\sigma d\ Z_0} \\ X_s &= \frac{1}{3}\, k_0\, d \end{aligned}\right\} \qquad \ldots \qquad \ldots \qquad (56)$$

may be used.

Thus, while at high frequencies R_s, X_s, $\propto \sqrt{\omega}$, at very low frequencies R_s becomes a constant while X_s becomes proportional to ω and hence tends to zero. Evidently for sufficiently small values of frequency α changes from its stationary value h_0 to h'_0 ($h'_0 < h_0$) while β changes its functional dependence from $k^{-\frac{1}{2}}$ to k^1, as indicated in Fig. 9.

With E-waves the analysis given in Sec. 1 holds at any frequency, however low. Consequently, in the case of metallic waveguides, to obtain expressions valid at very low frequencies the values of R_s and X_s given by Equ. (26) can be substituted therein, giving

$$\begin{aligned} \alpha &= h_0 \\ \beta &= \frac{1}{s}\,\frac{k_0}{h_0}\left\{\frac{1}{\sigma d\ Z_0}\right\} \qquad \ldots \qquad \ldots \qquad \ldots \qquad \ldots \qquad (57) \end{aligned}$$

Thus α remains unchanged but β changes its functional dependence from $k^{\frac{1}{2}}$ to k^1.

TABLE 1

The γ-coefficient of a cylindrical waveguide: $\gamma = \alpha + j\beta = \{h^2 - k_0^2\}^{\frac{1}{2}}$

Waveguide geometry	Mode	Coefficient h	Explanatory remarks
Planar *	E_{0n}	$j\, Z_s \cdot k_0 = h \tan ha$	a = separation between the guiding planes
	H_{0n}	$j\, \dfrac{Z_s}{k_0} = \dfrac{\tan ha}{h}$	
Circular	E_{mn}	$j\, Z_s \cdot k_0 = h J_0(hs)/J_0'(hs)$	s = radius of the waveguide
	H_{0n}	$j\, \dfrac{Z_s}{k_0} = \dfrac{J_0'(hs)/J_0(hs)}{h}$	
	H_{mn} ($m \neq 0$)	given by the solution of two simultaneous equations. (72) and (73) of Ref. (10).	
Rectangular **	H_{0n}	given by the solution of two simultaneous equations (see Ref. 10).	a = broad dimension of the waveguide.

* For one imperfect plane only.
** The small dimension, b, of the waveguide is assumed to be large in comparison with the skin depth.

TABLE 2

The simplified γ-coefficient of a cylindrical waveguide.
$\gamma = [\gamma_0^2 + 2\, M(j\, Z_s)]^{\frac{1}{2}}$

Waveguide geometry	Mode	Quantity M
Planar	E_{0n}	$\dfrac{k_0}{a} = \dfrac{2\pi}{\lambda_0 a}$
	H_{0n}	$\dfrac{k_0}{a}\left\{\dfrac{h_0}{k_0}\right\}^2 = \dfrac{2\pi}{\lambda_0 a}\left\{\dfrac{\lambda_0}{\lambda_c}\right\}^2$
	E_{mn}	$\dfrac{k_0}{s} = \dfrac{2\pi}{\lambda_0 s}$
Circular	H_{0n}	$\dfrac{k_0}{s}\left\{\dfrac{h_0}{k_0}\right\}^2 = \dfrac{2\pi}{\lambda_0 s}\left\{\dfrac{\lambda_0}{\lambda_c}\right\}^2$
	H_{mn} ($m \neq 0$)	$\dfrac{k_0}{s}\left\{\dfrac{h_0}{k_0}\right\}^2 + \dfrac{k_0}{s}\left\{\left\{\dfrac{h_0 s}{m}\right\}^2 - 1\right\}^{-1}$ $= \dfrac{2\pi}{\lambda_0 s}\left\{\dfrac{\lambda_0}{\lambda_c}\right\}^2 + \dfrac{2\pi}{\lambda_0 s}\left\{\left\{\dfrac{\nu_{mn}}{m}\right\}^2 - 1\right\}^{-1}$
Rectangular	H_{0n}	$2\,\dfrac{k_0}{a}\left\{\dfrac{h_0}{k_0}\right\}^2 + \dfrac{k_0}{b} = \dfrac{4\pi}{\lambda_0 a}\left\{\dfrac{\lambda_0}{\lambda_c}\right\}^2 + \dfrac{2\pi}{\lambda_0 b}$

ν_{mn} is the nth root of $J_m'(x) = 0$
At frequencies remote from cut-off,
$\gamma = \gamma_0 + M\,(j\, Z_s)$
At the cut-off frequency,
$\gamma = [2M\,(j\, Z_s)]^{\frac{1}{2}}$

4. Summary of Results

It is assumed that the waveguide is homogeneous and has a surface impedance Z_s. The wave propagation coefficient γ is given by

$$\gamma = \alpha + j\beta = \left\{h^2 - k_0^2\right\}^{\frac{1}{2}} \qquad \ldots \qquad \ldots \qquad \ldots \qquad (58)$$

where k_0 is the free space propagation coefficient and h is the cut-off coefficient.

In general, the quantity h can be evaluated as shown in Table 1. The coefficient γ then follows from Equ. (58) and, at least in principle, the calculation can be carried out at any frequency for any surface impedance.

In the frequency range from about 1,000 times less to 1,000 times larger than the cut-off a much simpler procedure can be adopted provided that throughout this range Z_s remains a small quantity (say $|Z_s| < 0{\cdot}01$). Under such conditions γ is given by the very much simpler expression

$$\left.\begin{aligned} \gamma &= \left\{h_0^2 - k_0^2 + 2\, M\,(j\, Z_s)\right\}^{\frac{1}{2}} \\ &= \left\{\gamma_0^2 + 2\, M\,(j\, Z_s)\right\}^{\frac{1}{2}} \end{aligned}\right\} \qquad \ldots \qquad \ldots \qquad (59)$$

as shown in Table 2.

Except in the vicinity of the cut-off, the expression (59) may be approximated by

$$\gamma = \gamma_0 + \frac{M}{\gamma_0}(j\, Z_s) \qquad \ldots \qquad \ldots \qquad \ldots \qquad (60)$$
$$(\gamma_0 \neq 0)$$

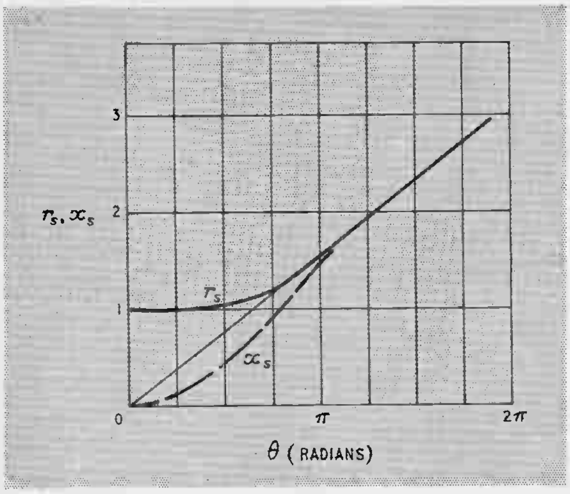


Fig. 11. *The real* (R_s) *and imaginary* (X_s) *components of the surface impedance of a metal crust, thickness d;*

$$Z_s = R_s + jX_s = \frac{1}{Z_0\,\sigma d}(r_s = jX_s);\ \theta = d\sqrt{2\omega\mu_m\sigma}$$

With plain metallic waveguides Z_s has the simple form

$$Z_s = \sqrt{\frac{\omega\mu m}{\sigma}}\sqrt{\frac{\epsilon_0}{\mu_0}}\frac{/\pi}{/4}$$

and this value of Z_s may be used in the formulae given in Tables 1 and 2 to calculate the α and β coefficients.

If, in any practical application, the waveguide walls are made up of a more complicated medium or are of a more complicated structure, then evidently a different expression for the surface impedance must be used[10]. Nevertheless the formulae given in Tables 1 and 2 apply so long as $|Z_s| \ll 1$.

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APPENDIX

The Surface Impedance of a Metal Crust of Finite Thickness

Consider a planar waveguide as shown in Fig. 10. The plane $x = 0$ (a perfectly conducting plane) forms one wall of the waveguide while the plane of $x = a$ forms the other waveguide wall. The space between the plane $x = a$ and $x = a + d$ is filled with a homogeneous metal of conductivity σ and permittivity μ_m.

If this structure is now made to guide an electromagnetic wave then provided the thickness, d, is not unreasonably small the surface impedance Z_s of the wall at $x = a$ will be small, while that of the plane at $x = a + d$ will be large and may be taken to be infinite. The latter is a simplifying assumption which is not exactly true, since it can be shown that part of the guided wave leaks through the metal crust to the outside of the waveguide and is carried there in the form of a surface wave. The amount of the energy actually carried in the surface wave is, however, under all practical cases vanishingly small and may therefore be neglected.

Assuming, therefore, that the surface impedance of the wall at $x = a + d$ is infinite it is easy to show that the surface impedance of the wall at $x = a$ is given by

$$Z_s = \frac{Z_m}{Z_0}\coth p\,d \quad \ldots \quad (61)$$

where Z_m is given by (11) and

$$p = \sqrt{j\omega\,\mu_m\sigma} \quad \ldots \quad (62)$$

For all good conductors σ is large and, therefore, at microwaves p is a very large number so that, even for very small values of d, the quantity $\coth p\,d$ becomes equal to unity and we may take

$$Z_s = Z_m/Z_0 \quad \ldots \quad (63)$$

However, at low frequencies (for practical waveguides much less than 1 Mc/s) the skin depth becomes comparable with the thickness of the waveguide wall, d, and the exact expression (61) must be used. For this case the normalized real and imaginary parts of Z_s are shown plotted in Fig. 11. It will be observed that provided the thickness of the metal, d, is more than about three times the skin depth, the real and imaginary parts of Z_s become equal and the approximation (63) is valid. Otherwise the exact expression (61) must be used.

At very low frequencies (for ordinary waveguides of the order of a few kc/s or less), the thickness of the metal, d, becomes small in comparison with the skin depth and the reactive part of Z_s disappears giving

$$\left.\begin{aligned} Z_s &= \frac{1}{\sigma d\,Z_0} + j\frac{1}{3}k_0\,d \\ R_s &= \frac{1}{\sigma d\,Z_0} \\ X_s &= \frac{1}{3}k_0 d \to 0 \end{aligned}\right\} \quad \ldots \quad (64)$$

LIST OF SYMBOLS

$k_0 = \frac{2\pi}{\lambda_0} = \frac{\omega}{c}$	= free-space propagation coefficient.
λ_0	= free-space wavelength.
h_0	= cut-off coefficient of a perfect waveguide.
$\gamma_0 = j\,\beta_0$	= propagation coefficient of a perfect waveguide.
$h = h_0 + \delta h$	= cut-off coefficient of an imperfect waveguide.
$\gamma = \alpha + j\beta$	= propagation coefficient of an imperfect waveguide.
α	= attenuation coefficient.
$\beta = \beta_0 + \delta\beta = \frac{2\pi}{\lambda_g}$	= phase propagation coefficient.
$Z_s = \|Z_s\| \angle \phi_s$	= surface impedance (normalized with respect to Z_0).
$Z_m = \sqrt{\omega\,\mu_m/\sigma}\ \angle 45°$	= intrinsic impedance of a metal conductivity σ and permeability μ_m (assumed real, otherwise the phase angle of Z_s will be altered accordingly).
a, b	= dimensions of a rectangular waveguide.
s	= radius of a circular waveguide.
d	= thickness of the waveguide wall.
ν_{mn}	= characteristic number of a wave mode.
$Z = \sqrt{c\epsilon_0/\sigma}$	= a parameter defined in Section 2.
M	= a coefficient introduced in Section 4.
χ, η	= quantities used in Section 3.

Operational Calculus—2: Sine-Wave Inputs

In an earlier article, a number of corresponding pairs of functions in the time world and in the "p-world" of operational calculus were listed, and the general principles of the operational approach to electrical problems were explained and applied in simple cases where the input voltage was a step function. Here, we consider first the effect of applying a sine-wave voltage at zero time to the circuit of Fig. 1 discussed in the earlier article and repeated here for convenience. Suppose then that the input voltage $V(t)$ is $v_0 \sin\omega t$, and that it is applied to the circuit of Fig. 1. The p-world counterpart of this voltage is $v_0\omega p/(p^2 + \omega^2)$, by item 5 of Table 1 of the earlier article and the impedance $Z(p)$ is $R + 1/(pC)$ as before. The current $I(t)$, therefore, has now the p-world counterpart

$$I(p) = \frac{v_0}{Z(p)} \cdot \frac{p\omega}{p^2 + \omega^2} = \frac{\omega v_0}{R} \frac{p^2}{(p + \alpha)(p^2 + \omega^2)} \quad \ldots \quad (1)$$

if we write α for $1/CR$, and the p-world counterpart $V_R(p)$ of the voltage across the resistance is R times this quantity.

If we examine Table 1 of the previous article, we find that this particular item is not included in it. However, item 15 has the same denominator but p in the numerator instead of p^2, and the corresponding time-expression has the value zero when t is zero. Furthermore, in the Table heading it was stated that $dh(t)/dt$ corresponds to $pf(p) - ph(0)$ which, in this case, is simply $pf(p)$. In other words, the required value of $I(t)$ is obtained simply by differentiating the time-expression given in item 15 of the Table. We thus obtain

$$I(t) = \frac{\omega v_0}{R} \frac{d}{dt} \left[\frac{\sin(\omega t - \theta) + e^{-\alpha t} \sin\theta}{\omega(\alpha^2 + \omega^2)^{\frac{1}{2}}} \right] \quad \ldots \quad (2)$$

where $\tan\theta = \omega/\alpha = \omega CR$ $(0 < \theta < \pi/2)$. (3)

Carrying out the differentiation, we get

$$I(t) = \frac{v_0/R}{(\alpha^2 + \omega^2)^{\frac{1}{2}}} [\omega \cos(\omega t - \theta) - \alpha e^{-\alpha t} \sin\theta] \quad (4)$$

The right-hand side of (4) is again zero when $t = 0$ in view of the definition (3) of θ.

Now consider the position when the input voltage is $v_0 \cos\omega t$ instead of $v_0 \sin\omega t$.

Table 1 shows that the p-expression for $\cos\omega t$ is p/ω times that for $\sin\omega t$, so that the expression for $I(p)$ is p/ω times that given by equation (1), that is

$$I(p) = \frac{v_0}{R} \frac{p^3}{(p + \alpha)(p^2 + \omega^2)} \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

If we wish, we can interpret this by differentiating equation (4) but, instead, we shall find the answer in a different way, which is more generally applicable.

In (5), the numerator and denominator of the right-hand side are both cubic in p. If we divide p^3 by $(p + \alpha)(p^2 + \omega^2)$, the quotient is clearly 1, and there will be a remainder $r(p)$ which is quadratic in p and, therefore, of lower degree than $(p + \alpha)(p^2 + \omega^2)$. Thus,

$$I(p) = \frac{v_0}{R} \left[1 + \frac{r(p)}{(p + \alpha)(p^2 + \omega^2)} \right] \quad \ldots \quad (6)$$

Now the denominator $(p + \alpha)(p^2 + \omega^2)$ has one linear factor $(p + \alpha)$ and one quadratic factor $(p^2 + \omega^2)$. We are therefore entitled to assume that the last term of (6) is the sum of two fractions

$$\frac{N_1}{p + \alpha} + \frac{N_2}{p^2 + \omega^2} \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

where N_1 is an expression of degree one less than its denominator $(p + \alpha)$, that is to say, a constant, and N_2 is an expression of degree one less than its denominator $(p^2 + \omega^2)$, that is to say of the first degree in p. Thus we may write

$$\frac{r(p)}{(p + \alpha)(p^2 + \omega^2)} = \frac{N_1}{p + \alpha} + \frac{Ap + B}{p^2 + \omega^2} \quad \ldots \quad (8)$$

where N_1, A and B are constants. This is only legitimate because $r(p)$ is of lower degree than the denominator $(p + \alpha)(p^2 + \omega^2)$ and, for that reason, it was necessary to divide the denominator of (5) into the numerator first. Thus, we are required to find N_1, A and B, so that

$$\frac{p^3}{(p + \alpha)(p^2 + \omega^2)} = 1 + \frac{N_1}{p + \alpha} + \frac{Ap + B}{p^2 + \omega^2} \quad (9)$$

or $$p^3 = (p + \alpha)(p^2 + \omega^2) + N_1(p^2 + \omega^2) + (Ap + B)(p + \alpha) \quad \ldots \quad \ldots \quad (10)$$

We can find the values of N_1, A, B by multiplying out the right-hand side of (10) and equating the coefficients of p^2, p and the constant term to zero. Alternatively, since (10) holds for all values of p, we can give any three values to p, and thus obtain three linear simultaneous equations for N_1, A and B. This last method is usually easier than the method of equating coefficients if we choose the arbitrary values of p carefully. In our

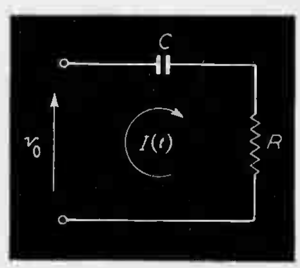


Fig. 1. *Simple RC circuit employed to illustrate the use of the Heaviside transforms*

particular case, the best values to choose are $p = -\alpha$, $p = j\omega$ and $p = -j\omega$. We thus obtain, for $p = -\alpha$,

$$-\alpha^3 = N_1 (\omega^2 + \alpha^2) \quad \ldots \quad (11)$$

and notice that (11) only contains N_1. The two equations derived from $p = j\omega$ and $p = -j\omega$ are, in effect, the same; the first gives

$$-j\omega^3 = (j\omega A + B)(\alpha + j\omega) \quad \ldots \quad (12)$$

Equating real and imaginary parts of (12) gives two equations involving A and B. If $p = -j\omega$, we merely obtain (12) with $-j$ for j. But the neatest way to obtain A and B from (12) is to multiply both sides of (12) by $\alpha - j\omega$, so that

$$-j\omega^3(\alpha - j\omega) = (j\omega A + B)(\alpha^2 + \omega^2) \quad \ldots \quad (13)$$

From (11) and (13) we obtain

$$N_1 = -\frac{\alpha^3}{\alpha^2 + \omega^2}; \; A = -\frac{\omega^2\alpha}{\alpha^2 + \omega^2}; \; B = -\frac{\omega^4}{\alpha^2 + \omega^2} \quad \ldots \quad (14)$$

so that, substituting back into (6)

$$I(p) = \frac{v_0}{R}\left[1 - \frac{\alpha^3}{\alpha^2 + \omega^2} \cdot \frac{1}{p + \alpha} - \frac{\omega^2\alpha p + \omega^4}{(\alpha^2 + \omega^2)(p^2 + \omega^2)}\right] \quad (15)$$

Now 1 corresponds to 1 (item 1 of Table)

$\dfrac{\alpha^3}{\alpha^2 + \omega^2} \cdot \dfrac{1}{p + \alpha}$ corresponds to

$$\frac{\alpha^2}{\alpha^2 + \omega^2} \cdot (1 - e^{-\alpha t}) \quad \text{(item 3)}$$

$\dfrac{\alpha\omega}{\alpha^2 + \omega^2} \cdot \dfrac{p\omega}{p^2 + \omega^2}$ corresponds to $\dfrac{\alpha\omega}{\alpha^2 + \omega^2} \sin \omega t$ .. .. (item 5)

$\dfrac{\omega^2}{\alpha^2 + \omega^2} \cdot \dfrac{\omega^2}{p^2 + \omega^2}$ corresponds to

$$\frac{\omega^2}{\alpha^2 + \omega^2}(1 - \cos\omega t) \quad \ldots \quad \text{(item 7)}$$

so that

$$I(t) = \frac{v_0}{R}\left[1 - \frac{\alpha^2}{\alpha^2 + \omega^2}(1 - e^{-\alpha t}) - \frac{\alpha\omega}{\alpha^2 + \omega^2}\sin \omega t - \frac{\omega^2}{\alpha^2 + \omega^2}(1 - \cos \omega t)\right]$$

$$= \frac{v_0}{R}\left[\frac{\alpha^2}{\alpha^2 + \omega^2} e^{-\alpha t} + \frac{\omega^2}{\alpha^2 + \omega^2}\cos \omega t - \frac{\alpha\omega}{\alpha^2 + \omega^2}\sin \omega t\right] \quad \ldots \quad (16)$$

If θ is the angle defined by (3), then $\sin \theta = \omega/(\alpha^2 + \omega^2)^{\frac{1}{2}}$, and $\cos \theta = \alpha/(\alpha^2 + \omega^2)^{\frac{1}{2}}$; (16) can be simplified to

$$I(t) = \frac{v_0}{R}\left[\frac{\alpha^2}{\alpha^2 + \omega^2} e^{-\alpha t} - \frac{\omega}{(\alpha^2 + \omega^2)^{\frac{1}{2}}}\sin(\omega t - \theta)\right] \quad (17)$$

In this case, it would have been much easier and quicker to have obtained the answer by differentiating (4) but we have worked it out the long way to illustrate the method if (4) had not been available to us.

It should particularly be noted that the amplitude of the sinusoidal term in both (4) and (17) is the same and can be obtained directly from $v_0/Z(p)$ by writing $p = j\omega$. Thus

$$I(j\omega) = \frac{v_0}{R + 1/pC} = \frac{v_0}{R + 1/j\omega C} = \frac{v_0}{R} \cdot \frac{1}{1 + 1/j\omega CR}$$

$$= \frac{v_0}{R} \cdot \frac{1}{1 + \alpha/\omega j}$$

and $$|I(j\omega)| = \frac{v_0}{R} \cdot \frac{1}{(1 + \alpha^2/\omega^2)^{\frac{1}{2}}} = \frac{v_0}{R} \cdot \frac{\omega}{(\alpha^2 + \omega^2)^{\frac{1}{2}}}$$

This is, of course, the steady-state a.c. component which exists when all switching transients have died away. It is given by the p-solution but, if we only want this, it is easier to obtain it directly by writing $p = j\omega$. The full p-solution gives the transient term as well.

What we have shown here is that the Heaviside system, although based upon a unit step input stimulus, is by no means confined to it. It can be used with other inputs by translating their time expressions into p-expressions.

It may be helpful to point out that in doing this we do not really lose the unit step. A sine wave is a continuous wave which exists at all times from minus infinity to plus infinity, at any rate in its mathematical form. A sine wave which is suddenly switched to a circuit can be regarded as one starting abruptly at zero time. The product of a continuous sine wave and a unit step is just such a wave. The step is zero prior to $t = 0$ and so the product is also zero. The step is unity after $t = 0$ and so the product has the value of the sine wave.

The fact that we are actually making this product is not always plain for, in the Heaviside system, the step is represented by unity.

Finally, it may be helpful to mention some general points. In (1), if p tends to infinity, the value of $I(p)$ tends to zero, whereas in (2), the value of the corresponding $I(t)$ is zero when t is zero. Again, the limiting value of $I(p)$ in (5) or (6) as p tends to infinity is v_0/R and, if t is replaced by zero in the corresponding $I(t)$ given by (17), we also find that $I(t)$ reduces to v_0/R in view of the definition (3) of θ. This equality is not an accident; the limit of a t-expression when t tends to zero is always the same as the limit of the corresponding p-expression when p tends to infinity, and it is also true that the limit of a t-expression when t tends to infinity equals the limit of the corresponding p-expression when p tends to zero. In the Table, we note that $1/p^n$ corresponds to $t^n/n!$; this indicates a kind of reciprocity between p and t which makes the above-mentioned limit relations seem reasonable, and they can be proved mathematically.

These limit relations are helpful in checking and also when only extreme values of an expression are needed.

So far, only p-expressions in which the degree of the numerator does not exceed that of the denominator have been considered. Occasionally, one may be found in which the numerator is of higher degree. By algebraic long division such an expression can be reduced to the sum of the normal kind and one or more terms which involve positive powers of p. The normal term is then interpreted in the ordinary way but the additional p-terms will be puzzling.

If the p-expression consists of the single term p or $p \times 1$, the Table heading implies that the t-expression is closely associated with the differentiation of unit step which corresponds to 1. Before and after the step transition the stimulus is either zero or constant and so the rate of change is obviously zero. At the step, the rate of change is infinite. A p-term thus represents a "unit impulse", which can be defined as a pulse of infinitesimal duration and infinite amplitude but of unit area. It may be regarded as the limit of a pulse of amplitude $1/\epsilon$ and duration ϵ as ϵ tends to zero.

It is easiest to think of an impulse in terms of current, for this is the rate of change of charge; i.e., $i = dq/dt$. A unit step of current means that the current is zero until $t = 0$ and unity thereafter. Charge is the integral of current and, for a unit step of current, charge increases linearly with time after $t = 0$.

The unit impulse is the differential of the unit step and so a unit impulse of current corresponds to a unit step of charge. The infinite current which flows for an infinitesimal time conveys unit charge.

The concept is valid elsewhere but is not so clearly visualized as in the case of current.

The significance of terms p^2, p^3, etc., can be understood by extending the above argument to the repeated differentiation of unit step. These terms are thus found to be associated with impulses of higher order, which can be regarded as zero except at zero time when they behave still more violently than the unit impulse.

We have mentioned this in passing because, in practice, unit impulses and impulses of higher order do not often arise. They usually correspond to impossible physical conditions, such as the instantaneous charging of a capacitor, and so they are likely to occur unexpectedly only when a circuit has been over-simplified.

Sometimes, however, unit impulse is deliberately adopted as an excitation for a circuit. If we want to know how a circuit behaves when it is shock-excited by a pulse of duration short compared with its time constants, it is simplest to assume that the pulse is of zero duration and the error in doing so is usually quite small. We then choose a unit impulse instead of a unit step as the stimulus and take pE as the p-expression for the input voltage instead of E.

The unit impulse is sometimes written as $\delta(t)$. It is the basis of the Laplace system. This is the main difference between the Heaviside and the Laplace systems. In the first, the p-expression for a unit step is 1 and for a unit impulse p; in the second, the unit impulse corresponds to 1 and the unit step to $1 \times 1/p$.

In a table of Laplace transforms, all the p-expressions are $1/p$ times those in a table of Heaviside transforms; impedances, however, are identical in the two systems. It is unfortunate that p is commonly used in both, but there is a growing tendency to keep p only for the Heaviside system and to adopt s for the Laplace.

Correction

In Table 1 of "Operational Calculas I—General Principles", September issue, p. 345, there are two errors, namely

Item 1 should read
$0(t<0)$, $a(t>0)$ corresponds to a

Item 13 should read
$(e^{-\alpha t}-e^{-\beta t})/(\beta-\alpha)$ corresponds to $p/\{(p+\alpha)(p+\beta)\}$

NEW SEMICONDUCTOR DEVICES

A device for storing images has recently been announced in the U.S. It consists essentially of a mosaic of elements, each made up of a photoconductive part and an electroluminescent part as shown diagrammatically in Fig. 1. If the device is set up in the dark, the application of a voltage as shown has no visible effect, because the high dark resistance of the photoconductor prevents enough voltage from being applied to the electroluminescent layer to excite it. If, however, light is allowed to fall on the device, the electroluminescent part is operated, and the resulting feedback of light to the photoconductor ensures that it remains 'on'. If, therefore, an image is momentarily projected on an array of such elements, it will be retained indefinitely. In practice, mosaics of such elements, each about $\frac{1}{32}$ in. square, have been produced by Sylvania in the form shown in Fig. 2. The elements are shielded from one another by an opaque layer to prevent adjacent ones from triggering one another.

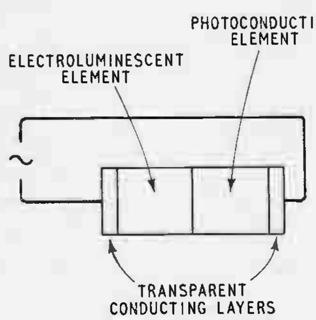


Fig. 1 (*left*). *Basic form of semiconductor element*

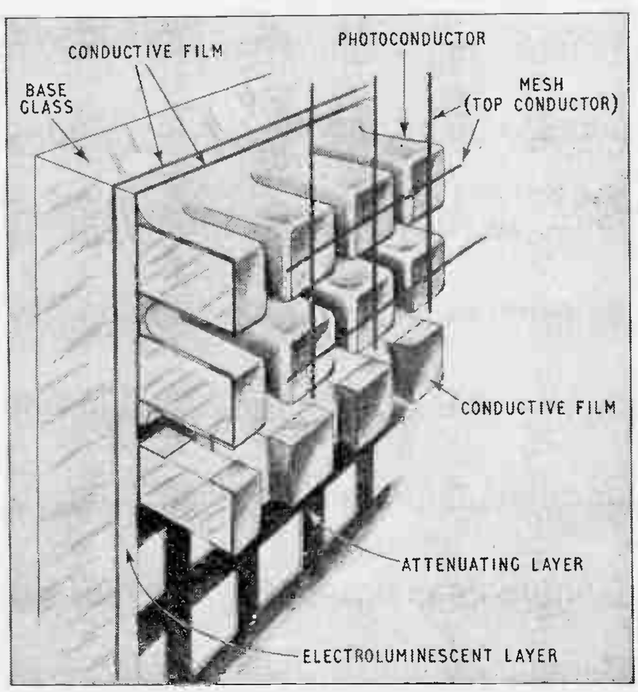


Fig. 2 (*right*). *Part of the mosaic of elements forming a storage plate*

Other applications of photoconducting and electroluminescent combinations under development are infra-red image converters and a coincident-voltage matrix in which any one electroluminescent element can be switched on by energizing two conductors which cross at that element.

An All-Pass Network

TRANSIENT RESPONSE IN TERMS OF A FINITE SERIES OF LAGUERRE FUNCTIONS

By W. Proctor Wilson, C.B.E., B.Sc.(Eng.), M.I.E.E.*

A study of the transient time-response of all-pass networks leads naturally to the choice of a Laguerre function as the 'preferred excitation'. The time response of a lattice all-pass network to an excitation of the form $\cos\omega t$, $\sin\omega t$ $(0<t<\infty)$ is discussed, the analysis being based on the expansion of these functions in terms of Laguerre functions. The transient response for an applied e.m.f. of unit function, $H(t)$, is derived as a special case of the $\cos\omega t$ excitation $(\omega = 0)$.

The expressions obtained for the transient time-responses are believed by the writer to be new. Since numerical values of the Laguerre functions are now available (see Bibliography), the finite series in which the transient responses are expressed can be summed without undue labour in computing.

We consider a network consisting of n stages of a non-dissipative dispersive lattice network, of the form shown in Fig. 1, terminated in its iterative impedance $R = \sqrt{(L/C)}$.

The transfer function of this composite network is readily shown to be†

$$(-1)^n \left[\frac{p-\gamma}{p+\gamma}\right]^n \quad \ldots \quad (1)$$

where $\gamma^2 = 1/LC$, and 'p' has its usual symbolic or operational significance.

Preferred Excitation

The general form of the transfer function leads us, in the interests of simplifying the problem of finding the transient response of the network, to find a class of time functions which can be regarded as a 'preferred excitation'. A time function $h_r(t)$ of which the operational image, $f_r(p)$ say, contains a factor

$$(-1)^r \left[\frac{p-\gamma}{p+\gamma}\right]^r ; \; 0<r<\infty$$

would evidently be of value, since the product of such an image with (1) would yield a new image $f_{r+n}(p)$, and consequently a time response $h_{r+n}(t)$, a function of the same class as the exciting function $h_r(t)$, but of the order $n + r$.

One such family of functions $f_r(p)$ is represented by the image of the Laguerre functions :

$$f_r(p) = \frac{p}{p+\gamma}\left[\frac{p-\gamma}{p+\gamma}\right]^r \doteqdot e^{-\gamma t} L_r(2\gamma t) \quad \ldots \quad (2)$$

where $L_r(t)$ is the Laguerre polynomial defined by*

$$L_r(t) = e^t \frac{d^r}{dt^r}\left[e^{-t}\frac{t^r}{r!}\right] = \sum_{s=0}^{r}(-1)^s \binom{r}{s}\frac{t^s}{s!} \quad \ldots \quad (3)$$

In general, therefore, the response of the network under discussion to a Laguerre excitation will itself be a Laguerre function. For an excitation

$$(-1)^r e^{-\gamma t} L_r(2\gamma t) \; ; \; 0<t<\infty \quad \ldots \quad (4)$$

the response will be

$$(-1)^{n+r} e^{-\gamma t} L_{n+r}(2\gamma t) \; ; \; 0<t<\infty \quad \ldots \quad (5)$$

We note that of the family of 'preferred excitations' (4) the simplest is that of order zero—since $L_0(t) = 1$, this reduces to the simple case of the exciting function

$$e^{-\gamma t} \doteqdot \frac{p}{p+\gamma} ; \; 0<t<\infty \quad \ldots \quad (6)$$

the response to which is evidently

$$(-1)^n e^{-\gamma t} L_n(2\gamma t) \; ; \; 0<t<\infty \quad \ldots \quad (7)$$

This preliminary consideration of 'preferred excitation'† can now assist materially in deriving the transient response of the network to specific excitations of the forms

$$\cos\omega t \; ; \; 0<t<\infty$$
$$\sin\omega t \; ; \; 0<t<\infty$$

of which the former includes excitation by 'unit step', $H(t)$, as a special case $(\omega=0)$.

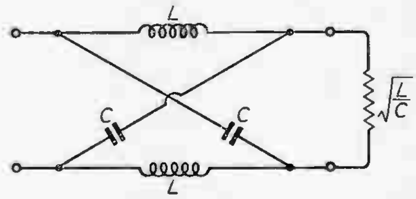


Fig. 1. *Dispersive lattice network*

Response to Excitation $\cos\omega t$, $\sin\omega t$; $\gamma<t<\infty$

The operational images of $\cos\omega t$, $\sin\omega t$ $(0<t<\infty)$, by which we mean $\cos\omega t\, H(t)$, and $\sin\omega t\, H(t)$ where $H(t)$ is Heaviside's unit function, are :

$$\cos\omega t \doteqdot \frac{p^2}{p^2+\omega^2} \; ; \; \sin\omega t \doteqdot \frac{p\omega}{p^2+\omega^2},$$

*Research Department, British Broadcasting Corporation

† The transfer function of the all-pass network derived by interchanging L and C in Fig. 1 is the expression (1) with the $(-1)^n$ factor omitted.

* See Appendix 1

† The preferred excitation in this case, a Laguerre function, can be compared to the preferred excitation in the case of a classic low-pass filter terminated in its iterative impedance. In this case van der Pol has shown that a Bessel function is the natural form of excitation.

and the images of the time-responses of the network to these excitations are respectively

$$(-1)^n \frac{p^2}{p^2+\omega^2}\left[\frac{p-\gamma}{p+\gamma}\right]^n \qquad \cdots \quad (8a)$$

$$(-1)^n \frac{p\omega}{p^2+\omega^2}\left[\frac{p-\gamma}{p+\gamma}\right]^n \qquad \cdots \quad (8b)$$

The formal transformation of these images to the corresponding time functions by means of the appropriate Bromwich contour integral presents considerable complication, because of the pole of order n at $p = -\gamma$. Algebraic manipulations of (8a) and (8b) to convert these expressions (e.g. by partial fractions) into more recognizable images do not appear readily feasible.

The writer has, however, shown that $\cos\omega t$, $\sin\omega t$ are readily expansible in series of Laguerre functions in which a real arbitrary parameter is involved. These expansions are* :

$$\cos\omega t = 2\left[1+\frac{\omega^2}{\alpha^2}\right]^{-\frac{1}{2}} e^{-\alpha t} \times$$
$$\sum_{s=0}^{\infty}(-1)^s\, T_{2s+1}\left\{\left[1+\frac{\omega^2}{\alpha^2}\right]^{-\frac{1}{2}}\right\}L_s(2\alpha t) \qquad \cdots \quad (9a)$$

$$\sin\omega t = 2\left[1+\frac{\omega^2}{\alpha^2}\right]^{-\frac{1}{2}} e^{-\alpha t} \times$$
$$\sum_{s=0}^{\infty}(-1)^s\, U_{2s+1}\left\{\left[1+\frac{\omega^2}{\alpha^2}\right]^{-\frac{1}{2}}\right\}L_s(2\alpha t) \qquad \cdots \quad (9b)$$

where $\alpha (\neq 0)$ is the arbitrary parameter, and $T_r\,(x)$, $U_r\,(x)$ are the normalized even and odd Tchebyshev functions defined as follows :

$$T_r(x) = \cos(r\cos^{-1}x)\ ;\ U_r(x) = \sin(r\cos^{-1}x)$$

Armed with this convenient pair of expansions, and identifying the arbitrary α of (9a, b) with the γ in the transfer function (1), we can now derive the transient response to impressed excitations $\cos\omega t$, $\sin\omega t$. We note that each term in the infinite series is of the form (4), associated with a Tchebyshev coefficient. It follows therefore, from previous reasoning, that the total time responses $w_n(t), x_n(t)$ of the network to these two excitations are :

For excitation $\cos\omega t$:

$$w_n(t) = 2\mu e^{-\gamma t}\sum_{s=0}^{\infty}(-1)^{n+s}\, T_{2s+1}(\mu)\, L_{n+s}(2\gamma t) \quad (10a)$$

For excitation $\sin\omega t$:

$$x_n(t) = 2\mu e^{-\gamma t}\sum_{s=0}^{\infty}(-1)^{n+s}\, U_{2s+1}(\mu)\, L_{n+s}(2\gamma t) \quad (10b)$$

where

$$\mu = \left[1+\frac{\omega^2}{\gamma^2}\right]^{-\frac{1}{2}}$$

We emphasise here that the expressions (10a), (10b) are the total time responses—they include the transient and the steady state components. We are interested in the *transient* response, and to obtain this we evaluate $w_n(\infty)$ and $x_n(\infty)$, the respective steady state responses, and subtract $w_n(t)$, $x_n(t)$, from them.

In (1) we now substitute $j\omega$ for p, and note that the transfer function of the network for a constant excitation of frequency $\omega/2\pi$ is

$$(-1)^n\left[\frac{j\omega-\gamma}{j\omega+\gamma}\right]^n = e^{-j2n\tan^{-1}\frac{\omega}{\gamma}} \qquad \cdots \quad (11)$$

It follows that the response of the network to a continuous excitation $e^{-j\omega t}$ is therefore

$$\exp.\left[-j\left(\omega t - 2n\tan^{-1}\frac{\omega}{\gamma}\right)\right] = \cos\left[\omega t - 2n\tan^{-1}\frac{\omega}{\gamma}\right]$$
$$-j\sin\left[\omega t - 2n\tan^{-1}\frac{\omega}{\gamma}\right] \qquad \cdots \quad (12)$$

The steady state response of the network to suddenly applied excitations $\cos\omega t$, $\sin\omega t$ corresponds to the real and imaginary expression on the r.h.s. of (12), under the assumption that these excitations were applied at $t = 0$, and that infinite time has elapsed. This being the case, we may make use again of the Laguerre expansions for $\cos\omega t$, $\sin\omega t$ $(t>0)$ (9a, b).

Bearing in mind that

$$\cos(\omega t - 2n\cos^{-1}\mu) = T_{2n}(\mu)\cos\omega t + U_{2n}(\mu)\sin\omega t, \qquad \cdots \quad (13a)$$

$$\sin(\omega t - 2n\cos^{-1}\mu) = -U_{2n}(\mu)\cos\omega t + T_{2n}(\mu)\sin\omega t \qquad \cdots \quad (13b)$$

it can readily be shown, by simple trigonometrical manipulations, that

$$w_n(\infty) = \cos(\omega t - 2n\cos^{-1}\mu)$$
$$= 2\mu e^{-\gamma t}\sum_{s=0}^{\infty}(-1)^s\, T_{2(s-n)+1}(\mu)\, L_s(2\gamma t) \qquad \cdots \quad (14a)$$

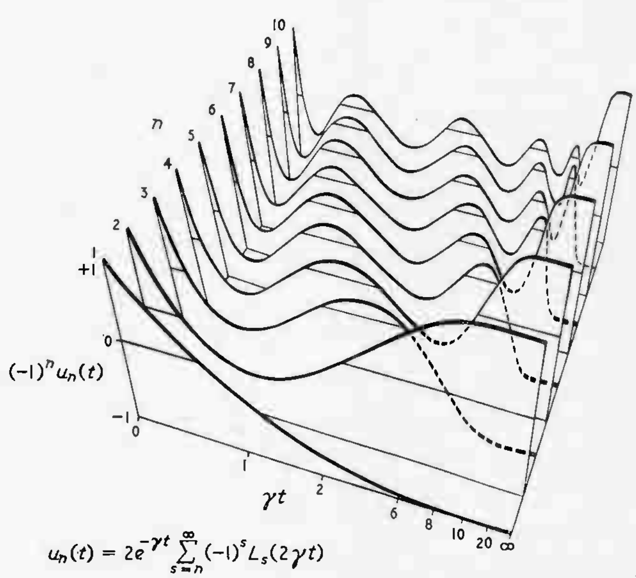
$$x_n(\infty) = \sin(\omega t - 2n\cos^{-1}\mu)$$
$$= 2\mu e^{-\gamma t}\sum_{s=0}^{\infty}(-1)^s\, U_{2(s-n)+1}(\mu)\, L_s(2\gamma t) \qquad \cdots \quad (14b)$$

where again

$$\mu = \left[1+\frac{\omega^2}{\gamma^2}\right]^{-\frac{1}{2}}$$

It is now convenient to reconsider the expressions for

Fig. 2. *Time response of n stages of network in Fig. 1 terminated by $\sqrt{L/C}$ to impressed excitation of unit function $H(t)$*
Note : Ordinates multiplied by $(-1)^n$ and abscissa drawn to a tangent-scale in order to simplify the diagram



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* For proof see Appendix 2.

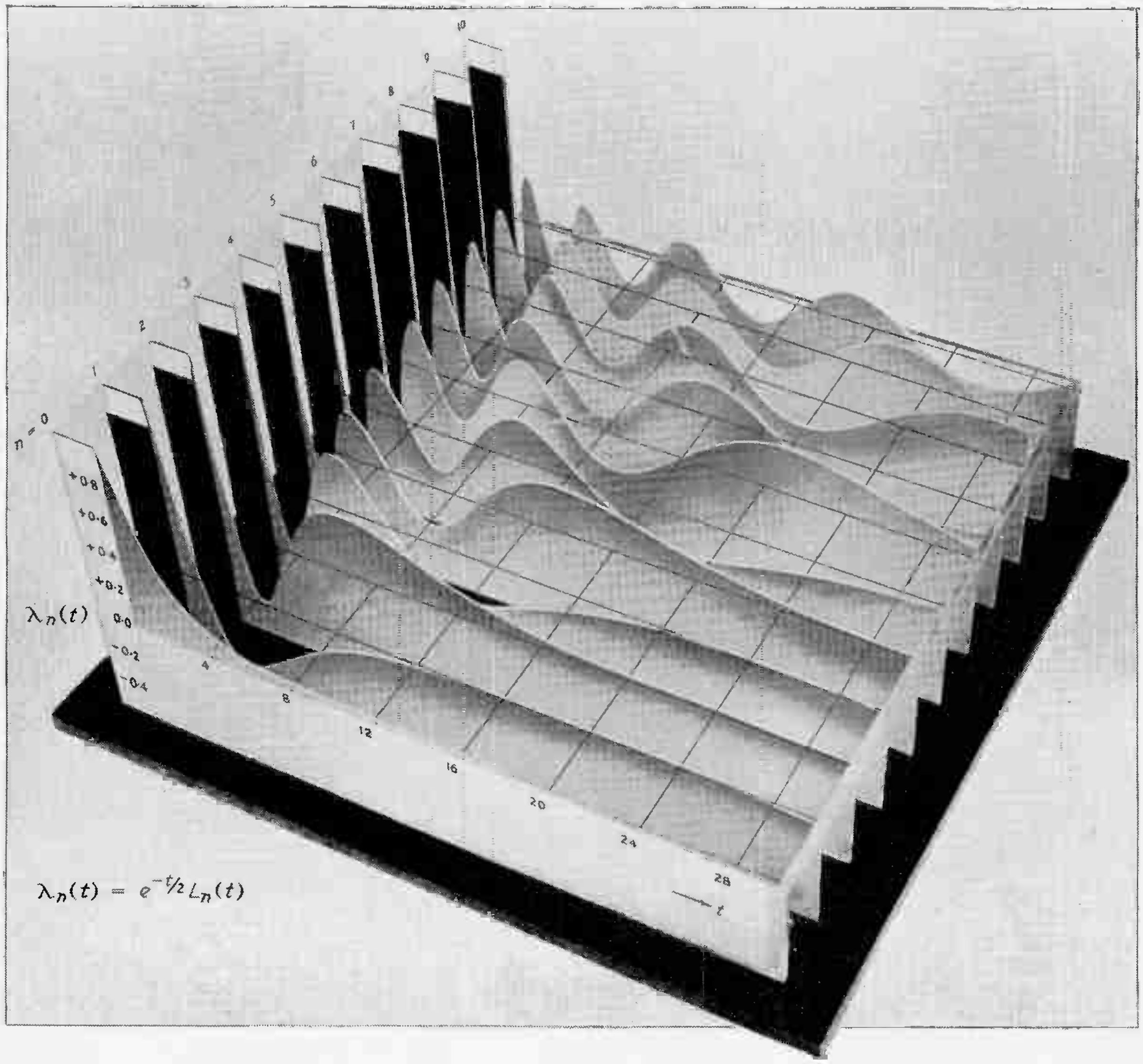

Fig. 3. *March of the Laguerre functions* $\lambda_n(t) = e^{-\frac{1}{2}t}L_n(t)$ (*of orders* 0 *to* 10)

$w_n(t), x_n(t)$ in (10a) and (10b), and to substitute in them a new variable of summation $s' = n + s$.

We then obtain

$$w_n(t) = 2\mu e^{-\gamma t} \sum_{s'=n}^{\infty} (-1)^{s'} T_{2(s'-n)+1}(\mu)\, L_{s'}(2\gamma t) \qquad (15a)$$

$$x_n(t) = 2\mu e^{-\gamma t} \sum_{s'=n}^{\infty} (-1)^{s'} U_{2(s'-n)+1}(\mu)\, L_{s'}(2\gamma t) \qquad (15b)$$

where μ has the same meaning as before.

Subtracting $w_n(t), x_n(t)$ from $w_n(\infty), x_n(\infty)$ in equations (15) and (14) we obtain the transient response of the network for suddenly applied $\cos\omega t$, $\sin\omega t$ excitation [dropping the dash in s' in (15)].

$$w_n(\infty) - w_n(t) = 2\mu e^{-\gamma t} \sum_{s=0}^{n-1} (-1)^s\, T_{2(n-s)-1}(\mu)\, L_s(2\gamma t) \quad \ldots \quad (16a)$$

$$x_n(\infty) - x_n(t) = -2\mu e^{-\gamma t} \sum_{s=0}^{n-1} (-1)^s\, U_{2(n-s)-1}(\mu)\, L_s(2\gamma t) \quad \ldots \quad (16b)$$

[noting that $T_{-r}(\mu) = T_r(\mu)$, $U_{-r}(\mu) = -U_r(\mu)$]

We note that these transient solutions are in terms of *finite* Laguerre series, and thus readily obtained numerically by the use of tables of these functions.

Response of Network to Unit Function $H(t)$

From (15a) and (16a) we may now derive the response of the network to applied unit step $H(t)$, by simply equating ω to zero, or μ to unity.

Carrying out this procedure, we find that the indicial* response $u_n(t)$†, is

$$u_n(t) = w_n(t)|_{\omega=0} = 2e^{-\gamma t} \sum_{s=n}^{\infty} (-1)^s\, L_s(2\gamma t) \quad \ldots \quad (17)$$

This is the total time response to excitation $H(t)$.

The transient response is evidently $u_n(\infty) - u_n(t)$ or

$$1 - u_n(t) = 2e^{-\gamma t} \sum_{s=0}^{n-1} (-1)^s\, L_s(2\gamma t) \quad \ldots \quad \ldots \quad (18)$$

* The term 'indicial' is borrowed from Carson's concept of indicial admittance, defining response to unit function excitation, $H(t)$.

† For an alternative expression for $u_n(t)$, see Appendix 3.

since $T_r(1) = 1$, and

$$u_n(\infty) = w_n(\infty)|_{\omega=0} = 2e^{-\gamma t} \sum_{s=0}^{\infty} (-1)^s L_s(2\gamma t) = H(t) \quad \ldots \quad \ldots \quad (19)$$

The (total) time response (17) is shown in Fig. 2, for values of n from 1 to 10. The ordinates of the three-dimensional diagram are $(-1)^n u_n(t)$. This scale has been chosen for convenience. It also corresponds to the time response of n stages of the network in Fig. 1, when L and C are interchanged. In order to show the whole time-range, from zero to infinity, a tangent-scale has been chosen for the abscissae of Fig. 2.

The author's thanks are due to Mr. J. W. Head for his kindness in checking the mathematical analysis in this paper, and to Mr. T. N. J. Archard for his advice and assistance in preparing Figs. 2 and 3.

Acknowledgment is also due to the Director of Engineering, British Broadcasting Corporation, for permission to publish this paper.

The expressions in (10a), (10b), (14a), (14b), (15a), (15b), (16a), (16b), (17) and (18) are believed by the writer to be new relationships.

The march of the Laguerre function of orders 0 to 10 is shown in Fig. 3.

APPENDIX 1

Laguerre Functions

The Laguerre functions e^{-t}. $L_n(2t)$, of argument $2t$, are based upon the normalized polynomials

$$L_r(t) = e^t \frac{d^r}{dt^r}\left[e^{-t}\frac{t^r}{r!}\right] = \left[\frac{d}{dt} - 1\right]^r \frac{t^r}{r!} \quad \ldots \quad \ldots \quad \ldots \quad (i)$$

The functions are orthonormal, in the sense that

$$\int_0^\infty e^{-\tau} L_m(\tau) L_n(\tau)\, d\tau = \begin{cases} 0\,;\ m \neq n \\ 1\,;\ m = n \end{cases} \quad \ldots \quad \ldots \quad \ldots \quad (ii)$$

It follows from (ii) that functions of time existing in the epoch $0 < t < \infty$ can be represented as Laguerre function series of the form

$$e^{-\frac{1}{2}at} \sum_{s=0}^{r} A_s L_s(at)\,;\ r < \infty,$$

(where A_s is a constant or a function of a) provided that, if $r = \infty$, the series converges.

The Laguerre functions $\lambda_r(t) = e^{-\frac{1}{2}t} L_r(t)$ are, for integer values of r, such that

$$\lambda_r(0) = 1\,;\ \lambda_r(\infty) = 0\,;\ \lambda_0(t) = e^{-\frac{1}{2}t}.$$

These functions decrease in an oscillatory manner in the first part of their range, the function of order r possessing r real finite roots. After passing through the rth zero, the modulus of the function decreases monotonically with t to zero at $t = \infty$. For r very large the functions tend to the approximation

$$e^{-\frac{1}{2}t} L_r(t) \approx J_0[2\sqrt{(rt)}] \quad \ldots \quad \ldots \quad (iii)$$

Tables of $\lambda_r(t)$ for $r = 0(1)20$ and $t = 0(0{\cdot}1)1(0{\cdot}2)3(0{\cdot}5)6(1)14(2)40(5)100$ have been published in a monograph by J. W. Head and the author. (See bibliography, reference 6.)

APPENDIX 2

Expansion of cos ωt, sin ωt in terms of Laguerre functions. $(0 < t < \infty)$

We make use of the known expansion

$$e^{-2at} = \sum_{s=0}^{\infty} \frac{a^s}{(1+a)^{s+1}} L_s(2t) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (i)$$

It follows that

$$e^{-(2a+1)t} = e^{-t} \sum_{s=0}^{\infty} \frac{a^s}{(1+a)^{s+1}} L_s(2t) \quad \ldots \quad \ldots \quad \ldots \quad (ii)$$

which is an expansion in Laguerre functions of the form $e^{-t} L_s(2t)$, or $\lambda_s(2t)$.

Let $2a + 1 = j\omega$. We now seek a transformation of the coefficients $a^s/(1+a)^{s+1}$ in terms of ω. It will be seen that

$$\begin{aligned}\frac{a^s}{(1+a)^{s+1}} &= 2\frac{(j\omega - 1)^s}{(j\omega + 1)^{s+1}} \\ &= (-1)^s\, 2(1+\omega^2)^{-\frac{1}{2}} \exp\{-j(2s+1)\tan^{-1}\omega\} \\ &= (-1)^s\, 2(1+\omega^2)^{-\frac{1}{2}} [\cos\{(2s+1)\cos^{-1}(1+\omega^2)^{-\frac{1}{2}}\} \\ &\qquad - j\sin\{(2s+1)\cos^{-1}(1+\omega^2)^{-\frac{1}{2}}\}] \\ &= (-1)^s\, 2\mu\,\{T_{2s+1}(\mu) - j\,U_{2s+1}(\mu)\} \quad \ldots \quad \ldots \quad (iii)\end{aligned}$$

where $\mu = (1+\omega^2)^{-\frac{1}{2}}$; $T_r(\mu)$, $U_r(\mu)$ are the even and odd normalized Tchebyshev polynomials, that is, $\cos(r\cos^{-1}\mu)$ and $\sin(r\cos^{-1}\mu)$ respectively.

It is now evident that by selecting the real and imaginary parts of the expansion of $e^{-j\omega t}$ in Laguerre functions with Tchebyshev coefficients we arrive at the two expansions:

$$\cos\omega t = 2\mu e^{-t} \sum_{s=0}^{\infty} (-1)^s\, T_{2s+1}(\mu)\, L_s(2t) \quad \ldots \quad \ldots \quad (iv)$$

$$\sin\omega t = 2\mu e^{-t} \sum_{s=0}^{\infty} (-1)^s\, U_{2s+1}(\mu)\, L_s(2t) \quad \ldots \quad \ldots \quad (v)$$

We may make the relationships (iv) and (v) more general by writing ωt as $\omega_0 t \cdot \frac{\omega}{\omega_0}$, say. Then

$$\cos\omega t = 2\mu e^{-\omega_0 t} \sum_{s=0}^{\omega} (-1)^s\, T_{2+1}(\mu)\, L_s(2\omega_0 t) \quad \ldots \quad \ldots \quad (vi)$$

$$\sin\omega t = 2\mu e^{-\omega_0 t} \sum_{s=0}^{\infty} (-1)^s\, U_{2s+1}(\mu)\, L_s(2\omega_0 t) \quad \ldots \quad \ldots \quad (vii)$$

where in this case $\mu = \left[1 + \frac{\omega^2}{\omega_0^2}\right]^{-\frac{1}{2}}$, and ω_0 is arbitrary.

APPENDIX 3

Alternative Derivation of Response to Unit Function Excitation

The transfer function $(-1)^n \left[\frac{p-\gamma}{p+\gamma}\right]^n$ is, of course, the 'image' of the time response of the network. We note that by virtue of the identities

$$\left[\frac{p-\gamma}{p+\gamma}\right]^n = \begin{cases} \dfrac{p+\gamma}{p} \cdot \dfrac{p}{p+\gamma} \cdot \left[\dfrac{p-\gamma}{p+\gamma}\right]^n & \ldots \quad \ldots \quad (i) \\[2ex] \dfrac{p-\gamma}{p} \cdot \dfrac{p}{p+\gamma} \cdot \left[\dfrac{p-\gamma}{p+\gamma}\right]^{n-1} & \ldots \quad \ldots \quad (ii) \end{cases}$$

we can arrive at two different forms for the time response. Remembering that

$$\frac{p}{p+\gamma} \cdot \left[\frac{p-\gamma}{p+\gamma}\right]^r \doteqdot e^{-\gamma t} L_r(2\gamma t) \quad \ldots \quad \ldots \quad \ldots \quad (iii)$$

we interpret (i) as

$$(-1)^n \left[\frac{p-\gamma}{p+\gamma}\right]^n \doteqdot u_n(t) = (-1)^n \{1 + \gamma\int_0^t dt\}\, e^{-\gamma t} L_n(2\gamma t) \quad (iv)$$

and (ii) as

$$(-1)^n \left[\frac{p-\gamma}{p+\gamma}\right]^n \doteqdot u_n(t) = (-1)^n \{1 - \gamma\int_0^t dt\}\, e^{-\gamma t} L_{n-1}(2\gamma t) \quad (v)$$

These solutions, involving the sum or difference of a Laguerre function and its integral, are due to K. W. Wagner[7]. The solution, although elegant in form, is inconvenient, since no tables of the integrals of Laguerre functions appear to exist.

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[2] B. van der Pol, *Phil. Mag.* 1929, Vols. VII, VIII.
[3] B. van der Pol and K. F. Niessen, *Phil. Mag.*, 1931, Vol. XI, 1932; Vol. XIII.
[4] B. van der Pol, "Operational Calculus based on the two-sided Laplace Integral" (Cambridge, 1950).
[5] K. W. Wagner, "Operatorenrechnung" (Leipzig, 1940).
[6] J. W. Head and W. Proctor Wilson, "Laguerre Functions: Tables and Properties. I.E.E. Monograph 183R, June 1956.
[7] K. W. Wagner, "Operatorenrechnung und Laplacesche Transformation," 2nd Edition, 1950.

New Books

Transistor Circuits and Applications

By JOHN M. CARROLL. Pp. 283 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 56s. 6d.

Mr. Carroll, an associate editor of the American periodical *Electronics*, has compiled a collection of 106 articles reprinted from that journal. The format of the book is uniform with earlier volumes from the same source, such as the well-known "Handbook of Industrial Electronic Circuits". Inevitably the articles are uneven in standard and there is some overlapping. Moreover, since the articles originally appeared during the years 1950 to 1956, some of the earlier circuits described are already obsolete (e.g., amplifiers using point-contact transistors). However, these defects are the result of the rapid progress which the transistor art has made during the last few years, and the book does contain a representative collection of suitable transistor applications.

Although the bulk of the contents is devoted to practical circuit applications, the first 14 articles deal with principles of circuit design. These include useful tables of fundamental design equations using both resistance parameters and h-parameters. The second group of 16 articles deals with amplifiers and includes articles on temperature stabilization and on negative feedback. The articles on negative feedback are valuable because this subject is neglected in current textbooks on transistor circuits. Twelve articles deal with oscillators and 14 with pulse circuits such as triggers, flip-flops and counters.

The remaining articles describe applications of transistor circuits to broadcast receivers, military and communications equipment, computers, servomechanisms and industrial, scientific and medical devices. The equipment described ranges from telemetering equipment for missiles to a toy electronic organ. Most of the diagrams give component values. Naturally, all the transistor types quoted are American, but this need give little concern as long as p-n-p alloyed junction types are used. When n-p-n transistors are employed, the circuit diagrams become of less immediate practical value to us. When tetrodes, silicon transistors, diffused-base transistors and double-base diodes are used the British reader can only read with envy and hope that our own manufacturers will ultimately produce a more extensive range of devices. J.E.F.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Coherent and Incoherent Detectors

SIR,—Mr. Kitai's paper in the March issue[1] seems to be misleading, if not wrong, in at least two respects. These are (a) the statement that the output signal-to-noise ratio (S/N) of a coherent detector is 3 dB greater than that of a linear diode detector, when no post-detector filtering is permitted; and (b) the implication that it is not possible to improve the output S/N of a coherent detector by filtering if the input signal is modulated.

In connection with (a) above, the mention of linear diode detectors appears irrelevant. The figure of 3 dB as shown by Smith[3], Tucker[2], and others, refers to the ratio of output to input N/S for a coherent detector. If a diode, or other non-coherent detector is to be compared with a coherent detector, account must be taken of the actual input S/N. This is because, as shown by Smith, and by Tucker, the output S/N for a diode detector is proportional to the square of the input S/N when this is less than unity. On the other hand, the output S/N for a coherent detector is directly proportional to the input S/N for all values.

So far as (b) above is concerned, if the modulation is random and covers a wide bandwidth, as in speech, it is not possible to use a filter either before or after the demodulator which preserves the signal and removes noise of the same frequency. In this I am in agreement with Kitai. However, there are other important types of modulation in which the S/N may be improved by post-demodulator filtering in a manner exactly analogous to that described by Kitai in connection with an unmodulated signal.

For example, consider the case of a carrier sinusoidally modulated by a single known frequency. A bandpass filter centred on the modulation frequency, and of as narrow a bandwidth as the signal will permit, will attenuate noise at other frequencies, and greatly improve the output S/N ratio. In fact, this principle can be extended to more complicated cases in which more than one pair of sidebands accompany the signal. I conclude that the appropriate filter following a coherent detector is always capable of improving the S/N ratio providing that the periodic modulation waveform is known beforehand.

It is perhaps relevant here to draw attention to recommendations[4] made by D. G. Tucker, who suggests the following usage:

"A demodulator is a device for producing the modulating signal from an envelope-modulated signal".

"A detector is a device for producing the modulating signal from an envelope-modulated signal by means of a circuit having a symmetrical response (e.g., a rectifier), without the introduction of a local oscillator".

Thus, it appears desirable to refer to coherent demodulators rather than coherent detectors, since the latter seems to be a contradiction in terms.

I personally would go further, and suggest that the term 'demodulator' be reserved for those devices which produce the waveform of the incoming envelope-modulated signal by multiplying (or modulating) it with an external signal. Thus coherent, synchronous, homodyne or product detectors, as they are variously known, would all become demodulators.

Dominion Physical Laboratory, D. D. CROMBIE
Lower Hutt, New Zealand.
7th August 1957

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[2] D. G. Tucker. *Wireless Engineer*, 1952, Vol. 29, p. 184.
[3] R. A. Smith. *Proc. Instn elect. Engrs*, Part IV, 1951, Vol. 98.
[4] D. G. Tucker. "Modulators and Frequency Changers", (MacDonald and Co., London), 1953, p. 17.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for August* 1957

Date 1957 August	MSF 60 kc/s 2030 G.M.T. parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. parts in 10^8
1	− 3	0
2	− 4	0
3	− 4	0
4	− 4	+ 1
5	− 4	+ 1
6	− 4	+ 1
7	− 3	N.M.
8	− 3	− 2
9	− 3	− 1
10	− 3	− 1
11	− 3	0
12	− 2	+ 1
13	N.M.	+ 1
14	− 2	+ 3
15	− 2	+ 2
16	− 2	+ 3
17	− 2	+ 3
18	N.M.	N.M.
19	N.M.	+ 4
20	− 2	+ 4
21	N.M.	+ 4
22	− 2	+ 5
23	− 2	− 1
24	− 2	N.M.
25	− 2	N.M.
26	− 2	− 2
27	− 1	− 1
28	N.M.	− 1
29	− 1	− 1
30	− 1	− 1
31	− 1	N.M.

* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

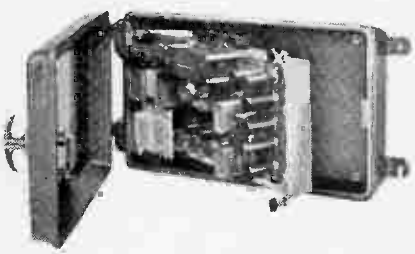
New Products

Ultrasonic Cleaning Bath

A radial magnetostriction transducer, which is claimed to focus energy in the centre of the cleaning-fluid container, is incorporated in the Mullard ultrasonic cleaning bath type L.276. The frequency employed is 20 kc/s: this is suitable where tenacious films have to be removed. The actual cleaning bath is a standard 250 c.c. beaker, which stands in the centre of the transducer annulus. The latter is immersed in water which serves both as a coolant and a means of coupling the transducer to the bath. About 50 W ultrasonic energy is supplied from a valve generator.
Mullard Ltd.,
Mullard House, Torrington Place, London, W.C.1.

Universal Batching Counter

The electronic counter illustrated is described by the makers as an instrument designed for both batching and counting. The maximum speed is 6,000 per minute, and any number of items from 1 to 1,000 may be batched. Input signals are derived from a photocell. The instrument will



operate electro-mechanical flaps, 'kickers', and diverting guides such as are incorporated in conveyor-belt systems, etc.
Atkins, Robertson & Whiteford Ltd.,
Industrial Estate, Thornliebank, Glasgow.

Preset Potentiometer

The Plessey type PP preset potentiometer is described by the manufacturers as a 'moulded track' component which costs no more than a conventional sprayed track potentiometer but has the advantages of low noise, better stability, greater resistance to humidity, and longer life. The power rating for the whole resistive element is 0·5 W, and the resistances available range from 1 Ω to 2 MΩ with a tolerance of −20%.

The type PP is stated to be a suitable component for use in a.c./d.c. equipment, and in equipment with tropical finish.
The Plessey Company Ltd.,
Ilford, Essex.

Temperature Control Unit

The temperature control unit type N241 is described as a highly sensitive robust

instrument suitable for incorporation in industrial installations. It operates over the temperature range −70° C to +600° C, control being effected by a heavy duty output relay fitted with three sets of contacts:— 8 amp. single-pole changeover, 1 amp. single-pole changeover and 1 amp. single-pole make-before-break.

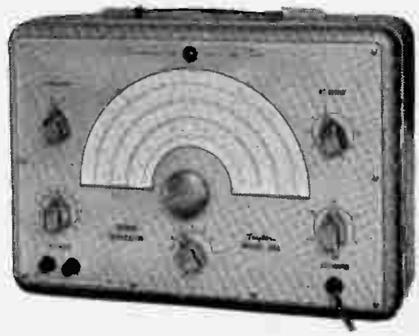
The sensitive element is a small platinum resistance, robust and stable. It is available in two forms, either ⅛ in. diameter by 2 in. long or ¼ in. diameter by 6 or 24 in. long.

Operation of the output relay is effected by a 0·1% change in the value of the temperature-sensitive resistor which represents a temperature change of about 0·4° C.
Airmec Limited,
High Wycombe, Buckinghamshire.

Signal Generator

The Model 68A signal generator is intended mainly for use in the servicing of radio and television receivers. The frequency coverage is 100 kc/s to 220 Mc/s. Two oscillators are incorporated; one of these covers the lower-frequency bands by means of a band-switching arrangement. The other covers the range 110–220 Mc/s without band-switching. This arrangement is intended to avoid the frequency errors which often result from band-switching at high frequencies.

The instrument is carefully screened to minimize radio-frequency leakage, and



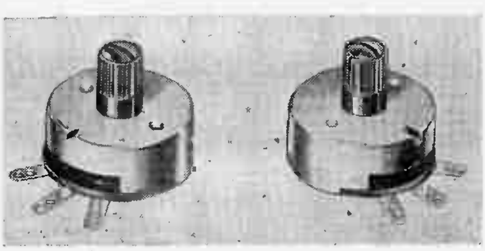
contains a multi-section mains filter effective over the entire frequency range of the oscillator.

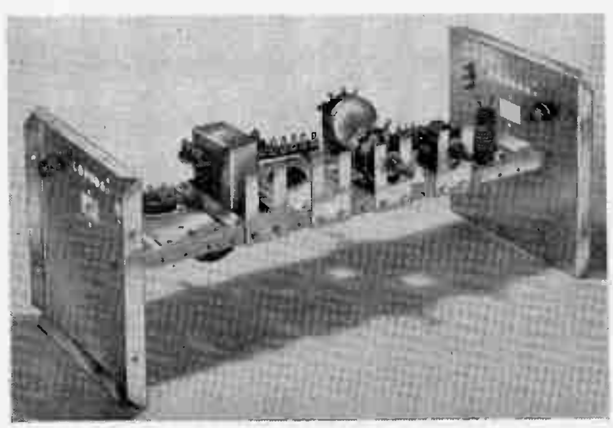
The audio modulating frequency of 400 c/s is also available for external testing.
Taylor Electrical Instruments Ltd.,
419-424 *Montrose Avenue, Slough, Bucks.*

Standard Chassis

The chassis illustrated is obtainable from Cowell Developments. The main framework consists of end-plates and two runners. Component plates of two different widths can be mounted on the runners. These plates are available either blank or with holes for mounting valve-holders. Slotted 'potentiometer plates' can be screwed to the runners as shown.

The end plates have holes for standard types of toggle switch, coaxial sockets, chassis connector, and terminal strip. If desired, the runners can be fixed so that the component plates are sloping instead of horizontal.
Cowell Developments,
67 *Long Drive, East Acton, London, W.3.*





Abstracts and References

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The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publishers concerned.*

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ACOUSTICS AND AUDIO FREQUENCIES

534.2 **3000**

Formulation of Wave Propagation in Infinite Media by Normal Coordinates with an Application to Diffraction.—M. A. Biot & I. Tolstoy. (*J. acoust. Soc. Amer.*, March 1957, Vol. 29, No. 3, pp. 381–391.) The mathematical basis of the technique is discussed fully. The method is applied to the example of diffraction by wedges and corners for an idealized point-source transient explosion.

534.2 **3001**

Perturbation of a Sound Field by a Rigid Cylinder.—O. Brosze. (*Elektronische Rundschau*, Jan. 1957, Vol. 11, No. 1, pp. 19–22.) A formula is derived for the sound field in the vicinity of a rigid, infinitely long cylinder. Results of acoustic measurements are compared in graphical form with calculated values; good agreement is found.

534.2 **3002**

Scattering of Sound Waves by Small Inhomogeneities in a Waveguide.—M. A. Isakovitch. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 37–45.) Analysis is presented for a waveguide (*a*) which is completely filled with a medium the refractive index of which varies from point to point, and (*b*) with rough interior walls and filled with a homogeneous medium.

534.2-14 **3003**

Phenomenological Theory of the Molecular Absorption and Dispersion of Sound in Fluids and the Relation between the Relaxation Time of the Internal Energy and the Relaxation Time of the Internal Specific Heat.—O. Nomoto. (*J. phys. Soc. Japan*, Jan. 1957, Vol. 12, No. 1, pp. 85–99.)

534.2-8 **3004**

Note on Finite-Amplitude Waves in Liquids.—R. T. Beyer & V. Narasimhan. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, p. 532.) See 1999 of August. The frequency dependence of the absorption-coefficient/acoustic-pressure relation is substantiated by other data.

534.2-8 : 539.37 **3005**

Frequency Dependence of Ultrasonic Attenuation and Velocity on Plastic Deformation.—A. Hikata & R. Truell. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 522–523.) The results of measurements at 5 and 10 Mc/s on Al are in accordance with those expected from dislocation damping theory.

534.23-14 **3006**

Measurements of the Attenuation of Low-Frequency Underwater Sound.—M. J. Sheehy & R. Halley. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 464–469.) Values for the range 20–200 c/s were obtained.

534.231 : 534.87 **3007**

Calculations of the Sound Field in the Focal Region of a Cylindrical Focusing System.—I. N. Kanevski & L. D. Rozenberg. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 46–61.) An analysis is made of sound fields with wave fronts of finite or infinite width near the axis of the system, assuming that the wavelength is small in comparison with the focal length of the system. Results show that in a direction normal to the axis of the wave front and in the focal plane the potential has null points, but in a direction perpendicular to both the axis and the focal plane the potential has minima but no null points.

534.232-8 : 546.431.824-31 **3008**

Barium Titanate Ultrasonic High-Intensity Radiator.—I. N. Kogan & L. I. Menes. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 62–64.) The high-power transducer described comprises a $BaTiO_3$ plate of 38 mm diameter backed by a 0·01-mm air film and a water-cooled reflector; the transducer is capable of continuous transmission in water at a frequency of 400–800 kc/s and an intensity of 15 W/cm².

534.26 **3009**

On the Diffraction of Sound Waves in a Viscous Medium.—J. B. Alblas. (*Appl. sci. Res.*, 1957, Vol. A6, No. 4, pp. 237–262.) The theory of the diffraction of a sound wave at a half-plane barrier is extended to the case of propagation in a viscous medium.

534.26 **3010**

Approximate Methods in High-Frequency Scattering.—D. S. Jones. (*Proc. roy. Soc. A*, 12th March 1957, Vol. 239, No. 1218, pp. 338–348.) In considering the diffraction of a high-frequency plane sound wave two approximate methods of deriving the scattering coefficient of two-dimensional obstacles without edges are described. In the first method a simple field which satisfies approximately the boundary condition near the points of glancing incidence is found. Elsewhere the geometrical acoustics field is used. The scattering coefficient is about 7% in error. In the second method Fourier transforms are employed to find a field which satisfies the boundary condition over a wider region. This leads to results which, for the circular cylinder, are in complete agreement with those of the exact theory.

534.6 : 621.395.623 **3011**

Contribution to the Study of the Artificial Ear. The New Model Artificial Ear at the Centre National d'Etudes des Télécommunications.—P. Chavasse. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2014–2017.) See also 1564 of 1950.

534.62 **3012**

Investigation of Sound-Absorbing Constructions for the Anechoic Chamber of the Physical Faculty of the Moscow State University.—K. A. Velizhanina & S. N. Rzhevkin. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 23–28.) Experimental results indicate that the best sound absorber for frequencies down to 80 c/s is a glass-fibre cone with a packing density of $0\cdot12$–$0\cdot14$ g/cm^3 and with an air gap between the wall and the base of the cone equal to $\frac{1}{3}$ of the cone height. Graphs show the frequency characteristics of the sound reflection coefficient of glass-fibre cones for various packing densities and cone dimensions.

534.75 **3013**

Downward Spread of Masking.—W. Spieth. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 502–505.) Bands of noise have a considerable masking effect on lower frequency levels, a pure tone has not.

534.75 **3014**

Noise - Masked Thresholds as a Function of Tonal Duration and Masking Noise Bandwidth.—P. M. Hamilton. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 506–511.)

534.75 **3015**

Remote Masking in Selected Frequency Regions.—B. H. Deatherage, R. C. Bilger & D.A. Eldredge. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 512–514.) Regular variations in the envelope of the masking sound at say 500 c/s produce remote masking in the 500-c/s region; while irregular variations produce equal amounts of masking everywhere outside the region.

534.75 **3016**

Simultaneous Two-Tone Pitch Discrimination.—W. R. Thurlow & S. Bernstein. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 515–519.) An investigation at frequencies of 200 c/s, 1, 4, 6 and 10 kc/s.

534.75 **3017**

Frequency Difference Limens for Narrow Bands of Noise.—R. M. Michaels. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 520–522.) None of the difference limens for noise approach that of the pure tone, all being at least twice as large.

534.75 **3018**

Signal Detection as a Function of Signal Intensity and Duration.—D. M. Green, T. G. Birdsall & W. P. Tanner, Jr. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 523–531.) An attempt to determine the surface of detectabilities in the space determined by signal duration, intensity and detectability. Results are consistent with data of previous research.

534.75 : 621.39 **3019**

Monitoring Task in Speech Communication.—J. P. Egan. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 482–489.) A quantitative description of the monitor's behaviour in terms of the operating characteristic and the articulation-criterion function.

534.78 : 621.39 **3020**

Estimates of the Maximum Precision Necessary in Quantizing Certain 'Dimensions' of Vowel Sounds.—J. L. Flanagan. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 533–534.) A discussion of quantization based on frequencies, formant amplitudes and fundamental vocal frequency for reduced-bandwidth transmission of speech.

534.78 : 681.613 **3021**

The Phonetograph and Subformants.—J. Dreyfus-Graf. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Feb. 1957, Vol. 35, No. 2, pp. 41–59. In French.) Details of the latest model and outline of future development of the phonetic speech-transcription equipment described earlier (1568 of 1953). Oscillograms resulting from investigations of the information content of various speech components are discussed.

534.844.5 **3022**

Correlation Criterion of the Optimum of Reverberation.—V. V. Furduev. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 74–80.) A discussion. The optimum reverberation time is related to the type of signal rather than the volume of the auditorium.

534.845 **3023**

Absorption of Sound by Patches of Absorbent Materials.—R. K. Cook. (*J. acoust. Soc. Amer.*, March 1957, Vol. 29, No. 3, pp. 324–329.) Exact solutions are found for plane waves incident on piston-like absorbers in the form of long strips or circles.

621.395.623.7 : 534.32 **3024**

Contribution to the Subjective Appreciation of Music Reproduction by means of Loudspeaker Combinations.—G. Kaufmann. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 4th Dec. 1956, Vol. 8, Nos. 5/6, pp. 93–102.) A series of tests is described in which the quality of reproduction of four pieces of music by three different loudspeaker combinations was judged by a number of listeners in two rooms of differing reverberation characteristics. The results are analysed in detail.

621.395.623.741 : 621.375.13 **3025**

Effect of a Negative-Impedance Source on Loudspeaker Performance.—R. E. Werner. (*J. acoust. Soc. Amer.*, March 1957, Vol. 29, No. 3, pp. 335–340.) Amplifiers with negative-impedance outputs can greatly improve the l.f. and transient response of moving-coil loudspeakers.

621.395.623.8 : 621.396.712.2 **3026**

New High-Grade Monitoring Equipment for [studio] **Control Rooms.**—Enkel. (See 3298.)

621.395.625.3 : 621.317.76 **3027**

Pulse Method of Measuring Magnetic-Tape Speed Fluctuations.—A. G. Al'mukhamedov. (*Akust. Zh.*, Jan.–March 1957, Vol. 3, No. 1, pp. 19–22.) An oscillographic method is described in which pulses recorded at a given constant repetition rate by the recording head are compared with the pulses arriving at the pickup head. The repetition rate is adjusted so that the interval between pulses equals the average time of travel of the tape from the recording to the pickup head.

621.395.625.3 : 621.397.6 **3028**

Picture - Synchronized Magnetic Sound Recording in Television.—Vollmer. (See 3314.)

AERIALS AND TRANSMISSION LINES

621.315.212 : 621.372.51 : 621.372.8 **3029**

Simple Technique for Diplexing 10 000 Mc/s and Video Signals on Coaxial Cables.—M. C. Thompson, Jr, & D. M. Waters. (*Rev. sci. Instrum.*, March 1957, Vol. 28, No. 3, pp. 206–207.) Constructional details of a diplexer used in an airborne refractometer.

621.372.2 **3030**

Electromagnetic Surface Waves, Guided by a Boundary with Small Curvature.—M. A. Miller & V. I. Talanov. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2755–2765.) The theory developed indicates the existence of (*a*) three-dimensional surface waves which maintain their surface characteristics independent of the radius of curvature, and (*b*) two-dimensional waves which are partly radiated even at small curvatures of the boundary. Curves showing the decrease of the attenuation of the latter type of wave with the increase in wave retardation and with the decrease in the curvature of the guiding surface are given. Several surface waveguides are illustrated.

621.372.2 **3031**

The Surface Wave on a Dielectric Cylinder: the Method of Launching it and its Characteristics.—C. Jauquet, (*Rev. HF, Brussels*, 1956, Vol. 3, No. 8. pp. 283–296.) The radiations produced by a magnetic current flowing in a ring concentric with a loss-free dielectric cylinder of infinite length are considered.

The ring forms an ideal coaxial electromagnetic horn and the cylinder behaves as a waveguide for surface waves. The calculated characteristics are verified experimentally. The use of a dielectric cylinder for transmitting electromagnetic power is investigated; the losses of such a system have been measured.

621.372.2 **3032**

Propagation of Microwaves on a Single Wire: Part 2.—S. N. Contractor & S. K. Chatterjee. (*J. Indian Inst. Sci.*, Section B, Jan. 1957, Vol. 39, No. 1, pp. 52–67.) An experimental investigation of the matching of the surface wave, at 3·2 cm λ, with three different types of termination. Part 1: 16 of 1956 (Chatterjee & Madhavan).

621.372.2.029.6 **3033**

Surface-Wave Propagation along Coated Wires.—T. Berceli. (*Acta tech. Acad. Sci. hungaricae*, 1957, Vol. 17, Nos. 3/4, pp. 219–251. In English.) Propagation along a wire with a coating possessing both dielectric and magnetic properties is considered theoretically and the application of the formulae derived is illustrated by the design of a surface-wave transmission line for the 2·5–3·5 kMc/s band.

621.372.22 : 621.375.121.2 **3034**

Artificial Transmission Lines.—Hudson. (See 3079.)

621.372.22 **3035**

Wave Propagation in a Disk Line.—G. Piefke. (*Arch. elekt. Übertragung*, Feb. 1957, Vol. 11, No. 2, pp. 49–59.) Theoretical investigation of wave propagation in a waveguide consisting of alternate copper disks of thickness D_2 with a central hole and lossy dielectric disks of thickness D_1, with $D_1 + D_2 \ll \lambda$. A formula is derived for the propagation constants of all possible modes. The attenuation of H_{0n} modes exceeds that of the homogeneous waveguide by the factor $(1 + D_1/D_2)^{\frac{1}{2}}$; the attenuation of all other modes is much greater.

621.372.8 **3036**

Variational Calculation Method for Waveguides with Periodic Inhomogeneities: Part 1.—Sh. E. Tsimring. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol 2, No. 1, pp. 3–14.) A mathematical analysis is presented of the propagation of electromagnetic waves in a corrugated waveguide, with perfectly conducting walls, filled with a loss-free medium. The method developed is applied to an approximate calculation of the dispersion equation of a system comprising two parallel surfaces one of which has rectangular corrugations of arbitrary dimensions.

621.372.8 **3037**

Propagation of the Circular H_{01} Low-Loss Wave Mode around Bends in Tubular Metal Waveguide.—H. E. M. Barlow. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 403–409.) Theoretical analysis shows that the required field distribution at a bend is obtained if the wave-front is radial with respect to the centre of curvature. This may be achieved (*a*) by a suitable variation of the permittivity over the cross-section of the dielectric medium, or (*b*) by varying the surface reactance of the guide around its circumference.

621.372.8 **3038**

Propagation of Electromagnetic Waves in a Waveguide, Partly Filled with a Dielectric, with a Helix.—V. P Shestopalov. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2749–2754.) The case is investigated theoretically of the effect of a monoenergetic electron beam of circular cross-section moving inside the helix and parallel to the waveguide axis. The dielectric layer covers the internal surface of the waveguide.

621.372.8 **3039**

Directional Coupler for the H_{01} Wave in a Waveguide with Circular Cross-Section.—M. V. Persikov. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 65–74.) Results of the analysis presented give the conditions for the separation of the H_{01} wave from a mixture of other modes, and formulae are derived for the coefficients of coupling, directivity and attenuation of unwanted modes. Theoretical calculations have been verified experimentally in the 8·6–9·6-kMc/s band.

621.372.8.029.65.002.2 : 621.357.6 **3040**

Techniques for Electroforming of Precision Waveguide Components in the Millimetre Wavelengths.—A. A. Feldmann. (*Rev. sci. Instrum.*, April 1957, Vol. 28, No. 4, pp. 295–296.)

621.396.67.014.1 **3041**

Integral Equation for Currents in the Theory of Metallic Aerials.—A. V. Gaponov & M. A. Miller. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2766–2770.) In the presence of a load in the aerial, or when the finite conductivity of the metal is taken into account, the tangential components of the electric field (E_T) and magnetic field (H_T) in the closed surface Σ are connected by the relation $E_T = f(H_T)$; this relation is taken into account in a discussion of the problem considered by King (2196 of 1955).

621.396.67.029.62 : 621.397.7 **3042**

The Crystal Palace Television Transmitting Station.—McLean, Thomas & Rowden. (See 3318.)

621.396.677 : 621.396.11 : 551.510.535 **3043**

The Gain of a Directive Receiving Aerial for Short-Wave Back-Scatter.—Beckmann & Vogt. (See 3276.)

621.396.677.32 **3044**

Directivity of End-Fire Arrays of Isotropic Antennas.—J. R. Lignon. (*Ann. Acad. bras. Sci.*, 31st Dec. 1956, Vol. 28, No. 4, pp. 439–446. In English.) Expressions are derived for the theoretical gain under conditions of normal and of increased directivity. The optimum number of elements for maximum gain for a given length of array can then be calculated.

621.396.677.75 **3045**

The Shape of Dielectric Beam Aerials.—G. von Trentini. (*Nachrichtentech. Z.*, Feb. 1957, Vol. 10, No. 2, pp. 60–64.) Report of experimental investigations made at 3·25 cm λ on various types of radiator consisting of dielectric rods and plates. The dimensions of the models used are given and their relative advantages are discussed. 20 references.

621.372 **3046**

Randwertprobleme der Mikrowellenphysik. [Book Review]—F. H. Borgnis & C. H. Papas. Publishers: Springer-Verlag, Berlin-Göttingen-Heidelberg, 1955, 266 pp., DM 48. (*Naturwissenschaften*, March 1957, Vol. 44, No. 6, p. 188.) The book provides an introduction, illustrated by examples, to methods of solving microwave field problems arising in the vicinity of aerials and the interior of waveguides or resonators. Solutions are obtained in the form of integrals.

AUTOMATIC COMPUTERS

681.142 **3047**

Digital Printer boosts Readout Time.—H. W. Gettings. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 182–185.) The system can print 180 lines of 12 characters per line in 1 sec.

681.142 : 511 **3048**

A System of Mathematical Symbols Suitable for Calculations by [digital] **Computers.**—L. V. Kantorovich. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1957, Vol. 113, No. 4, pp. 738–741. In Russian.)

681.142 : 621.318.57 : 621.3.042 **3049**

The Reading of Magnetic Stores without Loss of Information.—A. Darré. (*Frequenz*, Jan. & Feb. 1957, Vol. 11, Nos. 1 & 2, pp. 19–27 & 38–42.) A description and comparison of storage and read-out methods using various forms of the 'transfluxor' [see 3509 of 1955 (Rajchman & Lo)] and of toroidal core devices.

681.142 : 621.383.2 **3050**

An Electromagnetic Simulator for a Hill's Equation. — J. Valat. (*C. R. Acad. Sci., Paris*, 13th May 1957, Vol. 244, No. 19, pp. 2462–2465.) Description of a system based on a mirror galvanometer and twin-anode photocell with application to the calculation of cosmotron orbits. See also 3504 of 1955 (Blet).

681.142 **3051**

Analog Computer Techniques. [Book Review]—C. L. Johnson. Publishers: McGraw-Hill, London, 1956, 264 pp., 45s. (*Nature, Lond.*, 25th May 1957, Vol. 179, No. 4569, p. 1042.)

681.142 **3052**

Electronic Analog Computers. [Book Review]—G. A. Korn & T. M. Korn. Publishers: McGraw-Hill, London, 2nd edn, 1956, 452 pp., 56s. 6d. (*Nature, Lond.*, 25th May 1957, Vol. 179, No. 4569, p. 1042.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.016.35 **3053**

A Stability Criterion in the Form of an Integral Equation.—I. Gumowski. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2004–2007.) Wallman's criterion of realizability (see e.g. 1674 of June) can be interpreted as a stability criterion more general than those of Bode and Nyquist. Examples of its use are given.

621.319.4 : 621.315.614.6 **3054**

Change of Capacitance of a Stack of Capacitor Paper under Compression.—I. D. Fainerman & L. M. Vaisman. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2493–2497.) Results of an experimental investigation are reported. The optimum standard pressure for capacitance measurements is 0·5 kg.cm^{-2} or higher.

621.372.4 **3055**

Equivalence Theorem for Two-Terminal Networks with all Three Types of Impedance *R*, *C* and *L*.—K. H. R. Weber. (*Hochfrequenztech. u. Elektroakust.*, Jan. 1957, Vol. 65, No. 4, pp. 126–129.) The equivalence theorem formulated is based on earlier work [see 961 of 1955 (Weber & Schlegel)] and is applicable to series-parallel two-pole networks comprising R, C and L elements.

621.372.4 : 537.313 **3056**

The Calculation of Current Distribution in a Linear Network.—J. Vratsanos. (*Arch. elekt. Übertragung*, Feb. 1957, Vol. 11, No. 2, pp. 76–80.) It is proved that if a current I is flowing in a linear network of impedance R between its two terminals the current i flowing in any branch of impedance r is given by $i^2 = I^2 \partial R/\partial r$.

621.372.4 : 621.316.86 **3057**

Analysis of Electric Circuits containing Nonlinear Resistance.—L. A. Pipes. (*J. Franklin Inst.*, Jan. 1957, Vol. 263, No. 1, pp. 47–55.) Two methods of dealing with varistor circuits are illustrated : the 'reversion method' for the solution of nonlinear differential equations (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 202–207) is applied to determine transient response ; steady-state analysis is effected using a method of harmonic balance.

621.372.43 + 621.372.54].012 **3058**

Application of the Smith Chart to General Impedance Transformations.—H. N. Dawirs. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 954–956.) The use of the chart is extended for unsymmetrical networks having complex image impedances and propagation constant, e.g. filters in the rejection bands.

621.372.5 **3059**

Differentiating and Integrating Circuits.—G. Fodor & G. Temes. (*Acta tech. Acad. Sci. hungaricae*, 1957, Vol. 16, Nos. 1/2, pp. 73–103. In English.)

621.372.5 **3060**

Exact Electronic Differentiation.—H. Wittke. (*Elektronische Rundschau*, Jan. 1957, Vol. 11, No. 1, p. 7.) The addition of a feedback amplifier to the basic differentiating *RC* quadripole eliminates inherent phase and amplitude errors. See also 2574 of 1955.

621.372.5 : 621.318.134 **3061**

Ferrites at Microwaves.—P. E. V. Allin. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 292–296.) A review of the principles and applications of nonreciprocal microwave ferrite devices.

621.372.54 + 621.317.3] : 621.396.663 **3062**

Possible Applications of Goniometers in Telecommunications.—Frioke. (See 3228.)

621.372.543.2 **3063**

Design of Filters with Two Coupled Resonators for Wide Relative Pass Bands.—G. B. Stracca. (*Alta Frequenza*, Feb. 1957, Vol. 26, No. 1, pp. 41–89.) Design formulae and curves for Butterworth- and Tchebycheff-type inductively-coupled double-tuned filters are given. See also 2368 of August (Carassa).

621.372.56.029.6 : 621.372.8 : 621.318.134 **3064**

Nonreciprocal Ferrite Attenuator in a Rectangular Waveguide.—M. Vadnjal. (*Alta Frequenza*, Feb. 1957, Vol. 26, No. 1, pp. 3–24.) The increase in reverse loss relative to forward loss caused by a layer of dielectric material adjacent to the ferrite slab is investigated [see also 3675 of 1956 (Weisbaum & Seidel)]. Calculated results agree with experimental findings.

621.373 : 621.374.42 **3065**

Mutual Synchronization of Oscillators at Multiple Frequencies.—G. M. Utkin. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 44–56.) The dependence of the mode of operation of the system, the oscillation frequencies and the synchronization zone width on the frequency ratio of the two oscillators and oscillator parameters is calculated. Beat-frequency operation is also discussed.

621.373.43 : 513.83 **3066**

Study of a Nonlinear Oscillator by Topological Analysis.—P. Dagnelie. (*Rev. HF, Brussels*, 1956, Vol. 3, No. 8, pp. 275–282.) The oscillator analysed is of the universal type capable of producing quasi-sinusoidal, and continuous or discontinuous relaxation oscillations according to the adjustment of two parameters (resistances). The discontinuous solutions of the integral equations of the oscillator are discussed and the theory is verified by experimental measurements in the form of oscillograms.

621.373.431.1 : 621.387.4 **3067**

Experimental Investigations of the Dead Time of Univibrators.—D. Kiss & J. Szivek. (*J. sci. Instrum.*, March 1957, Vol. 34, No. 3, pp. 99–100.) The earliest time in which a monostable multivibrator can be triggered after a previous triggering is found to be strongly dependent on the amplitude of the triggering pulse.

621.373.52 **3068**

Transistor *RC* Oscillators.—M. K. Achuthan. (*Electronic Radio Engr*, Aug. 1957, Vol. 34, No. 8, pp. 309–310.)

621.374.32 : 621.314.7 **3069**

Decade Counters using Junction Transistors.—L. P. Morgan & W. L. Stephenson. (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 2–10.) A practical design with a counting speed of 100 kc/s uses one binary and four asymmetrical bistable circuits.

621.374.32 : 621.314.7 **3070**

High-Reliability Transistorized Counter.—H. C. Chisholm. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 171–173.) "Cascaded silicon-junction transistor binary stages energize neon-lamp indicators for digital frequency meter at counting rates up to 100 000 per second."

621.374.32 : 621.387.4 **3071**

A Soft-Valve Scaler for Intermittent Fast Counting.—G. W. Hutchinson. (*J. sci. Instrum.*, March 1957, Vol. 34, No. 3, pp. 109–110.) Counting information is stored as charge on a capacitor and is subsequently read off as evenly spaced pulses by discharging the capacitor with a standard repetitive waveform.

621.374.4 **3072**

Double-Lock Synchronizing Method for Frequency Division.—J. B. Earnshaw. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 282–283.) Both the maximum and the minimum of the timing waveform of a multivibrator are locked to an incoming signal.

621.374.4.029.63/64 **3073**

Microwave Mixing and Frequency Dividing.—R. W. Degrasse & G. Wade. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 1013–1015.) Use of the nonlinearity of an overmodulated electron beam in a travelling-wave valve.

621.375.024 : 538.639 **3074**

D.C. Amplifier with Converter based on the Effect of Change of Resistance of Semiconductors in a Magnetic Field.—V. N. Bogomolov. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2480–2486.) Theoretical and practical details are given of a d.c./a.c. converter (modulater) based on the $\Delta\rho/\rho$ effect in InSb, a Hall-effect mixer, and a Hall-effect demodulator.

621.375.029.422/.424 **3075**

Selective Amplification at Sub-audio Frequencies.—F. J. Hyde. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 260–265.) Direct transmission systems discussed are the resonant galvanometer and double *CR* (low-pass, high-pass) filter. Five feedback systems considered are based respectively on the twin-T filter, the Wien bridge, the double *CR* network, the zero-phase-shift oscillator and a phase-inversion method due to Schneider (1207 of 1946).

621.375.029.63/.64 : 538.221 : 538.569.4 **3076**

Proposal for a Ferromagnetic Amplifier in the Microwave Range.—H. Suhl.

(*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 384–385.) Discussion of a system based on the anomalous absorption effects in ferromagnetic insulators at high signal powers.

621.375.029.64 : 538.569.4 : 621.396.822 3077

Noise in a Molecular Amplifier.—M. W. Muller. (*Phys. Rev.*, 1st April 1957, Vol. 106, No. 1, pp. 8–12.) Extension of theory to systems not in thermal equilibrium.

621.375.121.1 3078

The Design of Phase-Linear Intermediate-Frequency Amplifiers.—A. van Weel. (*J. Brit. Instn Radio Engrs*, May 1957, Vol. 17, No. 5, pp. 275–286.) Both stagger-tuned single circuits and combinations of band-pass filters are discussed.

621.375.121.2 : 621.372.22 3079

Artificial Transmission Lines.—A. C. Hudson. (*Electronic Radio Engr*, Aug. 1957, Vol. 34, No. 8, pp. 297–299.) Relations are derived between coil dimensions and constants for lines of the negative-mutual-inductance type.

621.375.121.2 : 621.372.6 3080

Amplifier with Distributed Constants as a System of Multipoles.—Yu. N. Prozorovski. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 57–64.) Analysis is presented of an amplifier with distributed parameters considered as a system of a finite number of multipoles. Matrices connecting the input and output voltages and currents are derived and the transmission coefficient of the amplifier is determined.

621.375.132.3 3081

Multivalve Cathode - Follower Circuits.—J. G. Thomason. (*Wireless World*, July & Aug. 1957, Vol. 63, Nos. 7 & 8, pp. 310–313 & 373–377.) Methods of approaching the ideal buffer-stage performance are discussed and practical feedback circuits and their application are described. Examples include a precision differential amplifier.

621.375.2 3082

Alternatives to Cathode Bias for Vacuum Tubes.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 1011–1012.) Resistive feedback stabilizes negative grid operating potential.

621.375.4.039.3 : 621.314.7 3083

A 200-mW Amplifier employing Transistors Operated from a 6-V Supply.—O. J. Edwards. (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 32–33.) Modification of the circuit described earlier (3324 of 1956) operating on 4·5 V.

621.376.222.029.5 3084

A 16-kc/s Amplitude Modulator employing Envelope Feedback.—K. G. Corless. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 287–290.) Using a greater degree of feedback, modulation linearity within 1% is achieved for any frequency component in the range 0–30 c/s.

621.376.23 : 621.396.822 3085

Effect of Differentiation and Integration of Fluctuations on the Mean Number of Peaks.—V. I. Tikhonov. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 23–27.) An analysis is presented.

621.376.23 : 621.396.822 3086

Transformation of Amplitude and Phase Fluctuations of Oscillations by Tuned Systems.—G. S. Gorelik & G. A. Elkin. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 28–33.) A theoretical investigation is reported of the transmission of signals with randomly varying phase and amplitude by systems comprising a linear (tuned) and a nonlinear (detector) unit. Formulae are derived relating the statistical characteristics of phase and amplitude at the input of a tuned system with those at the output.

621.376.5 : 621.3.016.352 3087

Stability of Linear Pulse Systems with Variable Parameters.—G. P. Tartakovski. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 15–22.) The stability conditions for pulse systems with constant parameters are extended to systems with variable parameters and an equation is derived for the transmission function in pulse systems with variable parameters and feedback. The application of results is illustrated in a discussion of a p.f.m. signal in a variable-parameter system with pulse feedback.

GENERAL PHYSICS

535.566 : 539.23 3088

Calculation of the Amplitude of a Plane Wave Reflected by a Homogeneous Metal Lamina with Parallel Faces.—P. Dumontet. (*C. R. Acad. Sci., Paris*, 24th April 1957, Vol. 244, No. 17, pp. 2234–2236.) The work of previous authors [e.g. 1354 of 1949 (Reuter & Sondheimer)] is generalized to allow for finite thickness of film. Calculated values of complex wave amplitude for light reflected from metals such as Ag, Au, Al, whatever the film thickness, are in agreement with classical theory.

537.226 3089

Dipole Moment Fluctuations of a Dielectric Body.—B. K. P. Scaife. (*Proc. phys. Soc.*, 1st March 1957, Vol. 70, No. 447B, pp. 314–319.) The particular case of a dielectric which has two mechanisms of polarization (orientational and displacement) is considered. It is shown that Fröhlich's free-energy method of computing the fluctuations is independent of the dynamic properties of the dielectric.

537.226 3090

Theory of Dielectric Relaxation for the Three-Dimensional Polar Rotator: Lattice Models Leading to Bimodal Loss Curves.—J. D. Hoffman & B. M. Axilrod. (*J. Res. nat. Bur. Stand.*, Feb. 1957, Vol. 58, No. 2, pp. 61–73.)

537.226.33 3091

Relaxation Processes and Inertial Effects: Part 1—Free Rotation about a Fixed Axis. Part 2—Free Rotation in Space.—R. A. Sack. (*Proc. phys. Soc.*, 1st April 1957, Vol. 70, No. 448B, pp. 402–413 & 414–426.) The dielectric properties of systems of rigid dipoles are determined.

537.311.62 3092

Surface [skin] **Effect in a System of Conductors with Rectangular Cross-Sections.**—L. A. Tseitlin. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2771–2777.) Analysis is presented for (*a*) very thin bars and (*b*) bars with finite thickness.

537.52 : 621.385.13 3093

Filiform Anode in a Gas Discharge.—B. N. Klyarfeld & A. A. Frid. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2541–2547.) An experimental investigation is reported of a Hg-vapour discharge ($p = 10^{-3}$ mm Hg) in a long narrow cylindrical tube containing a cylindrical anode at one end, a hot oxide cathode surrounded by another cylindrical anode in the bulbous other end of the tube, and a filiform molybdenum electrode along the axis of the tube.

537.525.029.64 3094

Microwave Gas Discharge Breakdown in Air, Nitrogen, and Oxygen.—D. J. Rose & S. C. Brown. (*J. appl. Phys.*, May 1951, Vol. 28, No. 5, pp. 561–563.) The breakdown field for pure air, without contamination by oxides of nitrogen, is higher than previously observed and lies between the fields for nitrogen and oxygen. The results are now in accordance with theory.

537.533 + 535.215 3095

Electron Emission from Complex Surfaces.—P. V. Timofeev. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 85–91.) Photoelectric secondary and field emission from complex cathodes is discussed. Conclusions indicate that (*a*) the surface electron energy levels play an important part in electron emission, (*b*) the photoelectric emission from Cs photocathodes is largely determined by the distribution of free Cs in the surface layer, (*c*) high values of secondary electron emission are always observed in cases when particles with low conductivity are present in the surface layer, and (*d*) field and secondary emission depend on the presence of positive charges in the surface layer.

537.533 3096

The Electron Emission from Mechanically Worked Metal Surfaces under Oxygen Pressures Variable with Time.—J. Lohff. (*Naturwissenschaften*, April 1957, Vol. 44, No. 7, p. 228.) Curves have been plotted of electron emission from an aluminium surface which had been treated with a steel brush and subjected to varying pressures in an atmosphere of O_2. Further tests are necessary to resolve contradictions with theory based on earlier oxidation experiments (see 1723 of June).

537.533 3097

Electron Emission from Vapour-Deposited Metal Films.—J. Wüstenhagen. (*Naturwissenschaften*, April 1957, Vol. 44, No. 7, pp. 228–229.) Emission/time curves for aluminium films in an atmosphere of oxygen show a decrease of emission with decreasing O_2 pressures; a temporary evacuation of O_2 although stopping emission does not affect the continued decay of the emission after the O_2 pressure has been restored (see also 3096 above). Tests were also made on Be and Al films in the presence of inert gases.

537.533 : 537.226 3098

Mechanism of Electron Emission from Thin Dielectric Layers under the Influence of a Strong Electric Field (Malter Effect).—M. I. Elinson & D. V. Zernov. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 75–84.) Discussion of the Malter effect (*Phys. Rev.*, 1st July 1936, Vol. 50, No. 1, pp. 48–58.)

537.56 3099

On Bohm-Pines' Theory of Plasma.—W. Schützer. (*Ann. Acad. bras. Sci.*, 31st Dec. 1956, Vol. 28, No. 4, pp. 419–422. In English.) The method of treatment of electron interaction in a dense electron gas developed by Bohm & Pines (1375 of 1954 and back references) is modified so that media of finite extent can be dealt with.

537.56 3100

Derivation of the Fokker-Planck Equation for Plasma.—S. V. Temko. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1021–1026.)

537.56 3101

Hydrodynamic Description of Plasma Oscillations.—B. B. Kadomtsev. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1083–1084.) Brief discussion restricted to electron oscillations assuming ions and molecules to be of infinitely large mass; the amplitude of oscillations is assumed to be small so that the electron velocity distribution function is nearly Maxwellian.

537.56 : 538.6 3102

Stability of Plasma in a Strong Magnetic Field.—Yu. A. Tserkovnikov. (*Zh. eksp. teor. Fiz.*, Jan. 1957, Vol. 32, No. 1, pp. 67–74.)

537.56 : 538.63 3103

The Lateral Diffusion of an Electron Swarm in a Magnetic Field.—R. J. Bickerton. (*Proc. phys. Soc.*, 1st March 1957, Vol. 70, No. 447B, pp. 305–313.) The diffusion of electrons, moving through a gas under the influence of parallel electric and magnetic fields has been measured. Results are presented for the gases He and H_2 and currents between 10^{-10} and 10^{-5} A.

537.56 : 538.63 3104

On a Generalized Theory of a Lorentz Plasma under a Nonperiodic Force and taking Account of Inelastic Collisions.—R. Jancel & T. Kahan. (*C. R. Acad. Sci., Paris*, 20th May 1957, Vol. 244, No. 21, pp. 2583–2586.) First- and second-order approximations are made for a solution of Boltzmann's transport equation by the method described earlier (2420 of August).

537.56.083 : 621.372.413.029.63/.64 3105

Limitations of the Microwave Cavity Method of Measuring Electron Densities in a Plasma.—K. B. Persson. (*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 191–195.)

537.56.083 : 621.372.413.029.63/.64 3106

Microwave Measurements of High Electron Densities.—S. J. Buchsbaum & S. C. Brown. (*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 196–199.)

538.112.083.2 3107

Determination of the Value of the Bohr Magneton by a Method of Resonance in Ionized Air.—S. Procopiu & C. Papusoi. (*C. R. Acad. Sci., Paris*, 1st April 1957, Vol. 244, No. 14, pp. 1905–1907.)

538.221 3108

On the Propagation of Waves of the Classical Magnetic Form in a Fluid of Magnetic Doublets.—G. Karpman & J. P. Vigier. (*C. R. Acad. Sci., Paris*, 20th May 1957, Vol. 244, No. 21, pp. 2577–2580.) Application of theory proposed earlier [2421 of August (Karpman)].

538.23 : 538.221 3109

Interpretation of the Creep of Hysteresis Cycles.—L. Néel. (*C. R. Acad. Sci., Paris*, 27th May 1957, Vol. 244, No. 22, pp. 2668–2674.) If a magnetizing field varies cyclically about a mean value which is not zero successive cycles are displaced and creeping is said to occur. Principal experimental results can be interpreted by supposing that successive hysteresis cycles described between the same limits and identical macroscopically, differ microscopically and give rise to random coupling fields between elementary domains. See also 3110 below.

538.24 : 538.221 3110

Action of Successive Magnetic Fields of a Random Character on the Magnetization of Ferromagnetic Substances.—L. Néel. (*C. R. Acad. Sci., Paris*, 13th May 1957, Vol. 244, No. 19, pp. 2441–2446.) The idea of irreversible susceptibility is recalled and the variation of magnetization produced by n successive applications of a magnetic field of random character is calculated. The results of this research are applied to thermal fluctuations.

538.3 3111

Interaction between a Moving Current-Carrying Filament and a Conducting Wall.—A. I. Morozov. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1079–1080.) The forces acting on the filament, which is moving with a velocity $v \ll c$ parallel to the plane surface of the conducting medium are calculated.

538.311 3112

Note on Production of Strong Magnetic Fields of Short Duration and Measurement of their Intensity.—A. Piekara, J. Małecki, M. Surma & J. Gibalewicz. (*Proc. phys. Soc.*, 1st April 1957, Vol. 70, No. 448B, pp. 432–434.)

538.311 : 621.317.441 3113

Compensation of the Earth's Magnetic Field.—G. G. Scott. (*Rev. sci. Instrum.*, April 1957, Vol. 28, No. 4, pp. 270–273.) A system for reducing and holding the magnetic fields in a working space to 0·01% of the earth's horizontal component.

538.566 : 535.42] + 534.26 3114

Diffraction by an Aperture.—J. B. Keller. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 426–444.) Diffraction by an aperture of any shape in a thin screen is treated by an extension of geometrical optics introducing diffracted rays produced when incident rays hit the aperture edge. Explicit formulae and numerical results are given for slits and circular apertures.

538.566 : 535.42] + 534.26 3115

Diffraction by an Aperture: Part 2.—J. B. Keller, R. M. Lewis & B. D. Seckler. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 570–579.) The method described in Part 1 (3114 above) is compared with the Kirchhoff method (Huyghen's principle) and modifications of this [see 2909 of 1954 (Bouwkamp)].

538.566 + 534.2] : 535.43 3116

A New Method of Calculating Scattering with Particular Reference to the Circular Disc.—D. S. Jones. (*Commun. pure appl. Math.*, Nov. 1956, Vol. 9, No. 4, pp. 713–746.) A comparatively simple physical picture of the low-frequency problem is developed. In the mathematical formulation the field is split into a nonradiating field which becomes the static field in the limit of zero frequency and the remainder of the field. By the use of a Fredholm integral equation the use of an iteration scheme is avoided. 16 references.

538.566 : 535.43 3117

Back-Scattering from Dielectric-Coated Infinite Cylindrical Obstacles.—C. C. H. Tang. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 628–633.) The back-scattering properties were investigated both theoretically (using eigenfunction expansion) and experimentally; very close agreement was obtained. A parallel-plate region enclosing a disk-shaped obstacle to represent a section across the theoretically infinite cylinder was used. The disk was rotated eccentrically and the back-scatter detected by the Doppler-shift principle.

538.569.4 : 538.2 : 538.54 3118

The Effect of Eddy Currents on Nuclear Magnetic Resonance in Metals.—A. C. Chapman, P. Rhodes & E. F. W. Seymour. (*Proc. phys. Soc.*, 1st April 1957, Vol. 70, No. 448B, pp. 345–360.) The power absorption from a r.f. magnetic field is calculated as a function of the off-resonance absorption and the complex susceptibility for a material in the form of a plate, cylinder or sphere. The results thus obtained are confirmed by measurements on aluminium foil.

538.569.4 : 538.221 : 621.375.029.63/.64 3119

Proposal for a Ferromagnetic Amplifier in the Microwave Range.—Suhl. (See 3076.)

538.569.4 : 621.375.029.64 : 621.396.822 3120

Noise in a Molecular Amplifier.—Muller. (See 3077.)

539.11 : 061.3 3121

Physics of the Solid State.—(*Nature, Lond.*, 18th May 1957, Vol. 179, No. 4568, pp. 1004–1005.) Brief report of a meeting of the Physical Society held at the University of Nottingham, April 1957. Some twenty papers were given, the majority referring to magnetism or semiconduction. No full report is being prepared.

548.0 : 539.11 3122

Theory of Excitons in Ionic Crystals.—I. P. Ipatova. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2786–2788.)

548.0 : 539.11 3123

The Interaction of Electrons with Lattice Vibrations.—P. G. Harper. (*Proc. phys. Soc.*, 1st April 1957, Vol. 70, No. 448B, pp. 390–392.) Equations are derived for the frequency and zero-point energy of normal modes of lattice motion.

548.0 : 539.11 3124

Field Mass of Polarizing Excitons in Ionic Crystals.—I. M. Dykman, E. I. Kaplunova & K. B. Tolpygo. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2459–2466.) The effective mass is calculated assuming that it possesses a field characteristic and depends on the interaction of the exciton with lattice vibrations. For excitons with a large radius the calculation is made in a macroscopic approximation, for those with a small radius the calculation is made with reference to NaCl and KCl crystals taking into account the discrete structure of the crystals and the dispersion of natural vibrations.

548.0 : 539.11 : 537.311.1 3125

Capture of Conduction Electrons by Charged Defects in Ionic Crystals.—Yu. E. Perlin. (*Zh. eksp. teor. Fiz.*, Jan. 1957, Vol. 32, No. 1, pp. 105–114.) The probability of capture is calculated as a function of polaron velocity and the temperature dependence of the lifetime of the current carriers is determined.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 : 621.396.822 3126

Cosmic Radio-Noise Intensities below 10 Mc/s.—G. R. Ellis. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 229–234.) The increase in intensity with decreasing frequency ceases near 10 Mc/s and the spectrum is flat from 10 Mc/s down to 2 or even 0·9 Mc/s. The flux density is about 2×10^{-19} W/cm² per c/s.

529.77 (470) 3127

New Boundaries of Time Zones in the U.S.S.R.—P. N. Dolgov. (*Priroda, Moscow*, Jan. 1957, No. 1, pp. 57–61.) Historical introduction and map of new standard-time zones which came into effect on 1st March 1957 at midnight Moscow time.

55 3128

I.G.Y. World Warning Agency.—(*Tech. News Bull. nat. Bur. Stand.*, May 1957, Vol. 41, No. 5, pp. 65–66.)

550.385 : 523.74 3129

The Geomagnetic Influence of the Northern and Southern Solar Hemispheres.—W. Gleissberg. (*Naturwissenschaften*, March 1957, Vol. 44, No. 6, pp. 176–177.) An analysis of solar data for the period 1889 to 1943 shows that the northern hemisphere of the sun has a predominant effect on geomagnetic disturbances. This is probably due to more intense particle emission from that part of the sun.

551.51 3130

The Theory of Molecular Diffusion in the Atmosphere.—P. Mange. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 279–296.) Existing theory is extended to include a linear variation in scale height and other refinements. The application to particular problems is discussed.

551.51 3131

On the Atmospheric Dynamo Theory.—M. L. White. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 329–330.) Methods of avoiding earlier simplifying assumptions are suggested.

551.510.535 3132

Ionospheric Problems.—T. W. Bennington. (*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 365–368.) The future application of I.G.Y. data may help radio communication by providing information on sporadic E and night-time E layers and on ionospheric disturbances.

551.510.535 3133

Physical Properties of the Atmosphere from 90 to 300 km.—G. J. Simmons: H. K. Kallman, W. B. White & H. E. Newell, Jr. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 327–328.) Comment on 436 of February, giving a preferred formula for equation 16. See also *ibid*, March 1957, Vol. 62, No. 1, p. 168, for correction to original paper.

550.510.535 3134

Effective Recombination Coefficient in the Ionosphere.—V. I. Krasovski. (*Bull Acad. Sci. U.R.S.S., sér. géophys.*, April 1957, No. 4, pp. 504–511. In Russian.) Discussion of theoretical and experimental results. Conclusions are: (*a*) the coefficient depends on the electron concentration and rate of ion production and therefore cannot be assumed to be constant in time or symmetrical relative to noon, sunrise and sunset; (*b*) the coefficient depends on the nature of the primary ions and therefore is not necessarily identical with the coefficient for photoionization when ionization is produced by corpuscles, meteorites or gas discharges. The presence of infrasonic waves would cause an increase in the coefficient.

551.510.535 3135

Recombination Coefficient and Electron Production Rate from Total Electron Content in Unit Column below the Level of Maximum Ionization.—S. Datta. (*Indian J. Phys.*, Jan. 1957, Vol. 31, No. 1, pp. 43–52.)

551.510.535 3136

Drift Measurements of the E Layer.—L. Harang & K. Pedersen. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 183–198.) The Mitra method of recording drifts in the diffraction pattern was used at Kjeller (Norway) from June 1953 to September 1955. The diurnal and seasonal changes in drift are discussed and compared with results elsewhere. There is some evidence for a small lunar component.

551.510.535 3137

Anomalies in Ionosonde Records due to Travelling Ionospheric Disturbances.—G. H. Munro. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 325–326.)

551.510.535 : 550.385 3138

Relations among Radio Absorbing Regions, Geomagnetic Bay Disturbances and Slant Es in Auroral Latitudes.—S. Matsushita. (*J. Geomag. Geoelect.*, Dec. 1956, Vol. 8, No. 4, pp. 156–160.)

551.510.535 : 550.385 3139

Disturbances in the Ionospheric F_2 Region associated with Geomagnetic Storms: Part 1—Equatorial Zone.—T. Sato. (*J. Geomag. Geoelect.*, Dec. 1956, Vol. 8, No. 4, pp. 129–135.) The ionospheric disturbances are explained in terms of the vertical drift of electrons caused by the electric field arising from the disturbed component of the geomagnetic field. The ionospheric data for Huancayo during several disturbances are in accordance with the magnetic data.

551.510.535 : 621.396.812.3 3140

Fading and Random Motion of Ionospheric Irregularities.—Mitra & Srivastava. (See 3283.)

551.594.5 3141

The Geometry of Auroral Ionization.—T. R. Kaiser. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 297–298.) The pattern of radio echoes from auroral ionization agrees with the hypothesis that the ionization occurs in 'blobs' distributed along an arc which follows a parallel of magnetic latitude and not in narrow columns aligned along the magnetic lines of force.

551.594.5 **3142**

Radar Reflections from Aurorae.—Ya. G. Birfeld. (*Bull. Acad. Sci. U.R.S.S., sér. géophys.*, April 1957, No. 4, pp. 543–547. In Russian.) Report on experimental results obtained at the Loparskaya station near Murmansk. A frequency of 72 Mc/s was used and echoes were recorded from aurorae up to 1 000 km distant. Typical c.r.o. traces are shown and are briefly described.

551.594.6 **3143**

Lightning Mechanism and Atmospherics Radiation.—H. Isikawa. (*J. Geomag. Geoelect.*, Dec. 1956, Vol. 8, No. 4, pp. 136–146.) A discussion of the discharge processes as deduced from electric-field measurements. Particular attention is paid to the leader discharges.

551.594.6 : 523.75 **3144**

An Observation of Audio-Frequency Electromagnetic Noise during a Period of Solar Disturbance.—J. M. Watts. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 199–206.) An analysis of hiss recorded during a magnetic storm shows (*a*) a peak in the frequency spectrum near 3 kc/s, (*b*) that the high-frequency limit was variable, (*c*) a procession of narrow-bandwidth tones gliding upwards occurred during part of the period.

LOCATION AND AIDS TO NAVIGATION

621.396.933 **3145**

Vortac Beacons for Rho-Theta Navigation.—P. Caporale. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 156–159.) A short-distance navigation system providing information on azimuth and distance when challenged by beacon equipment in approaching aircraft.

621.396.933.1 **3146**

Doppler Navaid for Civil Aircraft.—(*Wireless World*, Aug. 1957, Vol. 63, No. 8, p. 396.) Brief description of a self-contained air navigational system entirely independent of ground station cooperation.

621.396.933.2 **3147**

Applying the Doppler Effect to Direction-Finder Design.—J. A. Fantoni & R. C. Benoit, Jr. (*Electronic Ind. Tele-Tech*, Jan. & Feb. 1957, Vol. 16, Nos. 1 & 2, pp. 75–77 . . 147 & 66–67 . . 128.) A circular array of fixed aerials is scanned in sequence to produce a direction-dependent phase. The i.f. from the receiver is applied to a discriminator and the resultant signal with scanning-switch modulation frequency of 42 c/s is applied to a c.r.-tube bearing display.

621.396.96.001.4 **3148**

Flight Simulator tests Fire-Control Radars.—D. L. DeMyer. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 168–170.) Target speeds of 200 to 1 000 m.p.h. in either direction can be simulated at distances of 800–24 000 yds.

621.396.963.3 : 621.396.822 **3149**

Detection of Pulse Signals in Noise: Trace-to-Trace Correlation in Visual Displays.—D. G. Tucker. (*J. Brit. Instn Radio Engrs*, June 1957, Vol. 17, No. 6, pp. 319–329.) Experiments show that side-by-side presentation of traces gives superior detection to that obtained by true integration. The theoretical reasons are discussed and previously published data are reviewed.

621.396.963.3 : 621.396.822 **3150**

Detection of Pulse Signals in Noise: the Effect on Visual Detection of the Area of the Signal Paint.—J. W. R. Griffiths. (*J. Brit. Instn Radio Engrs*, June 1957, Vol. 17, No. 6, pp. 330–338.) Available experimental data are reviewed in terms of physiological optics. Improvements in detection using side-by-side presentation may be explained by the increase in the solid angle subtended at the eye by the signal 'paint'.

621.396.969 **3151**

Rate-of-Climb Meter uses Doppler Radar.—S. H. Logue. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 150–152.) The direct-reading instrument uses a 10-kMc/s c.w. radar with the ground as reflecting surface. The equipment operates up to 2 300 ft.

621.396.969 : 061.3 **3152**

The International Annual Conference on Radio- and Sound-Location in Hamburg 1956.—H. Köppen. (*Nachr-Tech.*, Jan. 1957, Vol. 7, No. 1, pp. 34–39.) Summaries of most of the lectures delivered at this conference. See also *Elektronische Rundschau*, Dec. 1956, Vol. 10, No. 12, pp. 343–348 & Jan. 1957, Vol. 11, No. 1, pp. 27–29.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 **3153**

A New Electronic Circuit for a Hot-Cathode Ionization Gauge.—J. Schutten. (*Appl. sci. Res.*, 1957, Vol. B6, No. 4, pp. 276–284.) A circuit is described for measuring pressures in the range 10^{-3}–10^{-10} mm Hg. The quotient of ion current over emission current is measured, and at pressures below 10^{-4} mm Hg the meter indication varies by not more than 3% for emission currents between 10 μA and 1 mA.

535.21 : 546.23 **3154**

Structural Changes of Selenium under the Influence of Light.—H. Stegmann. (*Naturwissenschaften*, March 1957, Vol. 44, No. 5, pp. 108–109.) Experimentally observed changes in the transmission of light through Se films are briefly reported and discussed.

535.215 : 537.311.33 **3155**

The So-Called 'Secondary' and 'Through' Photocurrent in Semiconductors.—S. M. Ryvkin. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2439–2447.)

535.215 : [546.681.23 + 546.681.24 **3156**

Temperature Dependence of the Spectral Distribution of the Photoconductivity of Gallium Selenide and Telluride.—S. M. Ryvkin & R. Yu. Khansevarov. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2781–2783.) Results of an experimental investigation on polycrystalline GaSe and GaTe specimens with conductivities in the range 10^{-4}–$10^{-3}\,\Omega^{-1}.\text{cm}^{-1}$ are reported. Graphs showing the special characteristics at temperatures between about −100° C and + 100° C are given.

535.215 : 546.817.221 : 539.23 **3157**

Investigation of the Photoconductive Effect in Lead Sulphide Films using Hall and Resistivity Measurements.—J. F. Woods. (*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 235–240.) Measurements were made of the fractional change of Hall coefficient and of resistivity occurring under illumination of chemically deposited PbS films. Using a model of photoconductivity [1774 of June (Petritz)] the results show that the effect is due to an increase in majority-carrier density, with no modulation of barrier potentials.

535.37 : 534-8 **3158**

Luminescence Changes in Continuously Excited Zinc Sulphide Fluorescent Screens under the Influence of Ultrasonic Radiation.—M. Leistner & L. Herforth. (*Naturwissenschaften*, Feb. 1957, Vol. 44, No. 3, p. 59.) Brief preliminary note on tests carried out on ZnS-Cu and ZnS-Ag screens subjected to ultrasonic radiation at 800 kc/s and excited by ultraviolet light.

535.37 : 621.385.832 : 621.397.5 : 535.623 **3159**

Luminophores based on ZnS and ZnSe for Colour Television.—E. I. Blazhnova, A. I. Mokrintseva & V. I. Kas'yanova. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2784–2785.) Spectral intensity and brightness/electron-beam-density curves are shown of green, red and blue luminophores the compositions of which are given. The brightness decreased by less than 5% after 400 h at an anode potential of 15 kV and beam density of 1 μA.cm^{-2}. The effect of heat treatment in vacuum and in air on the colorimetric coordinates was also investigated and the results are tabulated.

537.226/.228.1 : 546.431.824-31 **3160**

Improved Barium Titanate Composition.—D. Schofield & R. F. Brown. (*J. acoust. Soc. Amer.*, March 1957, Vol. 29, No. 3, pp. 394–395.) Typical values of dielectric constant, radial coupling coefficient and piezoelectric constant are given for $BaTiO_3$ samples containing a small percentage of Co, the addition of which increases the power-handling capacity of the material.

537.226/.227 : 546.431.824-31 **3161**

Thermodynamic Theory of Ferroelectricity.—I. A. Izhak. (*Zh. eksp. teor. Fiz.*, Jan. 1957, Vol. 32, No. 1, pp. 160–162.) Results of measurements on polycrystalline $BaTiO_3$ in the paraelectric range confirm some predictions of thermodynamic theory. In the ferroelectric range satisfactory agreement was obtained only at temperatures not more than 10–12° C below the Curie point.

537.226/.227 : 546.431.824-31 **3162**

Theory of Phase Phenomena in Barium Titanate.—L. P. Kholodenko. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1034–1045.) The 'temperature hysteresis' in phase transitions is discussed and the temperature dependence of the permittivity near the phase transition points is calculated. The coefficients occurring in the thermodynamic theory are expressed in terms of measurable quantities.

537.226/.227 : 546.431.824-31 : 539.185 **3163**

Fast Neutron Effects in Tetragonal Barium Titanate.—M. C. Wittels & F. A. Sherrill. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 606–609.) An experimental study of the lattice properties of irradiated barium titanate. The neutron flux transformed the tetragonal form into the cubic form normally stable only above the Curie point. The control of ferroelectric properties by varied amounts of irradiation is suggested.

537.226 : 537.224 **3164**

New Inorganic Dielectric Electrets.—A. N. Gubkin & G. I. Skanavi. (*Zh. eksp. teor. Fiz.*, Jan. 1957, Vol. 32, No. 1, pp. 140–142.) The 14 different materials investigated include several titanates, steatite, porcelain, glass, quartz and KBr. For $BaTiO_3$ the surface charge density, σ, is $15 \cdot 4 \times 10^{-9}$ C/cm², 30 min after polarization in a field of 10 kV/cm, falling to $1 \cdot 1 \times 10^{-9}$ C/cm² after 7 months. $MgTiO_3$, with $\sigma = 1 \cdot 9 \times 10^{-9}$ C/cm² and $0 \cdot 5 \times 10^{-9}$ C/cm² after 7 months, was shown to have a 'lifetime' of over 1·5 years. The variations of σ with time are tabulated for periods up to 7 months after polarization.

537.226+537.311.33] : 537.29 **3165**

Influence of an Electric Field on the Properties of Thin Dielectric and Semiconductor Layers.—Yu. M. Volokobinski. (*C. R. Acad. Sci. U.R.S.S.*, 11th April 1957, Vol. 113, No. 5, pp. 1023–1024. In Russian.) The electrical conductivity and electrical strength of several oxides, sulphides and other compounds were investigated. Results indicate that the electrical breakdown is connected with the disruption of the layer by the heating produced by the passage of a current through the layer.

537.226 : 546.42.824-31 **3166**

Single Crystals of Strontium Titanate.—A. L. Khodakov, M. L. Sholokhovich, E. G. Fesenko & O. P. Kramarov. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2506–2507.) Brief note on the dielectric-constant/temperature characteristics of single crystals of $SrTiO_3$ prepared by two different methods.

537.226 : 621.315.612 **3167**

Classification of Perovskite and Other ABO_3-Type Compounds.—R. S. Roth. (*J. Res. nat. Bur. Stand.*, Feb. 1957, Vol. 58, No. 2, pp. 75–88.) A partial survey of the reactions occurring in binary oxide mixtures of the types $AO : BO_2$ and $A_2O_3 : B_2O_3$ has been conducted as part of a program of fundamental research on ceramic dielectrics.

537.226 : 621.315.612 **3168**

Breakdown of Ceramic Specimens under a High-Frequency Voltage at Different Temperatures.—I. E. Balygin. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2498–2505.) Various commercial-type ceramic materials were tested at a frequency of 1·5 Mc/s and the effects of porosity, chemical composition and structure of the material on the breakdown characteristics were noted.

537.226.2 : 546.331.31 **3169**

Influence of Dislocations of the Crystal Lattice on the Permittivity of Rock-Salt Crystals.—Yu. A. Sikorski. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2487–2492.)

537.227 **3170**

Phase Transitions in Ferroelectric Solid Solutions based on Strontium Pyrotantalate.—G. A. Smolenski, V. A. Isupov & A. I. Agranovskaya. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1957, Vol. 113, No. 4, pp. 803–805. In Russian.) The dielectric-constant/temperature characteristics were determined experimentally in solutions of $Sr_2Nb_2O_7$, $Ba_2Ta_2O_7$ and $Ca_2Ta_2O_7$ in $Sr_2Ta_2O_7$ at a frequency of 500 kc/s. Results are presented graphically.

537.227 : 546.431.824-31 **3171**

Solid Solutions of Barium Metaniobate and Metatantalate in Barium Titanate with Ferroelectric Properties.—G. A. Smolenski, V. A. Isupov & A. I. Agranovskaya. (*C. R. Acad. Sci. U.R.S.S.*, 11th April 1957, Vol. 113, No. 5, pp. 1053–1056. In Russian.) Results are reported of an experimental investigation of the dependence of the dielectric constant and loss angle on temperature in the range from −100° C to +200° C in solid solutions with various compositions. The measurements were made at a frequency of 1 kc/s. A composition/temperature phase diagram is also given.

537.227 : 547.476.3 **3172**

The Ferroelectric Domain Structure of Rochelle Salt.—E. Straubel-Fischer. (*Naturwissenschaften*, April 1957, Vol. 44, No. 7, pp. 230–231.) Brief report of optical tests.

537.228.1 : 548.0]. 001.4(083.7) **3173**

I.R.E. Standards on Piezoelectric Crystals—the Piezoelectric Vibrator: Definitions and Methods of Measurement, 1957.—(*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1010.) Correction to 1788 of June.

537.311.31 **3174**

Electrical Resistivity of Nickel-Palladium Alloys.—A. W. Overhauser & A. I. Schindler. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 544–546.) Experimental data for alloys of different relative concentrations are quoted. The results are discussed in terms of the electronic structure of the alloy.

537.311.33 **3175**

Evaporation of Impurities from Semiconductors.—K. Lehovec, K. Schoeni & R. Zuleeg. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 420–423.) The impurity distribution arising by evaporation from the surface is derived assuming the rate of evaporation to be proportional to the surface concentration. The proportionality constant is derived experimentally.

537.311.33 **3176**

A Contribution to the Recombination Statistics of Excess Carriers in Semiconductors.—P. T. Landsberg. (*Proc. phys. Soc.*, 1st March 1957, Vol. 70, No. 447B, pp. 282–296.) For inter-band transitions the conditions under which a recombination coefficient exists are investigated and the effect of degeneracy of hole and electron gas is discussed. For transitions involving traps, degeneracy and the case of impurities having several trapping levels are considered.

537.311.33 **3177**

Relation between Ratio of Diffusion Lengths of Minority Carriers and Ratio of Conductivities.—S. S. L. Chang. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 1019–1020.)

537.311.33 **3178**

Breakdown of Transition Layers in Semiconductors.—B. M. Vul. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2403–2416.) The electrical, thermal, and thermoelectric factors in the breakdown of a *p-n* junction are discussed theoretically.

537.311.33 **3179**

Measurement of Surface Recombination Velocity in a Thin Semiconductor with Qualitatively Differing Sides.—O. V. Sorokin. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2467–2472.) Formulae are derived from the solution of the continuity equation connecting the effective length of diffusion of non-equilibrium current carriers, L_α, in a plane semiconductor filament with the surface recombination velocities S_1 and S_2 of the upper and lower sides, respectively. The relations derived allow S_1 and S_2 to be calculated from the result of measurements of L_α.

537.311.33 **3180**

Measurement of the Lifetime, Diffusion Coefficient and Surface Recombination Velocity of Non-equilibrium Current Carriers in a Thin Semiconductor Specimen.—O. V. Sorokin. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2473–2479.) A thin, plane semiconductor specimen illuminated by a narrow beam of light moving with constant velocity along it is considered. Formulae are given connecting the experimentally determined distribution of non-equilibrium current carriers with lifetime, diffusion coefficient and the velocities of surface recombination.

537.311.33 : 53.083 **3181**

Comparator Method for Optical Lifetime Measurements on Semiconductors.—H. L. Armstrong. (*Rev. sci.*

Instrum., March 1957, Vol. 28, No. 3, p. 202.) An analogue comprising a vacuum photocell with a *CR* load is used to interpret the c.r.o. display in the light-pulse method described by Stevenson & Keyes (2008 of 1955).

537.311.33 : 535.215 **3182**

Measurement of Carrier Recombination Velocity by Conductivity Modulation.—K. D. Glinchuk, E. G. Miselyuk & E. I. Rashba. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2607–2613.) A photoconductivity-modulation method is described for the determination of minority-carrier lifetime τ and surface recombination velocity S. A 200-c/s square-wave-modulated beam of light of 40 000 lx intensity and about $0 \cdot 7 \times 0 \cdot 1$ mm² cross-section was used. The method is suitable for measurements of $S \approx 10^4$–10^5 cm/sec and $\tau \approx 10^{-1}$–10^{-2} μs. Preliminary experimental results show fair agreement with results obtained by other methods.

537.311.33 : 535.215 **3183**
: [546.28 + 546.289

The Properties of Semiconductor Devices.—A. A. Shepherd. (*J. Brit. Instn Radio Engrs*, May 1957, Vol. 17, No. 5, pp. 255–273.) A survey of recent trends in design including the production of germanium and silicon junctions during crystal growth and by alloying and solid-state diffusion. Semiconductor photocells are described. 33 references.

537.311.33 : 538.63 **3184**

Theory of Galvanomagnetic Phenomena in Semiconductors.—M. I. Klinger. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1055–1061.)

537.311.33 : 539.16/.18 **3185**

The Effect of Nuclear Radiation on Selected Semiconductor Devices.—G. L. Kleister & H. V. Stewart. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 931–937.) A range of commercial Ge and Si transistors and diodes has been exposed to neutron and gamma radiation. Deterioration of performance ensues. Both the surface- and bulk-controlled properties of the devices suffer change. In general, surface changes are of a transient nature whereas bulk changes (i.e. decrease of lifetime and change of resistivity) are permanent.

537.311.33 : 539.16.08 **3186**
: 621.387.462

Mechanism of the Forming of Pulses in Semiconductor Crystal Counters (Motion of Charges in Pulse Ionization in Semiconductors).—S. M. Ryvkin. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2667–2683.)

537.311.33 : 546.24 : 538.63 **3187**

Investigation of Influence of Pressure on Galvanomagnetic Properties of Tellurium at Low Temperatures.—N. E. Alekseevski, N. B. Brandt & T. I. Kostina. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 943–946.) An experimental investigation of the effect of 1 700 atm pressure on Te specimens with various impurity concentrations, at temperatures in the ranges $1 \cdot 4$–$4 \cdot 2$, 14–$20 \cdot 4$ and 60–78° K, and magnetic fields up to 20 000 oersted.

537.311.33 : 546.28 **3188**

Carrier Concentration Changes in *p* Si Induced by Heat Treatment.—Y. Matukura. (*J. phys. Soc. Japan*, Jan. 1957, Vol. 12, No. 1, pp. 103–104.) The effects of heat treatment (400°–1 000°C) on carrier concentration are given, including dependence on heating and cooling times, and on the ambient gas.

537.311.33 : [546.28 + 546.289] **3189**
: 539.169

Etching Behaviour of Pile-Irradiated Germanium and Silicon Single Crystals.—R. Chang. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 385–387.) Etching behaviour changes after pile irradiation, and preliminary work suggests a new experimental approach to the study of radiation damage.

537.311.33 : 546.289 **3190**

Effective Mass of Electrons and Holes in Germanium.—Z. Kopets. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2451–2458.) The effective mass is calculated from experimental data on the thermo-e.m.f. α, Hall constant and electrical conductivity, for the case when α increases with temperature. At a temperature of 125° K and electron concentration of $2 \cdot 7 \times 10^{15}$ the effective electron mass is $0 \cdot 47$, and the effective mass of the hole is $0 \cdot 23$ at a hole concentration of $1 \cdot 3 \times 10^{17}$. Results of effective electron and hole mass calculations are tabulated for concentrations of $2 \cdot 7 \times 10^{15}$–$6 \cdot 6 \times 10^{17}$ and $1 \cdot 3 \times 10^{17}$–$5 \cdot 5 \times 10^{18}$, respectively, at temperatures between 125 and 250° K.

537.311.33 : 546.289 **3191**

Precipitation of Cu in Ge.—A. G. Tweet. (*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 221–224.) Results of an experimental study are expressed in terms of the exponential decay with time of the unprecipitated fraction of Cu. The dependence of the time constant on temperature and dislocation density in the samples is discussed.

537.311.33 : 546.289 **3192**

Effect of Heavy Doping on the Self-Diffusion of Germanium.—M. W. Valentor & C. Ramasastry. (*Phys. Rev.*, 1st April 1957, Vol. 106, No. 1, pp. 73–75.) Measurements of the self-diffusion coefficients for intrinsic, heavily-doped *n*- and *p*-type Ge suggest that self-diffusion occurs by the vacancy mechanism.

537.311.33 : 546.289 **3193**

Influence of Electric Field in Diffusion Region upon Breakdown in Germanium *n-p* Junctions.—E. M. Pell. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 459–466.) The electric field arising from the *IR* drop in the diffusion region can (*a*) decrease the voltage at which a junction breaks down, (*b*) increase the photocurrent if the junction current is increased by any mechanism. An allied phenomenon of 'soft breakdown' has also been observed.

537.311.33 : 546.289 **3194**

Study of Injecting and Extracting Contacts on Germanium Single Crystals.—L. Y. Lin. (*Rev. sci. Instrum.*, March 1957, Vol. 28, No. 3, pp. 187–188.) The preparation of reliable contacts for determining lifetimes by the Many bridge method is described. Rectifying and ohmic contacts to a single crystal must be considered together.

537.311.33 : 546.289 : 538.6 **3195**

Oscillatory Magneto-absorption on the Direct Transition in Ge.—S. Zwerdling & B. Lax. (*Phys. Rev.*, 1st April 1957, Vol. 106, No. 1, pp. 51–52.) Measurements on intrinsic Ge at λ $1 \cdot 4$–6 μ, in magnetic fields up to 36 000 G.

537.311.33 : 546.289.241 **3196**

Electrical Properties of GeTe.—Y. Moriguchi & Y. Koga. (*J. phys. Soc. Japan*, Jan. 1957, Vol. 12, No. 1, p. 100.) Measurements of the Hall coefficient, resistivity and thermoelectric power of polycrystalline and single-crystal specimens of GeTe show that this material does not conform well to ordinary semiconductor theory, but shows semi-metallic properties.

537.311.33 : 546.482.21 **3197**

Ohmic and Rectifying Contacts to Semiconducting CdS Crystals.—W. C. Walker & E. Y. Lambert. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 635–636.)

537.311.33 : [546.57.23 + 546.57.24 **3198**

Rectification Properties of Silver Selenide and Telluride.—N. G. Klyuchnikov. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, p. 2603.) The rectifiers prepared possess a high forward/reverse current ratio. In Ag_2Te this ratio is approximately 18 000 : 1. With a potential difference $1 \cdot 4$ V the forward current is 40 A.cm⁻²; the reverse current is 12 mA.cm⁻² for 12 V. For Ag_2Se the forward current is 17 A.cm⁻² for 2 V, the reverse current 16 mA.cm⁻² for 12V. The rectifiers are unsuitable for rectification of an alternating current owing to the unstable nature of the barrier layer under forward current. Weak rectifying properties were also observed in other selenides and tellurides.

537.311.33 : 546.682.86 : 621.396.822 **3199**

Current Noise in Indium Antimonide.—D. J. Oliver. (*Proc. phys. Soc.*, 1st March 1957, Vol. 70, No. 447B, pp. 331–332.) Measurements in the frequency range 20 c/s–20 kc/s showed that excess noise was not detectable. It is estimated that the noise ratio (total-noise/thermal-noise) was less than 7 at 20 c/s, and less than 2 above 1 kc/s for currents up to 500 mA through a 1-Ω filament. See also 1142 of April (Suits et al.).

537.311.33 : 546.817.241 **3200**

Diffusion of Lead in Lead Telluride.—B. I. Boltaks & Yu. N. Mokhov. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2448–2450.) Results of preliminary experiments are reported. The temperature dependence of the diffusion coefficient is given by the relation $D = 2 \cdot 9 \times 10^{-5} \exp(-0 \cdot 6/kT)\text{cm}^2.\text{sec}^{-1}$ at temperatures well below the melting point of PbTe (905°C). The small activation energy (0·6 eV) and the high diffusion velocity indicate that the diffusion by migration of positively charged lead ions is of the interstitial type, similar to that of the diffusion of Cu in Ge.

537.311.33 : 546.817.241 **3201**

Influence of Impurities on the Electrical Properties of Lead Telluride.—T. L. Koval'chik & Yu. P. Maslakovets. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2417–2431.) A considerable increase of the concentration of free electrons in PbTe can only be obtained by the introduction of two impurities, e.g. Br and excess Pb. Introduction of single Br atoms, by substituting a $PbBr_2$ group for a 2PbTe group, may result in the production of lattice vacancies and a consequent decrease in current-carrier mobility. Different heat treatments of *p*-type PbTe give different current-carrier numbers. The relation between the carrier concentration and annealing temperature is given by $n_+ = A \exp(-\Delta E/2kT)$, where ΔE is near the value of the forbidden zone width for PbTe as determined from the temperature characteristic of the Hall constant.

537.311.33 : 621.314.7 **3202**

On the Injection of Carriers into a Depletion Layer.—W. G. Matthei & F. A. Brand. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 513–514.) Results are given of experiments which verify and extend the concept of injection into a depletion layer of a reverse-biased *p-n* junction.

537.311.33 : 621.385.833 **3203**

Image of *p-n* Junctions in Semiconductors Obtained by a Reflection Electron Microscope.—G. Bartz & G. Weissenberg. (*Naturwissenschaften*, April 1957, Vol. 44, No. 7, p. 229.) Micrograms are shown which were obtained by electron reflection and secondary emission from a Si *p-n* junction subjected to different bias voltages affecting the visibility and contrast of the transition zone.

538.2 : 539.23 **3204**

Magnetic Domain Patterns on Thin Films.—H. J. Williams & R. C. Sherwood. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 548–555.) The films were deposited in the presence of a magnetic field to establish a uniaxial direction of easy magnetization. This direction could be changed by reheating in a magnetic field with a new orientation, contrary to the behaviour with bulk specimens.

538.22 **3205**

Magnetic Properties of $MnAu_3$.—A. J. P. Meyer. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2028–2031.)

538.22 **3206**

Neutron Diffraction Study of the Magnetic Structures for the Perovskite-Type Mixed Oxides La(Mn,Cr)O_3.—U. H. Bents. (*Phys. Rev.*, 15th April 1957, Vol. 106, No. 2, pp. 225–230.) Investigation of the series La [x Mn, $(1-x)$ Cr]O_3 with x varying from 0 to 1.

538.221 **3207**

The Transformation of Haematite (α-Fe_2O_3) into Ferromagnetic Iron Oxide (γ-Fe_2O_3).—F. J. Lecznar. (*Acta tech. Acad. Sci. hungaricae*, 1957, Vol. 16, Nos. 3/4, pp. 383–398.) The different properties of the two varieties are outlined and experimental methods of producing the ferromagnetic form are described.

538.221 **3208**

Orientation of the Precipitations of Cobalt in a Cu-Co Alloy.—L. Weil, L. Gruner & A. Deschamps. (*C. R. Acad. Sci., Paris*, 15th April 1957, Vol. 244, No. 16, pp. 2143–2146.) Magnetic measurements made at low temperatures show the segregated masses of Co in a drawn specimen to be oriented.

538.221 **3209**

Orientational Superstructures in Fe-Ni Alloys.—E. T. Ferguson. (*C. R. Acad. Sci.*, Paris, 6th May 1957, Vol. 244, No. 19, pp. 2363–2366.) Uniaxial anisotropy induced in Fe-Ni alloys by reheating in a magnetic field is determined as a function of composition and of temperature and duration of reheating. See 1101 and 2983 of 1954 (Néel).

538.221 : 621.318.134 **3210**

Domain Patterns on Ferrite Single Crystals.—R. F. Pearson. (*Proc. phys. Soc.*, 1st April 1957, Vol. 70, No. 448B, p. 441, plates.) Powder patterns for $BaFe_{12}O_{19}$ (magnadur) and $Ba_3Zn_2Fe_{24}O_{41}$ are shown.

538.221 : 621.318.134 **3211**

Effects of Annealing on the Saturation Induction of Ferrites containing Nickel and/or Copper.—L. G. Van Uitert. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 478–481.) Real deviations from the smooth functions usually described occur in ferrites subjected to differing annealing treatments when the induction measurements are made at room temperature.

538.221 : 621.318.134 **3212**

Mechanism of Magnetization Processes in Very Weak Fields in some Ni-Zn Ferrites.—L. A. Fomenko. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 1092–1093.) Comment on paper by Rathenau & Fast (2136 of 1956).

538.221 : 621.318.134 **3213**

Form of Polder Tensor for Single-Crystal Ferrite with Small Cubic-Symmetry Anisotropy Energy.—H. Seidel & H. Boyet. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 452–454.) The tensor is derived for the 110 plane, and its elements are given as a function of several parameters.

538.221 : 621.318.134 **3214**

Ferrimagnetic Resonance of Gadolinium Garnet at 9 300 Mc/s.—J. Paulevé. (*C. R. Acad. Sci., Paris*, 1st April 1957, Vol. 244, No. 14, pp. 1908–1910.) Experimental results were obtained for temperatures from 4° K to 700° K. The compensation temperature is 290° K at which there is a discontinuation in the value of the field strength *H*. See 183 of 1956.

538.221 : 621.318.134 : 538.569.4 **3215**

Domain Structure Effects in an Anomalous Ferrimagnetic Resonance of Ferrites.—R. C. LeCraw & E. G. Spencer. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 399–405.) Measurements of the intrinsic tensor permeability of unsaturated Ni ferrite at 9·3 kMc/s have revealed an anolamous resonance for negative (anti-Larmor) circularly polarized fields. Effects are explained by theory. Since the anomalous resonance depends on domain structure and appears to occur generally, it is useful for studying magnetization processes and h.f. phenomena in ferrites.

538.63 : 546.59 **3216**

Galvanomagnetic Properties of Gold.—N. E. Alekseevski & Yu. P. Gaĭdukov. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 947–950.) Electrical conductivity and the effect of magnetic fields up to 23 000 oersted were determined experimentally in the temperature range 0·05–20·4° K.

538.632 : 546.87 **3217**

On the Hall Effect in Bismuth at Low Temperature.—H. Hasegawa, S. Nakano & N. Hashitsume. (*J. phys. Soc. Japan*, Jan. 1957, Vol. 12, No. 1, p. 104.) Theoretical discussion and re-analysis of the experimental data of Reynolds et al. (*Phys. Rev.*, 1st Dec. 1954, Vol. 96, No. 5, pp. 1203–1207).

538.652 : 534.232 **3218**

Dynamic Magnetostrictive Properties of Ni-Fe Alloys.—C. M. Davis, Jr, H. H. Helms & S. F. Ferebee. (*J. acoust. Soc. Amer.*, April 1957, Vol. 29, No. 4, pp. 431–434.) The suitability of Ni-Fe alloys containing from 35% to 67·3% Ni for use in electromechanical transducers is investigated.

539.234 : 546.26 **3219**

Electrical Properties of Arc-Evaporated Carbon Films.—M. D. Blue & G. C. Danielson. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 583–586.) The electrical properties of the films before heat treatment indicate that they are more truly amorphous than any other form of carbon.

621.3.032.12 **3220**
: [621.387 + 621.372.56.029.64

On Processing Titanium Hydride Replenishers.—D. Walsh & P. M. Shearman. (*J. sci. Instrum.*, April 1957, Vol. 34, No. 4, pp. 161–162.) Methods of processing replenishers are described, with particular reference to their use in microwave coaxial-diode attenuators.

669.046.5 : 537.533.9 **3221**

The Floating-Zone Melting of Refractory Metals by Electron Bombardment.—A. Calverley, M. Davis & R. F. Lever. (*J. sci. Instrum.*, April 1957, Vol. 34, No. 4, pp. 142–147.) See 2147 of 1956.

MATHEMATICS

517.3 3222

The Fermi-Dirac Iutegrals

$$F_p(\eta) = (p!)^{-1}\int_0^\infty \epsilon^p(e^{\epsilon-\eta}+1)^{-1}d\epsilon.$$ — R. B. Dingle. (*Appl. sci. Res.*, 1957, Vol. B6, No. 4, pp. 225–239.) Complete expansions are developed and values are tabulated for orders −1 and 0 for positive and negative arguments, and for orders 1, 2. 3, 4 for positive arguments.

517.3 3223

The Bose-Einstein Integrals

$$B_p(\eta) = (p!)^{-1}\int_0^\infty \epsilon^p(e^{\epsilon-\eta}-1)^{-1}\,d\epsilon$$ — R. B. Dingle. (*Appl. sci. Res.*, 1957, Vol. B6, No. 4, pp. 240–244.) Complete expansions are developed so that integrals of all orders can be calculated without numerical integration.

517.3 3224

The Integrals

$$\mathsf{E}_p(x) = (p!)^{-1}\int_0^\infty \epsilon^p(1+x\epsilon^3)^{-1}\,e^{-\epsilon}\,d\epsilon$$

and $$\mathsf{F}_p(x) = (p!)^{-1}\int_0^\infty \epsilon^p(1+x\epsilon^3)^{-2}\,e^{-\epsilon}\,d\epsilon$$

and their Tabulation.—R. B. Dingle, D. Arndt & S. K. Roy. (*Appl. sci. Res.*, 1957, Vol. B6, No. 4, pp. 245–252.) The applications and properties of the integrals are discussed and integrals are tabulated for arguments $x = 0(\cdot 002)\cdot 01$, $\cdot 02(\cdot 02)\cdot 1(\cdot 1)1(\cdot 2)2(\cdot 5)5(1)10(2)20$ at half-integer spacings of the order p.

517.9 3225

Sufficient Conditions for Non-oscillation and Oscillation of the Solution of the Equation $y'' + p(x)y = 0$.—V. A. Kondrat'ev. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1957, Vol. 113, No. 4, pp. 742–745. In Russian.)

51 : 621.3 3226

Mathematics for Electronics with Applications. [Book Review]—H. M. Nodelman & F. W. Smith. Publishers: McGraw-Hill, New York, 1956, 391 pp., \$7. (*J. Franklin Inst.*, Jan. 1957, Vol. 263, No. 1, pp. 84–85.) The book is divided into five parts including equation testing, circuit analysis and series.

MEASUREMENTS AND TEST GEAR

53.087 : 519.25 : 621.396.822 3227

The Analysis of Finite-Length Records of Fluctuating Signals.—M. J. Tucker. (*Brit. J. appl. Phys.*, April 1957, Vol. 8, No. 4, pp. 137–142.) The problem is that of estimating the characteristics of a signal by Fourier-type analysis or by direct measurement, and its fundamentals are presented with the simplest possible mathematics. The paper is concerned mainly with the accuracy of estimates of the power spectra and r.m.s. amplitudes of stationary random Gaussian processes.

621.317.3 + 621.372.54] : 621.396.663 3228

Possible Applications of Goniometers in Telecommunications.—H. Fricke. (*Nachrichtentech. Z.*, Feb. 1957, Vol. 10, No. 2, pp. 65–73.) Survey of applications ncluding methods of measuring frequency, admittance (see 806 of 1955), line reflection coefficients and quadripole characteristics, and the use of goniometers in frequency separating networks and as band-pass filters.

621.317.32 : 537.311.4 3229

An Improved Technique for the Measurement of Contact Potential Differences.—K. A. Macfadyen & T. A. Holbeche. (*J. sci. Instrum.*, March 1957, Vol. 34, No. 3, pp. 101–105.) A development of the vibrating-capacitor method is described. It is used for the measurement of contact potential differences between metals immersed in an insulating liquid to an accuracy of 0·1 mV.

621.317.335.2 3230

A Bridge Network for the Precise Measurement of Direct Capacitance.—A. C. Lynch. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 363–366. Discussion, pp. 366–367.) This comparison bridge, with both oscillator and detector connected to earth, is suitable for use at frequencies from 4 c/s to 20 Mc/s. The balance point is independent of frequency and of stray capacitance and conductance to earth.

621.317.35.029.3 3231

Construction and Investigation of a Spectral Analyser of High Selectivity in the Low-Frequency Region.—M. Savelli & J. C. Solera. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2020–2023.) A note on the theory and performance of an analyser for noise measurements in which the signal after heterodyning is filtered at near-zero frequency. The output voltage is applied to a thermocouple followed by a d.c. amplifier.

621.317.352.029.5/.6 : 621.315 3232

Measurement of Line Attenuation in the Frequency Range from 20 to 250 Mc/s.—J. F. Bornhardt & G. Buhmann. (*Elektrotech. Z., Edn A*, 1st Jan. 1957, Vol. 78, No. 1, pp. 6–12.) Two test methods of universal application based on the use of the line resonance characteristics are described.

621.317.39 3233

The [present] **State of Electrical Measuring Techniques with Particular Reference to their Application for the Measurement of Non-electrical Quantities.**—W. Hunsinger. (*Elektrotech. u. Maschinenb.*, 1st Feb. 1957, Vol. 74, No. 3, pp. 49–57.)

621.317.42 3234

Instrument for Relative Measurements of Alternating Magnetic Fields.—I. S. Shpigel', M. D. Raizer & E. A. Myae. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 111–119.) The instrument described is based on the principle of resonance absorption, and is designed for relative measurements of weakly inhomogeneous magnetic fields with maximum differences of field strength between a pair of points in the field of $\Delta H_{max} = 3\% \ H_0$. The errors of the instrument do not exceed $\pm 3\% \ \Delta H_{max}$ at $H_0 \approx 160$ oersted.

621.317.42 3235

Some Possibilities of Measuring Magnetic Field Strength with Film-Type Hall-E.M.F. Generators prepared from HgSe, HgTe and their Solid Solutions.—O. D. Elpat'evskaya & A. R. Regel'. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2432–2438.)

621.317.616 + 621.317.77 3236

A Phase-Characteristic Wobbulator for the Video-Frequency Range.—E. Legler. (*Rundfunktech. Mitt.*, Feb. 1957, Vol. 1, No. 1, pp. 20–23.) The instrument described has a range of about 0·3–10 Mc/s for phase angles up to 1 440°. The phase characteristics are displayed on a c.r.o. screen. Frequency response curves are also obtainable.

3237

621.317.616 : 621.373.42 : 621.376.3

A Simple F.M. Signal Generator and Wobbulator with Very Large Frequency Deviation.—E. G. Woschni. (*NachrTech.*, Feb. 1957, Vol. 7, No. 2, pp. 51–55.) A relative frequency deviation of about 50% in the l.f. and m.f. range is obtainable with the single-stage oscillator circuit described.

621.317.7 : 621.372.8 : 538.569.4 3238

Directional-Coupler Arrangement for Paramagnetic - Resonance Apparatus using Circularly Polarized Waves.—A. Charru. (*C. R. Acad. Sci., Paris*, 15th April 1957, Vol. 244, No. 16, pp. 2146–2147.) Description of apparatus for operation at 3 kMc/s making use of a coupled rectangular-waveguide section perpendicular to the axis of a circular waveguide, whereby measurement can be made of the reflected-wave energy alone. Measurements on diphenyl picryl hydrazyl indicate a gain in sensitivity of the order of 30 over that of apparatus previously described (*ibid.*, 13th Aug. 1956, Vol. 243, No. 7, pp. 652–654.)

621.317.725 : 621.397.62 3239

Line Timebase Measurements: E.H.T. Rectifier Heater Voltages.—A. Ciuciura. (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 25–29.) A thermocouple method gives accuracy within 1%.

621.317.729 3240

Flux Plotting Analogue for an Axially Symmetric Potential Field.—W. L. Beaver. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 579–582.) The use of an electrolyte tank is described.

621.317.733.029.3 : 621.314.7 **3241**

Simple Transformer Bridge for the Measurement of Transistor Characteristics.—W. F. Lovering & D. B. Britten. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 368–373. Discussion, p. 373.) The real and imaginary components of the impedance parameters of point-contact or junction transistors at 1 kc/s, connected in the common-base circuit, may be quickly measured.

621.317.755 **3242**

A Versatile Oscilloscope.—(*Electronic Applic. Bull.*, Oct. 1956, Vol. 17, No. 1, pp. 1–10.) The instrument described uses a timebase generator that can be converted into an amplifier for horizontal deflection.

621.317.755 **3243**

The Ultimate Performance of the Single-Trace High-Speed Oscillograph.—M. E. Haine & M. W. Jervis. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 379–384. Discussion, pp. 390–392.) An improvement of 10–100 times the present resolution is possible theoretically, without loss of deflection sensitivity.

621.317.755 **3244**

The Design and Performance of a New Experimental Single-Transient Oscillograph with Very High Writing Speed.—M. E. Haine & M. W. Jervis. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 385–390. Discussion, pp. 390–392.) The oscillograph has a limiting resolution of 2×10^{-13} sec per spot width.

621.317.755 : 621.385.832 **3245**

A New Type of Multibeam Cathode-Ray Oscillograph.—Fert, Lagasse & Clot. (See 3363.)

621.317.755.089.6 **3246**

Line Timebase Measurements : Oscilloscope Calibration Unit.—P. L. Mothersole (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 30–31.) Provides square waves variable in amplitude from 100 mV to 100 V.

621.317.784 **3247**

A Logarithmic Wattmeter.—J. A. Bennet. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 266–271.) An electronic wattmeter using logarithmic and anti-logarithmic circuits for the measurement of arc loss in mercury-arc rectifiers is described.

621.385.001.4 : 534.1 **3248**

Standardized White-Noise Tests.—J. Robbins. (*Electronic Ind. Tele-Tech*, Feb. 1957, Vol. 16, No. 2, pp. 68–69 . . 122.) A noise generator is used to operate an accelerometer to produce wide-frequency, high-*g* vibrations for performance tests on valves and circuit components.

3249

621.385.032.216.001.4 : 621.317.018.75

A Pulse Method for Measuring the Interface Layer of Oxide Cathodes.—Lieb. (See 3355.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.2 : 621.396.9 **3250**

Surveying Instrument for the Precise Measurement of Length.—(*Engineer, Lond.*, 5th April 1957, Vol. 203, No. 5280, p. 538.) Description of the " tellurometer", which has a range of about 35 miles and probable error 1 part in 300 000 ± 2 in. Measurement is based on the phase shift between the outgoing and incoming modulation of a wave transmitted by the 'master' station, and re-radiated from the 'remote' station. The wavelength used is 10 cm, with 'pattern' modulation frequencies of 10, 9·99, 9·9 and 9 Mc/s.

531.78.087.252 : 621.314.7 **3251**

Transistorized Strobe measures Shaft Torque.—J. Patraiko. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 147–149.) An improved stroboscope for torque measurements on shafts rotating up to 60 000 r.p.m. For a more detailed account, see *Trans. Inst. Radio Engrs*, March 1956, Vol. PGIE-3, pp. 3–11.

621-52 : 621.314.7 : 623.98 **3252**

Transistors Stabilize Missile Ships.—R. Scheib, Jr. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 138–143.) A description of measuring, computing and servo-mechanism techniques used in controlling underwater fins to reduce roll of a missile-launching ship.

621-57 : 537.228.4 **3253**

High-Speed Electrostatic Clutch.—C. J. Fitch. (*Product Engng*, Feb. 1957, Vol. 28, No. 2, pp. 189–191.) Description and performance details of a 3-clutch unit based on the Johnsen-Rahbeck effect.

621.317.39.082 : 621.38 **3254**

Electromechanical Devices.—L. A. Goncharski. (*Uspekhi fiz. Nauk*, Feb. 1957, Vol. 61, No. 2, pp. 277–302.) See also 613 of 1957 and back references.

621.317.79 : 531.7 **3255**

Transducer Indicator System.—W. C. Vaughan. (*Electronic Radio Engr*, Aug. 1957, Vol. 34, No. 8, pp. 286–290.) Theoretical analysis of a displacement indicator using an inductance transducer and moving-coil dynamometer indicating instrument. See also 895 of March (Spratt).

621.365.5 : 621.385.3 **3256**

The Design and Operation of High-Power Triodes for Radio-Frequency Heating.—Pohl. (See 3359.)

621.373.029.4/.51 : 621.383.2 **3257**

Generation of Oscillations in Photocells and Photomultipliers.—P. V. Makovetski. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2652–2660.) Stable ultrasonic-frequency oscillations can be produced within narrow ranges of illumination intensities and electrode potentials ; the frequency of oscillations is a function of the circuit time constant. Experimental results are presented graphically.

621.384.6 **3258**

Envelope Method for Investigating Free Oscillations in Accelerators.—A. M. Baldin, V. V. Mikhailov & M. S. Rabinovich. (*Zh. eksp. teor. Fiz.*, Dec. 1956, Vol. 31, No. 6 (12), pp. 993–1001.) The envelope of particle trajectories for a large number of revolutions is used in the calculations, rather than the individual trajectories themselves. The method is illustrated by an application to accelerators with a sectional magnet and with strong focusing.

621.384.612 **3259**

Excitation of Betatron Oscillations by Synchrotron Momentum Oscillations in a Strong-Focusing Accelerator.—Yu. F. Orlov. (*Zh. eksp. teor. Fiz.*, Jan. 1957, Vol. 32, No. 1, pp. 130–134.)

621.384.7 **3260**

An Application of a Difference-Type Electrostatic Field in the Spectroscopy of Beams of Charged Particles.—M. I. Korsunski & V. A. Bazakutsa. (*C. R. Acad. Sci. U.R.S.S.*, 11th April 1957, Vol. 112, No. 5, pp. 1029–1031. In Russian.)

621.385.833 **3261**

Electron-Optical Systems, the Fields of which are Independent of One Coordinate.—Yu. V. Vandakurov. (*Zh. tekh. Fiz.*, Nov. 1956, Vol. 26, No. 11, pp. 2578–2594.) Electron-optical systems with cylindrical or rotationally symmetrical fields are considered. In these systems one of the two differential equations determining the deviation of the trajectory of any given particle from the axial trajectory can be replaced by a first-order equation. The results obtained are applied to the investigation of fields for which the electric potential is symmetrical and the scalar magnetic component antisymmetrical to the central plane. In particular, fields of (*a*) a linear current, (*b*) a spherical condenser, (*c*) a cylindrical condenser and (*d*) a combination of an inhomogeneous magnetic field with that of a cylindrical condenser, are considered. In the latter case fields with large dispersion are found in the absence of second-order aberrations depending on the divergence angle of the beam in the central plane.

621.385.833 **3262**

Third-Order Paths in Electron Mirrors.—P. Schiske. (*Optik, Stuttgart*, Jan. 1957, Vol. 14, No. 1, pp. 34–45.)

621.385.833 **3263**

Experimental Study of the Velocity Spectrum of Electrons Transmitted by the Diaphragm of an Emission-Type Electron Microscope.—F. Pradal & R. Simon. (*C. R. Acad. Sci., Paris*, 15th April 1957, Vol. 244, No. 16, pp. 2150–2152.) The velocity spectrum becomes narrower as the diaphragm diameter is reduced. See also 2264 of July (Fert & Simon).

621.385.833 : 537.533.72 **3264**

Investigations of a Special Deflection System for Electron Beams.—R. Gain. (*Optik, Stuttgart*, Feb. 1957, Vol. 14, No.

2, pp. 49–71.) The system described, which has the deflection characteristics of an inhomogeneous prism, consists of a wire and a flat plate at a potential different from that of the adjacent electrodes. The equipotential lines are plotted for various electrode arrangements and analytical formulae are derived. Calculations of deflection and aberration are in good agreement with results of tests on an experimental c.r. tube incorporating this system; comparisons with a normal c.r. tube are also made.

621.385.833 : 537.533.72 **3265**

Mechanically Adjustable Electrostatic Unipotential Lenses.—B. Rajewsky & W. Lippert. (*Optik, Stuttgart*, Feb. 1957, Vol. 14, No. 2, pp. 72–73.) In the lens system outlined one of three cylindrical electrodes is movable in the direction of its axis to change magnification.

621.385.833 : 537.533.72 **3266**

Asymptotic Image Distortions.—F. Lenz. (*Optik, Stuttgart*, Feb. 1957, Vol. 14, No. 2, pp. 74–82.) A modified theoretical treatment of aberration is developed which takes account of the interaction between the lenses, e.g. in an electron microscope.

621.387.4 : 612 **3267**

Some Nucleonic Instruments for Clinical Use.—E. W. Pulsford & N. Veall. (*J. Brit. Instn Radio Engrs*, June 1957, Vol. 17, No. 6, pp. 299–307.) Detailed descriptions are given of a four-channel logarithmic ratemeter, a portable Geiger-counter-type clinical monitor, a recording ratemeter for observing transient phenomena, and a β-γ ionization chamber.

3268

621.387.462 : 539.16.08 : 537.311.33

Mechanism of the Forming of Pulses in Semiconductor Crystal Counters (Motion of Charges in Pulse Ionization in Semiconductors).—S. M. Ryvkin. (*Zh. tekh. Fiz.*, Dec. 1956, Vol. 26, No. 12, pp. 2667–2683.)

621.398 : 551.46 **3269**

Buoy telemeters Ocean Temperature Data.—R. G. Walden, D. D. Ketchum & D. N. Frantz, Jr. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 164–167.) On reception of an appropriate tone-modulated signal a transistor receiver triggers a transmitter in the buoy. A 15-sec transmission cycle includes low and high reference tones and a tone determined by a thermistor indicating water temperature. A range of 600 miles has been obtained under favourable conditions.

621.398 : 621.317.361.029.4 **3270**

Methods of Telemetering L.F. Oscillatory Phenomena.—W. Nicolai. (*Elektronische Rundschau*, Jan. 1957, Vol. 11, No. 1, pp. 8–12.) A comparison shows f.m. methods to be superior to a.m. methods for telemetering frequencies ranging from zero to about 1 kc/s. Details are given of single-channel experimental equipment and of a two-channel telemetering system used for medical applications.

681.188 : 413 : 8.03 **3271**

Automatic Programming of Operations in Translation from One Language into Another.—S. N. Razumovski. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1957, Vol. 113, No. 4, pp. 760–761. In Russian.)

621.398 + 621.317.083.7 **3272**

Radio Telemetry. [Book Review]—M. H. Nichols & L. L. Rauch. Publishers: Wiley & Sons, New York and Chapman & Hall, London, 2nd edn 1957, 461 pp., 96s. (*Electronic Engng*, June 1957, Vol. 29, No. 352, p. 301.) ". . . a comprehensive treatment of the basic theory of radio telemetry together with a review of current equipment." Originally published for the U.S.A.F.

PROPAGATION OF WAVES

538.566.2 **3273**

On the Theory of Reflection from a Wire Grid Parallel to an Interface between Homogeneous Media.—J. R. Wait. (*Appl. sci. Res.*, 1957, Vol. B6, No. 4, pp. 259–275.) The plane-wave solution for arbitrary incidence is generalized for cylindrical-wave excitation. The problem of energy absorption from a magnetic line source by a grid situated on the surface of a dissipative half-space is examined, this being a two-dimensional analogy of a vertical aerial with radial wire earth system.

621.396.11 **3274**

Fading of Radio Waves Scattered by Dielectric Turbulence.—R. A. Silverman. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 506–511.) Fading is attributed to (*a*) time variation of the scattering eddies as seen in a coordinate system moving with the local wind velocity, and (*b*) Doppler shifting produced by the convection of the scattering eddies by the mean wind and by the macro-eddies.

621.396.11 : 551.510.535 **3275**

Influence of Collisions on Ionospheric Reflection.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2031–2033.) Further analysis of the propagation of plane waves in a stratified medium for linear variation of ionization with height (see 3510 of 1956). An expression for viscous damping is introduced representing the effect of collisions. See also *ibid.*, 29th April 1957, Vol. 244, No. 18, pp. 2298–2299.

3276

621.396.11 : 551.510.535 : 621.396.677

The Gain of a Directive Receiving Aerial for Short-Wave Back-Scatter.—B. Beckmann & K. Vogt. (*Nachrichtentech. Z.*, Feb. 1957, Vol. 10, No. 2, pp. 90–91.) The aerial gain measurements briefly reported were made in West Germany with rhombic aerials of 115-m sides for normal reception of Ankara and New York and for back-scatter reception of transmissions of the nearby station at Bonames. A comparison of results obtained confirms the coherent nature of back-scatter radiation. See also 1888 of June.

621.396.11 : 621.396.96 **3277**

Back-Scattering from Water and Land at Centimetre and Millimetre Wavelengths.—C. R. Grant & B. S. Yaplee. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 976–982.) Describes measurements of the average radar cross-section of echo per unit area of surface, made at various angles of incidence at wavelengths of 3·2 cm, 1·25 cm, and 8·6mm using vertical polarization. Over sea, the cross-section was found mostly to increase with frequency and wind velocity, a considerable specular component being present at normal incidence. Over land, it varied in a complex manner with the type of terrain, but usually increased with frequency and a specular component was sometimes present. Dry vegetation was practically an isotropic scatterer.

621.396.11.029.6 **3278**

Influence of the Form of the Structural Function of Inhomogeneities of the Permittivity of Air on Long-Distance Tropospheric Propagation of Ultra Short Waves.—V. N. Troitski. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 34–37.) Expressions are derived for the median field strength and the possible transmission bandwidth for tropospheric propagation, assuming a structural inhomogeneity function of the form $\overline{(\epsilon_1-\epsilon_2)^2} = B_\beta a^\beta$, where ϵ_1 and ϵ_2 are the dielectric constants at points 1 and 2, respectively, a is the distance between the points, and B and β are constants. The influence of the form of this function on field strength and distribution is analysed; results indicate that it is small for the former, but considerable for the latter.

621.396.11.029.62 **3279**

Evidence regarding the Mechanism of Long-Range Propagation at Metre Wavelengths by Measurements of Tropospheric Drift.—L. Klinker. (*Z. Met.*, Feb. 1957, Vol. 11, No. 2, pp. 43–49.) A comparison of propagation measurements beyond optical range over a sea path and records of wind speed and direction at various altitudes appears to indicate that propagation is due predominantly to partial reflections at free inversions.

621.396.11.029.62 **3280**

Contribution on V.H.F. Propagation over the Sea.—B. Abild. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 4th Dec. 1956, Vol. 8, Nos. 5/6, pp. 103–108.) Measurements in the 3-m band made in the period from 1952 to 1954 for both sea and land paths in North Germany are used for a comparison of daily and yearly variations of field strength. Over land, particularly in summer, daily variations are large; over the sea only slight variations occur. The seasonal fluctuation over the sea is greater than over land. The difference is explained by climatic conditions.

621.396.11.029.62 **3281**

Forward - Scatter Observations at 50 Mc/s.—K. Bibl, H. A. Hess & K. Rawer. (*Arch. elekt. Übertragung*, Feb. 1957, Vol. 11, No. 2, pp. 59–62.) Report on field strength measurements over two paths of v.h.f. scatter propagation at 51·3 Mc/s. A 10-kW c.w. transmitter at Kootwyk (Netherlands) and a special receiver of 70-c/s bandwidth distant 500 km (Neversshausen, near Freiburg) and 1 000 km (Toulon) were used. The mean propagation loss compared with free-space propagation was 90–100 dB for either path; short bursts due to meteor trails were also observed. Some typical records are shown.

3282

621.396.11.029.62 : 551.510.535

Diurnal Variations of Signal Level and Scattering Heights for V.H.F. Propagation.—A. D. Wheelon. (*J. geophys. Res.*, June 1957, Vol. 62, No. 2, pp. 255–266.) Scattering in the lower ionosphere is due to turbulent fluctuations of the electron plasma. The theory of gradient mixing is applied to the afternoon and early evening periods when solar control is probably more important than meteoric influences. The theory is used to explain observational data.

621.396.812.3 : 551.510.535 **3283**

Fading and Random Motion of Ionospheric Irregularities.—S. N. Mitra & R. B. L. Srivastava. (*Indian J. Phys.*, Jan. 1957, Vol. 31, No. 1, pp. 20–42.) An analysis of the fading observed on transmissions at 1 020 kc/s shows that the r.m.s. velocity in the line-of-sight of the ionospheric irregularities varied between 4 and 25 m/s.

RECEPTION

621.376.23 **3284**

Low-Distortion A.M. Demodulation.—U. Köhler. (*NachrTech.*, Feb. 1957, Vol. 7, No. 2, pp. 56–60.) Considerations, such as the choice of detection system, underlying the design of a high-grade test receiver are discussed.

621.376.23 : 621.396.822 **3285**

Analysis of a General System for the Detection of Amplitude-Modulated Noise.—E. Parzen & N. Shiren. (*J. Math. Phys.*, October 1956, Vol. 35, No. 3, pp. 278–288.) Various statistics of the output, with and without modulation, are computed for a system involving square-law detectors, and a criterion for the design of detection systems is considered.

621.376.333 **3286**

Dynamic Testing of Ratio Detectors.—G. Rösler. (*Funk-Technik, Berlin*, Feb. 1957, Vol. 12, No. 3, pp. 68–71.) Description of method and circuits which use pulsed a.m. for checking the characteristics of ratio detectors.

621.396.621 : 621.376.33 **3287**

Limiters and Discriminators for F.M. Receivers: Part 5.—G. G. Johnstone. (*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 378–384.) The measurement of the a.m. suppression ratio and various types of limiter are described. Part 4: 2580 of August.

621.396.621.029.53/.55 **3288**

Unconventional Communications Receiver.—(*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 388–389.) A description of the Racal Type-RA17 receiver which gives continuous coverage from 500 kc/s to 30 Mc/s without band switching.

621.396.621.54 : 621.376.2/.3 **3289**

I.F. Amplifiers in F.M./A.M. Receivers.—L. W. Hampson. (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 11–24.) Circuit design is discussed with particular reference to the Type-EF89 variable-mu r.f. pentode.

621.396.82 : 621.376.3 **3290**

V.H.F. Broadcasting.—R. D. A. Maurice. (*Electronic Radio Engr*, Aug. 1957, Vol. 34, No. 8, pp. 300–309.) The problem of impulsive noise and man-made interference to v.h.f. broadcasting is discussed and methods of improving signal/noise ratio with particular reference to f.m. reception are described. A practical illustration of the problem shows field strengths required to protect a given reception from motor-car ignition interference.

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 **3291**

Distortion in F.M. Systems.—A. Ditl. (*Hochfrequenztech. u. Elektroakust.*, Jan. 1957, Vol. 65, No. 4, pp. 136–148.) Following on theoretical investigations of transient characteristics of f.m. systems (*ibid.*, May 1956, Vol. 64, No. 6, pp. 184–193.) the signal distortion is calculated with reference to bandwidth limitations particularly in relaying television signals. An approximate formula for calculating the maximum permissible frequency shift for a given amplifier characteristic and amount of distortion is obtained.

621.39.011.1 **3292**

Coding and Compression of Codes.—L. N. Korolev. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1957, Vol. 113, No. 4, pp. 746–747.) In Russian.)

621.391.5.029.45 **3293**

Selective A.F. Induction Signalling.—L. E. Philipps. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 180–181.) A one-way signalling system operating in the range 6–20 kc/s.

621.394.441 **3294**

The Voice - Frequency Telegraphy System WT24/1.—H. H. Voss & J. Arnold. (*Nachrichtentech. Z.*, Feb. 1957, Vol. 10, No. 2, pp. 81–87.)

621.396.001.11 **3295**

Improvement of Binary Transmission by Null-Zone Reception.—F. J. Bloom, S. S. L. Chang, B. Harris, A. Hauptschein & K. C. Morgan. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 963–975.) Single-null-zone (3-level) detection in a binary system of equal and opposite signals is shown theoretically to be capable of achieving about one-half of the improvement in information rate attainable by an infinite number of levels. Double-null (4-level) detection offers only a slight additional increase in rate, but may be better in practice because it is much less sensitive to variations in null level.

621.396.2 : 621.372.5 **3296**

Bandwidth Occupied in Pulse Transmission.—M. S. Gurevich. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 38–43.) Analysis of the distribution of the pulse energy in a given frequency band is presented. The spectral distribution of energy is calculated for rectangular, trapezoidal, triangular, cos and $\cos^2$ pulses.

621.396.65 **3297**

Some Information on the Radio Link Monte Erice [Sicily] **–Bou Kornine (Tunisia).**—(*Poste e Telecomunicazioni*, Feb. 1957, Vol. 25, No. 2, pp. 157–159.)

621.396.712.2 : 621.395.623.8 **3298**

New High-Grade Monitoring Equipment for [studio] **Control Rooms.**—F. Enkel. (*Elektronische Rundschau*, Feb. 1957, Vol. 11, No. 2, pp. 51–54.) The installation described has characteristics closely corresponding to those of modern high-fidelity domestic receivers. A spherical loudspeaker system is used for the upper frequencies. Associated matching, correcting and compensating devices are briefly outlined.

621.396.931 **3299**

Recent Developments in Mobile Radio in Britain.—J. R. Brinkley. (*J. Brit. Instn Radio Engrs*, May 1957, Vol. 17, No. 5, pp. 287–293.) Some comments on the relation between modulation systems and channel spacing are included.

SUBSIDIARY APPARATUS

621-52 **3300**

The Inductosyn and its Application to a Programmed Coordinate Table.—L. H. R. Harrison, B. A. Horlock & F. D. Hunt. (*Electronic Engng*, June & July 1957, Vol. 29, Nos. 352 & 353, pp. 254–259 & 331–335.) Description of a new control element developed for the U.S.A.F. with an accuracy within 5 seconds of arc in its rotary form and within 0·0001 in. in its linear form. Its application to position control systems is described, particularly for machine tools. For a similar description, see *J. Brit. Instn Radio Engrs*, July 1957, Vol. 17, No. 7, pp. 369–383 (Finden & Horlock).

621-526 : 621.375.088.7 **3301**
A Zero-Correcting Servo for Use with D.C. Amplifiers.—B. Shackel & M. Beaney. (*Electronic Engng*, June 1957, Vol. 29, No. 352, pp. 284–286.) This device is designed primarily to compensate drift voltages which have developed in the external output circuit.

621.314.63 : 537.311.33 **3302**
Tunnel Effect in Sulphide Rectifiers.—Yu. M. Volokobinski. (*C. R. Acad. Sci. U.R.S.S.*, 21st April 1957, Vol. 113, No. 6, pp. 1239–1242. In Russian.) Results of a series of experiments on Cu_2S/Al and Cu_2S/Mg rectifiers confirm some fundamental conclusions of the tunnel-effect theory developed by Frenkel (*Phys. Rev.*, 1st Dec. 1930, Vol. 36, No. 11, pp. 1604–1618) and others.

621.316.72 : 621.314.7 **3303**
A Regulated Power Unit with Transistor Control.—R. E. Reynolds. (*Mullard tech. Commun.*, Feb. 1957, Vol. 3, No. 21, pp. 34–36.) A transistor controls the series valve element eliminating the necessity for a negative rail.

621.316.9 : 621.396.61 : 621.385 **3304**
Electronic Crowbar Protects Transmitter.—R. G. Wenner. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 174–176.) Details of an electronic switch for protecting high-power c.w. transmitting valves against flash-arc destruction.

621.318.57 **3305**
Electronic Time-Delay Relay with Ionization Chamber as Timing Device.—H. Jucker. (*Elektronische Rundschau*, Jan. 1957, Vol. 11 No. 1, pp. 13–14.)

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 : 389.6 **3306**
The New Television Standards of the C.C.I.R.—H. A. Laett. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, Jan. 1957, Vol. 35, No. 1, pp. 1–6.) The standards applicable to the 625-line system as established or revised by the Warsaw conference (Aug./Sept. 1956) are summarized.

621.397.5 : 535.623 **3307**
Colour Television Transmission.—K. Teer. (*Electronic Radio Engr*, Aug. & Sept. 1957, Vol. 34, Nos. 8 & 9, pp. 280–286 & 326–332.) Further details of principles and circuits of a system [see 1224 of 1956 (Haantjes & Teer)], using two subcarriers combined with a luminance signal, originally demonstrated before C.C.I.R. Study Group XI in 1955. Subsequent modifications and improvements are also described.

621.397.5 : 535.623 **3308**
Colour Television marks Time.—(*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 354–355.) Research trends from the Paris International Symposium suggest the N.T.S.C. system is the most suitable for European use.

621.397.5 : 535.623 **3309**
An Alternative Colour TV System.—E. J. Gargini. (*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 361–364.) The signal is compatible for monochrome receivers in a similar manner to the N.T.S.C. system, but no brightness information is carried by the colour signal. This is achieved by instantaneously dividing the chrominance signal by a brightness signal which is the mean of the three-colour component signals.

621.397.5 : 535.623 : 621.385.832 : 535.37 **3310**
Luminophores based on ZnS and ZnSe for Colour Television.—Blazhnova, Mokrintseva & Kas'yanova. (See 3159.)

621.397.5 : 535.7 **3311**
Television Images.—T. G. Crookes. (*Nature, Lond.*, 18th May 1957, Vol. 179, No. 4568, pp. 1024–1025.) A viewer 6 ft away from a screen 10½ in. × 8 in. rolls his eyes rapidly first in a horizontal and then in a vertical arc. Pictures are seen, which, although displaced from their place of origin, are first images not after-images. They are due to the method by which a television image is formed on the c.r.-tube screen, and are not seen with cinematograph or still pictures.

621.397.5 : 621.396.4 **3312**
Study of Multichannel Sound Transmission from a Single Transmitter : Application to Bilingual Television.—L. Bourassin. (*Électronique, Paris*, Jan. 1957, No. 122, pp. 15–20.) Two methods of modulation suitable for French standards are discussed. See also 2605 of August (Pouzols) and 2300 of July (Dubec).

621.397.5 : 621.396.4 **3313**
Simple Device for Sound Reception of the Bilingual Television System of Algeria.—P. Rogues. (*TSF et TV*, Jan. 1957, Vol. 57, No. 7, pp. 16–17.) Brief outline of system. See also 3312 above and *Télévision*, Jan. 1957, No. 70, pp. 22–23.

621.397.6 : 621.395.625.3 **3314**
Picture-Synchronized Magnetic Sound Recording in Television.—H. Vollmer. (*Elektronische Rundschau*, Feb. 1957, Vol. 11, No. 2, pp. 33–37.) Problems arising from the use of a pilot frequency on magnetic recording tape are discussed, and some suitable equipment is described.

621.397.611 **3315**
Slow-Scan Adapter for Conventional TV Signals.—S. K. Altes & H. E. Reed. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 153–155.) A description of a converter providing a slow-scan video output signal with a bandwidth compression of 800 to 1.

621.397.621.2 **3316**
Improvements in Television Receivers : Part 1—Stabilization of the Line and Frame Deflection Circuits.—(*Electronic Applic. Bull.*, Oct. 1956, Vol. 17, No. 1, pp. 12–25.)

621.397.621.2 : 535.623 : 621.385.832 **3317**
Deflection and Focusing in Local [post-deflection] Spot Position Control for Electron Beams.—U. Pellegrini. (*Alta Frequenza*, Feb. 1957, Vol. 26, No. 1, pp. 25–40.) Further investigation of the two systems previously discussed (2616 of August). For the Lawrence grid system the deflection and focusing characteristics are determined separately by an approximation method. For the other system the differential equations of electron motion are solved numerically to obtain the deflection sensitivity.

621.397.7 : 621.396.67.029.62 **3318**
The Crystal Palace Television Transmitting Station.—F. C. McLean, A. N. Thomas & R. A. Rowden. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 392–393.) Discussion on 3895 of 1956.

621.397.7.029.63/.64 : 621.396.65 **3319**
Study of the Television Link Monte Generoso–Monte Ceneri [Switzerland].—F. Grandchamp. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, Jan. 1957, Vol. 35, No. 1, pp. 14–22. In French & Italian.)

621.397.8 **3320**
The Problem of Assessing Picture Quality, particularly of Television Pictures.—W. Kroebel & F. Below. (*Rundfunktech. Mitt.*, Feb. 1957, Vol. 1, No. 1, pp. 2–6.) On the basis of amount of detail recognizable, as a function of object size with the contrast between object and background as a parameter, the assessment of quality can be made independent of aesthetic considerations or picture content. For descriptions of such methods, see 3321 and 3322 below.

621.397.8 **3321**
An Objective Method of Determining Television Picture Quality.—F. Below, W. Kroebel & H. Springer. (*Rundfunktech. Mitt.*, Feb. 1957, Vol. 1, No. 1, pp. 7–11.) The effect of contrast and other objectively measurable parameters on the recognizability of picture detail is discussed. A method of determining picture sharpness is outlined but requires experimental confirmation. 22 references.

621.397.8 **3322**
Measurements for the Investigation of Picture Quality of Television Systems in the Case of Moving Objects.—F. Arp & H. Baurmeister. (*Rundfunktech. Mitt.*, Feb. 1957, Vol. 1, No. 1, pp. 12–16.) A method is described of objectively testing an observer's recognition of moving detail in a television picture. The test picture of moving circular disks is produced by projecting an optical or television image of black spheres rolling over an inclined illuminated screen. Results are shown in graphical form and discussed.

621.397.8 : 535.61 **3323**
Contrast and Grey Scale in the Television Image.—R. Suhrmann. (*Elektronische Rundschau*, Feb. 1957, Vol. 11, No. 2, pp. 43–46.) The influence of room lighting on contrast and gradation is examined ; the effectiveness of compensating gradation losses by an increase in control voltage or brightness was checked by subjective tests.

621.397.813 **3324**

Delay Distortions in Vestigial-Sideband Television Transmission and their Elimination.—F. Kirschstein. (*Hochfrequenztech. u. Elektroakust.*, Jan. 1957, Vol. 65, No. 4, pp. 119–126.) The effects of phase-delay errors on the quality of received pictures is investigated. The merits of remedies such as phase pre-correction at the transmitter [see also 584 of 1956 (Peters)] and phase-linear receivers [2235 of 1956 (van Weel)] are discussed.

621.397.5 **3325**

Television Engineering Principles and Practice: Vol. 3—Waveform Generation. [Book Review]—S. W. Amos & D. C. Birkinshaw. Publishers: Iliffe, London, 1957, 226 pp., 30s. (*Electronic Engng*, June 1957, Vol. 29, No. 352, p. 301.)

TRANSMISSION

621.396.61.029.62 **3326**

Radio Transmitter for Ionospheric Scatter.—J. L. Hollis, W. H. Collins & A. R. Schmidt. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 144–146.) The transmitter operates in the range 30–65 Mc/s with an output power of 60 kW. A neutralized triode driver operates as a linear amplifier to deliver 8–12 kW to an 8-valve grounded-grid power amplifier.

VALVES AND THERMIONICS

621.314.63 **3327**

Exact Current/Voltage Relation for the Metal/Insulator/Metal Junction with a Simple Model for Trapping of Charge Carriers.—G. H. Suits. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 454–458.)

621.314.63 : 546.289 **3328**

Very-Narrow-Base Diode.—R. H. Rediker & D. E. Sawyer. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 944–953.) The method of construction of a Ge planar-alloy-junction diode, having an active base width of the order of microns, is described. The theoretical treatment includes the effect of the non-ideal 'ohmic' contact, which is essential to its performance as a rectifier, but limits its high-frequency performance.

621.314.63+621.314.7] : 621.396.822 **3329**

Theory of Shot Noise in Junction Diodes and Junction Transistors.—A. van der Ziel. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1011.) Note of correction and comment on 600 of 1956.

621.314.632 : 537.311.33 : 546.289 **3330**

On the Current/Voltage Characteristics of Metal/Germanium Rectifying Contacts.—G. Mesnard & A. Dolce. (*C. R. Acad. Sci., Paris*, 8th April 1957, Vol. 244, No. 15, pp. 2025–2028.) A rapid increase of current for high reverse voltages is deduced theoretically from known characteristics of *n*-type Ge.

621.314.632 : 537.311.33 : 546.289 **3331**

The Variation of the Inverse Current of Metal/Germanium Rectifying Contacts when the Carrier Concentrations in the Interior of the Semiconductor Deviate from Thermal-Equilibrium Values.—G. Mesnard & A. Dolce. (*C. R. Acad. Sci., Paris*, 15th April 1957, Vol. 244, No. 16, pp. 2141–2143.)

621.314.7 **3332**

A Study of High-Speed Avalanche Transistors.—J. R. A. Beale, W. L. Stephenson & E. Wolfendale. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 394–402.) Transistors operated above a critical supply voltage appear as a 2-terminal negative resistance and give a relaxation oscillation at a much higher speed than achieved in conventional use. The design, application, and static and dynamic properties of transistors operated in this way are discussed.

621.314.7 : 537.311.33 **3333**

On the Injection of Carriers into a Depletion Layer.—Matthei & Brand. (See 3202.)

621.314.7 : 621.317.733.029.3 **3334**

Simple Transformer Bridge for the Measurement of Transistor Characteristics.—Lovering & Britten. (See 3241.)

621.314.7 : 621.385.4 **3335**

High-Frequency Circuits use Meltback Tetrodes.—D. W. Baker. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 177–179.) A description of the design and application of the tetrode-type transistor [see 677 of 1953 (Wallace et al).].

621.314.7 : 621.385.4 : 546.28 **3336**

High-Performance Silicon Tetrode Transistors.—R. F. Stewart. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1019.)

621.314.7 : 621.396.822 **3337**

Behaviour of Noise Figure in Junction Transistors.—E. G. Nielsen. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 957–963.) The work is based on a simplified version of a noise equivalent circuit developed by van der Ziel (600 of 1956). Calculations show that for minimum noise figure the ohmic base resistance and emitter current should be small, while the current gain and its cut-off frequency should be large.

621.383.27 **3338**

Fluctuations in Photomultipliers.—P. Moati. (*C. R. Acad. Sci., Paris*, 6th May 1957, Vol. 244, No. 19, pp. 2366–2368.) Previous theoretical work (2948 of September) is modified to allow for incomplete electron capture by the different dynodes.

621.383.5 **3339**

Improving the Linearity of Barrier-Layer Photocells.—D. G. Wyatt. (*J. sci. Instrum.*, March 1957, Vol. 34, No. 3, pp. 106–108.) "A modification of the Campbell-Freeth circuit is described, whereby the effective series resistance of a barrier-layer photocell, in so far as it is constant, may be eliminated. The results suggest that the series resistance includes part of the barrier layer."

621.385 (083.74) **3340**

I.R.E. Standards on Electron Tubes: Definitions of Terms, 1957.—(*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 983–1010.) Standard 57 I.R.E. 7.S2.

621.385 : 621.395.64 **3341**

Electron Tubes for the Transatlantic Cable System.—J. O. McNally, G. H. Metson, E. A. Veazie & M. F. Holmes. (*P.O. elect. Engrs' J.*, Jan. 1957, Vol. 49, No. 4, pp. 411–419.) See 1951 of June.

621.385.001.4 : 534.1 **3342**

Standardized White-Noise Tests.—Robbins. (See 3248.)

621.385.029.6 **3343**

International Congress on Microwave Valves.—(*Onde élect.*, Feb. 1957, Vol. 37, No. 359, pp. 86–193.) Further selection of papers presented at the 1956 Congress in Paris. See also 2643 of August.

621.385.029.6 **3344**

Analysis of the Energy Interchange between an Electron Stream and an Electromagnetic Wave.—V. N. Shevchik. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 104–110.) Analysis is presented of the electron bunching process in a travelling wave, which is identical with that occurring in a klystron. The maximum of the first harmonic of the current in the travelling wave was calculated and the efficiency of the interaction between the electron stream and the travelling wave was determined. The concept of synchronization was clarified. The calculations have been extended to the travelling-wave valve and the optimum small-amplitude working conditions were found for a backward-wave generator.

621.385.029.6 **3345**

High-Frequency Oscillations in Electron Beams with a Periodically Varying Velocity.—P. V. Bliokh. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 92–103.) The interaction of e.m. waves with a compensated electron beam in a longitudinal periodic electric field is examined by means of kinetic equations. Dispersion equations, derived for small signals, are used in determining the stability conditions of the electron beam.

621.385.029.6 **3346**

Barkhausen Electron Oscillations as the Basis of Velocity-Modulated Valves.—H. E. Hollmann. (*Hochfrequenztech. u. Elektroakust.*, Jan. 1957, Vol. 65, No. 4, pp. 112–119.) A phenomenological survey of the history of this type of microwave valve. Over 24 references, mainly to German literature.

621.385.029.6 **3347**

Space-Charge Effects in Beam-Type Magnetrons.—R. W. Gould. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 599–605.) A small-signal theory for magnetron-type travelling-wave valves is developed assuming a thin electron beam. The starting conditions for an M-type backward-wave oscillator are deduced. If the tube is long in space-charge wavelengths the starting current is appreciably reduced.

621.385.029.6 **3348**

Widening of Oscillation Zones of Decimetre-Waveband Magnetrons.—G. M. Gershtein & G. L. Vitel's. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 120–121.)

621.385.029.62 : 621.376.32 **3349**

A Frequency-Modulated Magnetron.—E. Petrasco & I. I. Vasilesco. (*C. R. Acad. Sci., Paris*, 29th April 1957, Vol. 244, No. 18, pp. 2296–2298.) A diagram and performance curves are shown for a special construction of magnetron with a truncated-cone anode. At 110 Mc/s a highly linear frequency excursion approaching 30 Mc/s is obtained for a 30% variation of anode voltage (950–1 350 V); a minimum voltage change of 1·5 V gives a 200-kc/s deviation with negligible amplitude modulation. Output power is about 1 W.

621.385.029.63/.64 : 621.374.4 **3350**

A Travelling-Wave Frequency Multiplier.—D. J. Bates & E. L. Ginzton. (*Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, pp. 938–944.) The valve has two helices in cascade, the output helix being of dispersive, forward-wave type with voltage tuning for selection of a particular harmonic in the range 2–4 kMc/s. Useful input range is 0·1–1 kMc/s. Multiplication ratios up to 10 or 15 are feasible with substantial gain. Output of a few milliwatts has been obtained at harmonics as high as the 40th.

621.385.029.64 **3351**

Space Harmonics of an Electron Wave.—G. A. Bernashevski. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 124–125.) Brief note on systems of the type described by Kleinwächter (2075 of 1952 and 279 of 1953).

621.385.032.2 : 537.533 **3352**

Transverse Scaling of Electron Beams.—G. Herrmann. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 474–478.) A discussion of scaling operations and their practical applications, particularly in analysing beams when thermal velocity effects are to be considered.

621.385.032.213 : 621.396.822 **3353**

On the Cause of the Anomalous Flicker Effect.—W. W. Lindemann & A. van der Ziel. (*J. appl. Phys.*, April 1957, Vol. 28, No. 4, pp. 448–451.) Experiments are described which indicate that the flicker effect is caused by sudden bursts of positive ions emitted by the cathode in a period much less than 1μs.

621.385.032.216 **3354**

The Effect of Sulphur and Oxygen on the Electrical Properties of Oxide-Coated Cathodes.—G. S. Higginson. (*Brit. J. appl. Phys.*, April 1957, Vol. 8, No. 4, pp. 148–149.) Poisoning and recovery experiments substantiate the Loosjes-Vink theory of conduction (see 3208 of 1950).

621.385.032.216.001.4 : 621.317.018.75 **3355**

A Pulse Method for Measuring the Interface Layer of Oxide Cathodes.—A. Lieb. (*Nachrichtentech. Z.*, Feb. 1957, Vol. 10, No. 2, pp. 88–89.) In the method described pulses are applied in turn to the valve under test and to a reference valve without interface layer. From a direct comparison of the resulting pulse shapes on a c.r.o. screen and the consequent adjustment of a resistor and capacitor the layer characteristics can be determined. For an earlier pulse method, see 2052 of 1951 (Eisenstein).

621.385.1 **3356**

A New Method of Investigating the Microphony of Valves: Part 2—Test Arrangements and Results.—I. P. Valkó, A. Kemény & L. Szécsi. (*Hochfrequenztech. u. Elektroakust.*, Jan. 1957, Vol. 65, No. 4, pp. 129–136.) Details of the test apparatus and procedure are given and test results are discussed and compared with those obtained by other methods. Part 1: 1981 of July (Valkó).

621.385.2 : 621.3.011.4 **3357**

The Capacitance between Diode Electrodes in the Presence of Space Charges.—C. S. Bull. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 374–378.) The capacitance of the active area of a space-charge-limited diode is zero; that of a saturated diode varies with a sudden transition from twice the cold value to the cold value as the anode potential is varied from the saturation value to the value at which space charge is negligible.

621.385.3 **3358**

The Determination of the Space-Charge Field, the Space-Charge Capacitances and the Characteristics of a Planar Triode by means of a Resistance Network with Current Sources.—G. Čremošnik & M. J. O. Strutt. (*Arch. elekt. Übertragung*, Feb. 1957, Vol. 11, No. 2, pp. 63–75.) Results obtained by means of the network analogue, particularly applied to the triode-connected valve Type EL6, are compared with results of precise calculations and with published data; close agreement is found. 37 references.

621.385.3 : 621.365.5 **3359**

The Design and Operation of High Power Triodes for Radio-Frequency Heating.—W. J. Pohl. (*Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, No. 16, pp. 410–416.) A new range of power triodes and the principles of their design.

621.385.5 **3360**

Input Conductance of a Pentode.—Yu. N. Prozorovski. (*Radiotekhnika i Elektronika*, Jan. 1957, Vol. 2, No. 1, pp. 121–123.) The effect of the heater/cathode capacitance on the input conductance is briefly discussed.

621.385.83 : 537.533.082.7 **3361**

Electron Beam Analyzer.—A. Ashkin. (*J. appl. Phys.*, May 1957, Vol. 28, No. 5, pp. 564–569.) The beam is swept across a pinhole in crossed electric and magnetic fields. The charge distribution is shown visually on a c.r. tube.

621.385.832 **3362**

A Survey of Image Storage Tubes.—H. G. Lubszynski. (*J. sci. Instrum.*, March 1957, Vol. 34, No. 3, pp. 81–89.) Half-tone storage tubes operate either with charge restoration or with charge modulation. The construction, performance, and applications of a representative selection of these tubes are discussed.

621.385.832 : 621.317.755 **3363**

A New Type of Multibeam Cathode-Ray Oscillograph.—C. Fert, J. Lagasse & J. Clot. (*C. R. Acad. Sci., Paris*, 15th April 1957, Vol. 244, No. 16, pp. 2148–2150.) A new version of the multibeam c.r.o. described in 1624 of 1954 (Fert et al.) and 2377 of 1955 (Olte). The improvements are low h.v. consumption, no beam interaction and automatic time marking.

MISCELLANEOUS

061.4 : [621.396.933+621.396.96 **3364**

French Air Show.—(*Wireless World*, Aug. 1957, Vol. 63, No. 8, pp. 368–370.) New electronic developments at the 22nd Salon International de l'Aéronautique.

538.569.2.047 **3365**

Eye Protection in Radar Fields.—W. G. Egan. (*Elect. Engng, N.Y.*, Feb. 1957, Vol. 76, No. 2, pp. 126–127.) "Tests on animals indicate that eyes exposed to microwave radiation may develop cataracts. The design of protective goggles, utilizing transparent microwave shielding, is discussed."

621.3.002.2 : 551.58 **3366**

Climatic Resistance Code of Components for Electronic Equipment.—E. Ganz. (*Bull. schweiz. elektrotech. Ver.*, 16th Feb. 1957, Vol. 48, No. 4, pp. 137–141. In French.) The system of coding developed by the Commission Électrotechnique Internationale is briefly explained with examples.

621.3.049.75 : 681.142 **3367**

Three-Dimensional Printed Wiring.—E. A. Guditz. (*Electronics*, 1st June 1957, Vol. 30, No. 6, pp. 160–163.) A new technique, using collimated light sources, for the production of etched wiring in the holes in ferrite cores of storage matrices.

621.38.019.3 **3368**

The Reliability of Electronic Equipment and Installations.—E. Ganz. (*Elektrotech. Z., Edn A.*, 11th March 1957, Vol. 78, No. 6, pp. 218–225.) The statistical results of typical life and performance tests are discussed.

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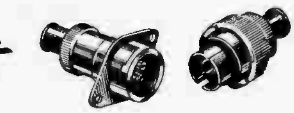

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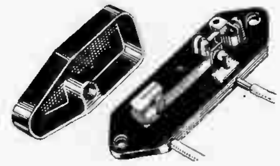

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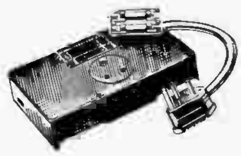

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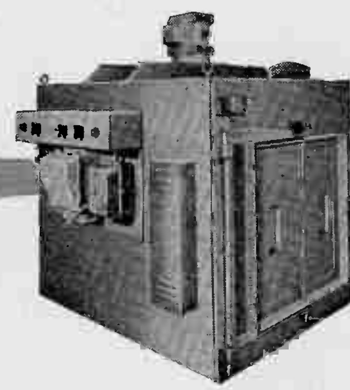

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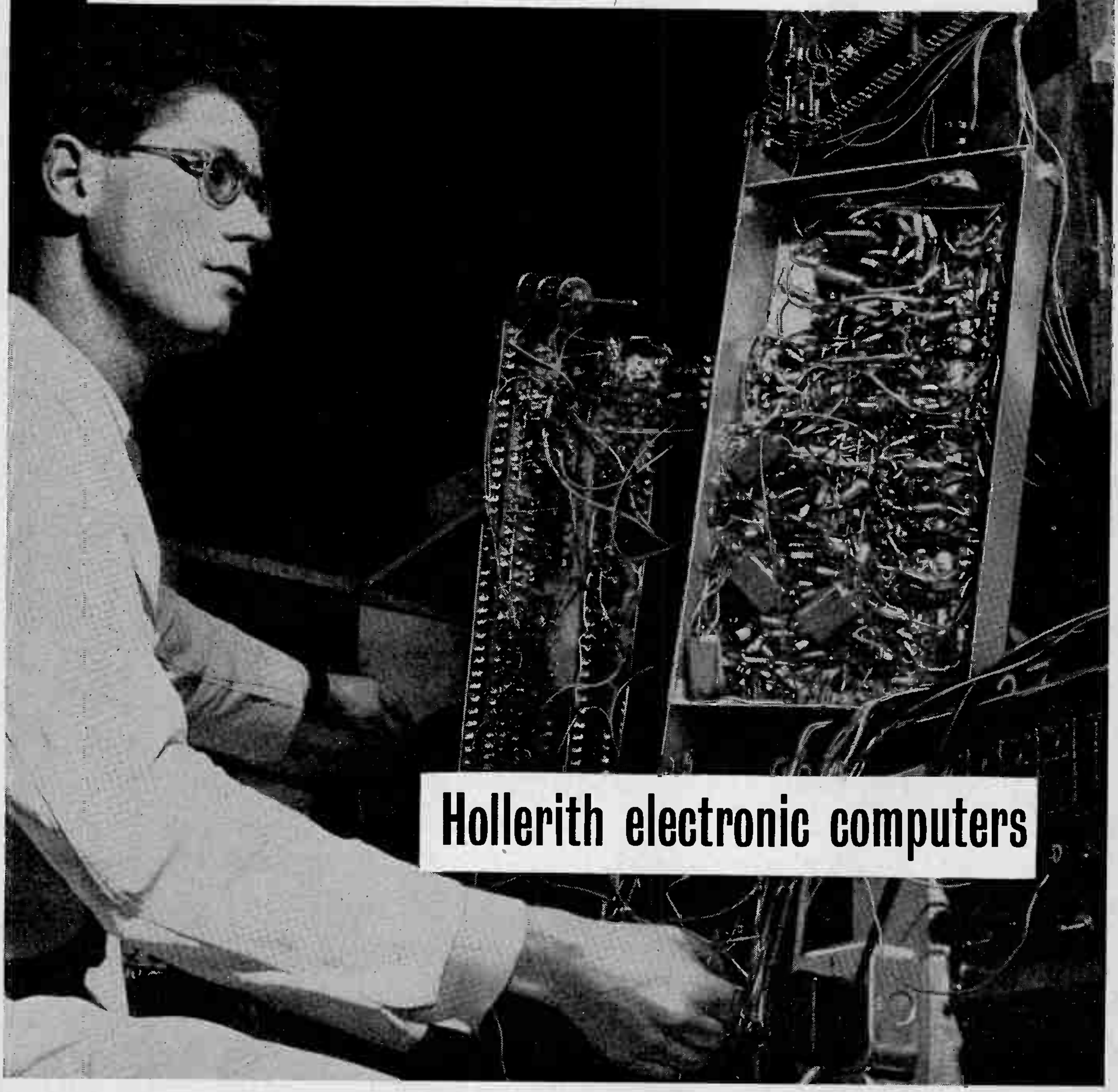

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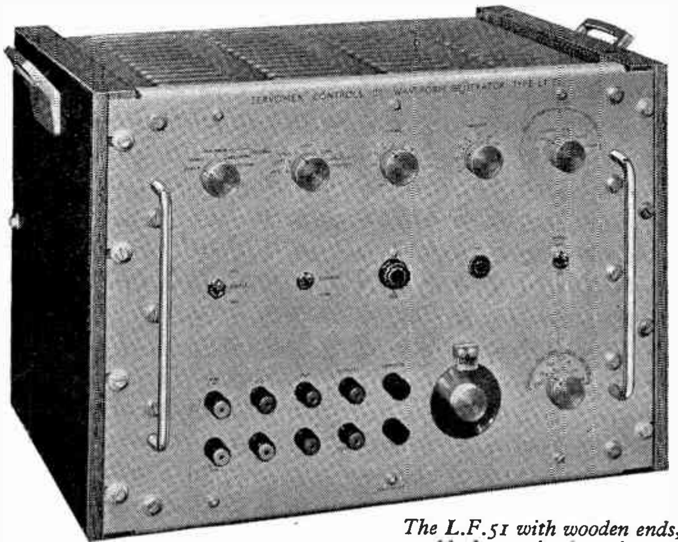

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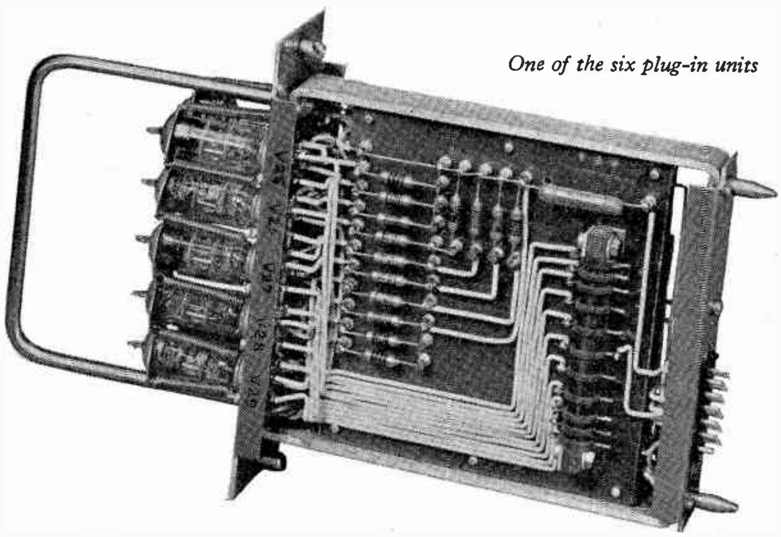

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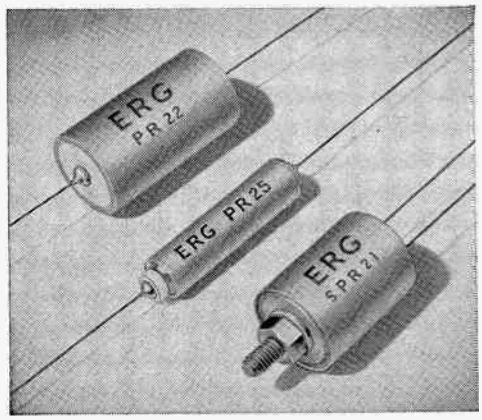

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PR 22	SPR 22	1⅜"	¾"	1·0	·65	·32	0·1	500K
PR 22A	SPR 22A	1½"	¾"	1·0	·65	·32	0·1	750K
PR 23	SPR 23	1¾"	1"	1·5	1·0	·5	0·1	2 Meg
PR 24		1⅞"	½"	1·0	·65	·32	0·1	350K
PR 25		1⅜"	⅜"	0·5	·32	·16	0·1	100K
PR 26	SPR 26	2½"	1"	2·0	1·32	·65	0·1	2·5 Meg

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Tropical Exp. ...	84 days	RCS II	·027
Climatic Cycles ...	6 Cycles	RCS II Climatic	·01
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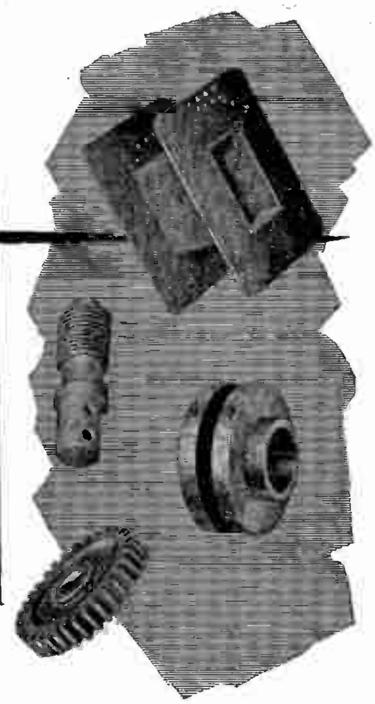

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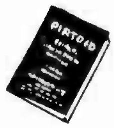

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