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THE JOURNAL OF RADIO RESEARCH & PROGRESS

APRIL 1952

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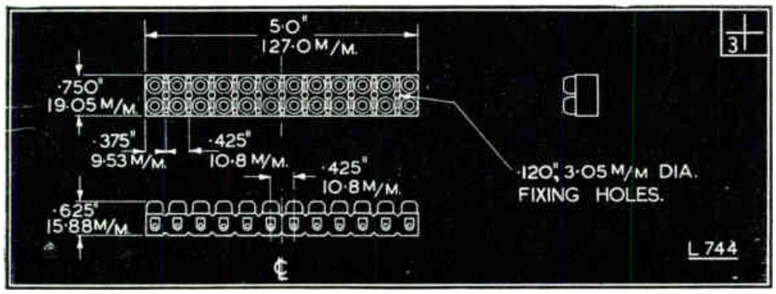

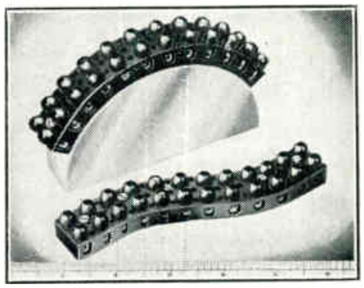

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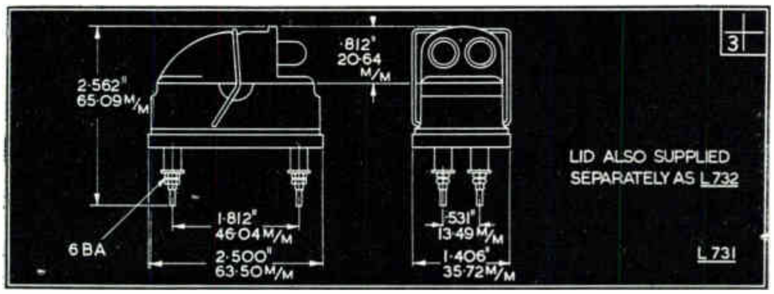

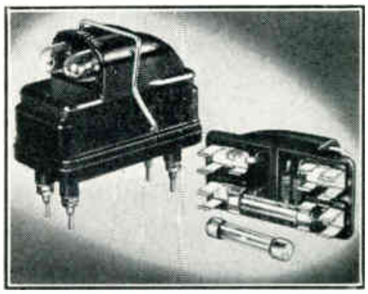

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Mechanical Force on the Short Side of a Long Rectangular Circuit

In the Editorial of November 1945, we explained a very simple method of determining this force, based on the fact that the force is transmitted from one end of the rectangle to the other by the compressional stress in the medium. If a cross-section of the line or rectangle be taken at a point not too near the end, the lines of force lie entirely in this plane and the Maxwellian stress normal to the plane is equal to $H^2/8\pi$ dynes per cm². The integral of this over the plane gives the total longitudinal force, but since the energy stored in each cubic centimetre is also equal to $H^2/8\pi$ ergs, the integral gives the energy stored in a centimetre length of the line or rectangle. But this is also equal to $0{\cdot}5\,LI^2$ ergs, where L is the inductance of unit length and I the current in absolute units. Hence, the total mechanical force on the cross-connection at the end must be equal to $0{\cdot}5\,LI^2$ whatever its shape. Since $L = 4\log_e D/r + 1$ for two parallel conductors of radius r with a distance D between centres, the end force for unit current must be equal to $2\log_e D/r + 0{\cdot}5$ dynes.

This force can be calculated in various ways, but unless this result is obtained something has gone wrong, as has often happened in the past. This is the total end force; the actual measured force on a movable cross-bar may be somewhat different from this, since it may not experience the total force. In most practical cases the accurate calculation of the force on the movable element is very complicated, especially at the corners where the distribution of current and magnetic field is very complicated.

In another Editorial of October 1945, we discussed this corner effect and described a method of avoiding it in a fictitious experiment by terminating the two long conductors with discs of infinitely-conducting material. The end connection can be pictured as a rod going either from edge to edge of the discs or from centre to centre. Since the discs are of infinite conductivity the current density in the conductors will be uniform right up to the discs, and the force on the cross-rod can be accurately calculated. As a numerical example we assumed D/r to be 60, for which $L = 17{\cdot}38$, and the total longitudinal force $8{\cdot}69\,I^2$ dynes. Assuming the cross-rod to be a fine wire and applying the Biot-Savart formula, the force on it when going from edge to edge will be $2\int_r^{59r} dx/x = 8{\cdot}15\,I^2$; when going from centre to centre this will be increased to $1 + 2\int_r^{60r} dx/x = 9{\cdot}19\,I^2$. As we pointed out, the mean of these two values is $8{\cdot}67\,I^2$ which is almost exactly the same as the value of the total longitudinal force.

We are indebted to Dr. G. F. C. Searle, of Cambridge, for the suggestion that the discs should be regarded as a part of the cross-connection and the forces on them calculated in both cases. As he pointed out, this can be done very simply even when the rod goes from edge to edge, since the force on an element of current going from A to P in Fig. 1(a) is the same as it would be if the current followed the path AOP. The magnetic flux is in the plane of the disc and the flux crossing the line AP is the same as that crossing

AOP. Considering first the radial components such as AO, let AO $= x$, then at A the magnetic field $H = Ix/r^2$, and the current crossing the circle of radius x on its way to the centre O will be $I(r^2 - x^2)/r^2$. Hence, the force on the circle of radial width dx will be $(Ix/r^2)\{I(r^2 - x^2)/r^2\}dx$, that is $I^2(x/r^2 - x^3/r^4)dx$, and the total force on the disc due to these radial currents will be the integral of this from r to O which is equal to $- I^2/4$. The minus sign means that the force is away from us towards the conductor.

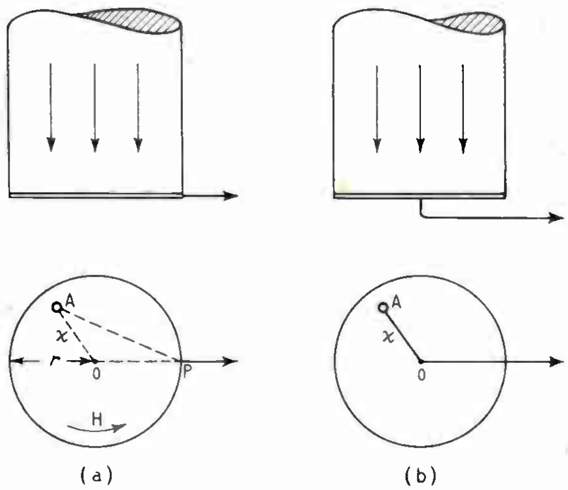

Fig. 1.

Considering now the path OP, the current along which is the whole current I, and the mean value of H on which is $I/2r$ we have for the force on OP $I(I/2r)r = I^2/2$. This force is towards us and away from the conductor. Hence the resultant force on all the paths such as AP will be $I^2/2 - I^2/4 = I^2/4$ away from the conductor.

If the cross-connection is made to the centres of the discs instead of the edges, as shown in Fig. 1(b), the component OP does not exist in the disc, and we only have the components such as AO for which we have seen that the total force is $- I^2/4$ on each disc.

When the connection is made from edge to edge each disc experiences a small force due to the magnetic field of the other conductor. In Fig. 1(a) the resultant force on all the radial components is zero, but the whole current I flows from O to P, and the magnetic field due to the other conductor will have a mean value of $(I/r) \log D/(D - r)$ giving a force on OP of $I^2 \log D/(D-r)$; which in most cases will be very small.

Summarizing our results we have for unit current:

A, with edge-to-edge connection,

$$\text{force on cross bar} = 2 \log \frac{D - r}{r}$$

$$\text{force on each disc due to main field} = 0{\cdot}25$$

force on each disc due to other field

$$= \log \frac{D}{D - r}$$

Total force

$$= 2 \log \frac{D - r}{r} + 2\left[0{\cdot}25 + \log \frac{D}{D - r}\right]$$

$$= 2 \log \frac{D}{r} + 0{\cdot}5,$$

B, with centre-to-centre connection,

$$\text{force on cross-bar} = 1 + 2 \log \frac{D}{r}$$

$$\text{force on each disc} = - 0{\cdot}25$$

$$\text{Total force} = 2 \log \frac{D}{r} + 0{\cdot}5,$$

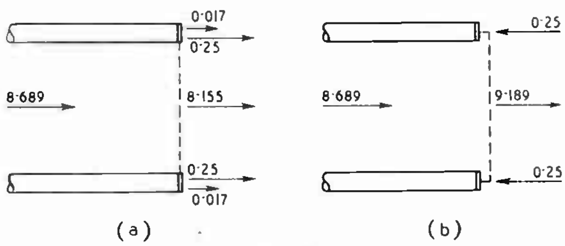

Fig. 2.

thus substantiating our statement that the total end force for unit current must be $2 \log D/r + 0{\cdot}5$ in all circumstances.

In Fig. 2 the numerical values are given for the case in which $D/r = 60$. The small component in Fig. 2(a) is that due to the field of the other conductor. The forces are given in dynes for a current of 1 absolute unit, that is, of 10 A; the inductance L is given throughout in absolute units (1 henry $= 10^9$ units).

G. W. O. H.

HIGHLY-SELECTIVE AMPLIFICATION AT LOW FREQUENCIES

By F. J. Hyde, M.Sc., Grad. I.E.E.

(Communication from the Director of Radio Research, D.S.I.R.)

SUMMARY.—The unbalanced twin-T resistance-capacitance filter, having series arms R and C and shunt arms $aR/2$ and $2C$, is analysed. The locus of the transmission vector, as a function of frequency, is given in complex form with a as parameter. For $0{\cdot}5 < a < 1$, the locus is approximately circular and the phase angle varies continuously from 0 to 2π radians. A single-stage amplifier with series feedback through a balanced twin-T filter ($a = 1$) has a Q of $A/4$, when A, the loaded stage gain without feedback, is large. Deliberate unbalancing of the filter, by making a less than unity, gives improved selectivity by virtue of the selective positive feedback introduced, and a working Q of 20 can be readily obtained. Experimental results are given for filters having resonant frequencies of approximately 0·25 c/s, 0·5 c/s and 360 c/s.

1. Introduction

THE problem of selective amplification at audio and sub-audio frequencies can be solved by using twin-T filters to provide feedback. The transmission characteristic of a balanced twin-T filter shows null-transmission at the resonant frequency, so that, if the filter is connected between grid and anode of a valve, there will be no feedback at resonance and increasingly large feedback on either side of resonance. The feedback is negative at all frequencies.

It has been found,[1] however, that a reduction of the shunt resistance arm to give an unbalanced filter, will cause the phase of the transmitted voltage to vary continuously from 0 to 2π radians with respect to the input voltage, with frequency. With the filter connected between grid and anode of a valve, it is then possible to use the selective positive feedback (in the frequency range where the real component of the transmission vector is negative) to give a considerably higher Q than with the balanced twin-T filter.

2. Transmission Characteristic of a Twin-T Filter

2.1 *Balanced Filter*

For the balanced twin-T filter, the resistance and capacitor values are as given in Fig. 1, with $a = 1$. The resonant frequency is $f_0 = 1/2\pi RC$. With $x = f/f_0$, the filter transmission is,

$$\beta = \frac{1}{1 - j\dfrac{4x}{x^2 - 1}} = |\beta| e^{j\theta} \quad .. \quad .. \quad (1)$$

where

$$|\beta| = \left\{\frac{1}{1 + \dfrac{16x^2}{(x^2-1)^2}}\right\}^{\frac{1}{2}}; \ \theta = \tan^{-1}\frac{4x}{x^2 - 1} \quad (2)$$

Plotted in the complex plane, the locus of the transmission vector is a circle, tangential to the imaginary axis at the origin, with its centre on the positive real axis. (Curve 1, Fig. 2).

For x increasing from 0 to 1, $|\beta|$ decreases from 1 to 0 and θ varies from 0 to $-\pi/2$. At $x = 1$, θ changes abruptly from $-\pi/2$ to $\pi/2$. For x increasing from 1 to ∞, $|\beta|$ increases from 0 to 1 and θ varies from $+\pi/2$ to 0.

2.2 *Unbalanced Filter*

The transmission is,

$$\beta = |\beta| e^{j\theta} = \alpha + j\gamma, \quad \text{where}$$

$$\left.\begin{aligned}
\alpha &= \frac{(1 - ax^2)\{1 - x^2(3a+2)\} + ax^2(1 - x^2)(a + 4 - ax^2)}{\{1 - x^2(3a+2)\}^2 + x^2(a + 4 - ax^2)^2} && \text{(a)}\\
\gamma &= \frac{ax(1 - x^2)\{1 - x^2(3a+2)\} - x(1 - ax^2)(a + 4 - ax^2)}{\{1 - x^2(3a+2)\}^2 + x^2(a + 4 - ax^2)^2} && \text{(b)}\\
\beta &= \left\{\frac{(1 - ax^2)^2 + a^2x^2(1 - x^2)^2}{\{1 - x^2(3a+2)\}^2 + x^2(a + 4 - ax^2)^2}\right\}^{\frac{1}{2}} && \text{(c)}\\
\theta &= \tan^{-1}\frac{2x\{x^4(a^2 + a) + x^2(a - a^2) - 2\}}{x^6a^2 + x^4(a^2 - 2a) + x^2(a^2 - 2) + 1} && \text{(d)}
\end{aligned}\right\} \quad (4)$$

The locus of the transmission vector for $a = 0{\cdot}5$ is given by curve 2 in Fig. 2. Over a limited frequency range θ lies between $\pi/2$ and $3\pi/2$. This range encloses the shaded area between the locus and the imaginary axis. Antiphase transmission for this value of a occurs at $x = 1{\cdot}2$, representing a frequency 20% higher than f_0, the frequency of

MS accepted by the Editor, June 1951

null transmission of the balanced filter. For the limiting case of $a = 0$, the locus is given by curve 3 in Fig. 2. This is tangential to the negative real axis at the origin at infinite frequency.

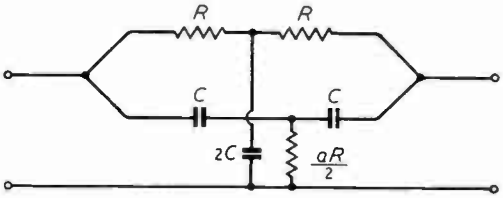

Fig. 1. *Twin-T filter.*

When the filter is connected between grid and anode of an amplifier, to provide feedback, the maximum gain corresponds to antiphase transmission ($\alpha = \alpha_\pi/\gamma = 0$) and the Q of the amplifier may be derived from the shape of the locus diagram in the antiphase region, coupled with a knowledge of the frequencies corresponding to the various points on the locus. The conditions for quadrature transmission ($\alpha = 0$, $\gamma = \gamma_{\pm\pi/2}$) are relatively easy to deduce and form a convenient reference for defining the shape of the locus and the corresponding frequency shifts. The antiphase transmission α_π, the quadrature transmissions $\gamma_{\pm\pi/2}$ and the corresponding frequency shift ratios x_π and $x_{\pm\pi/2}$ will accordingly be studied as functions of a.

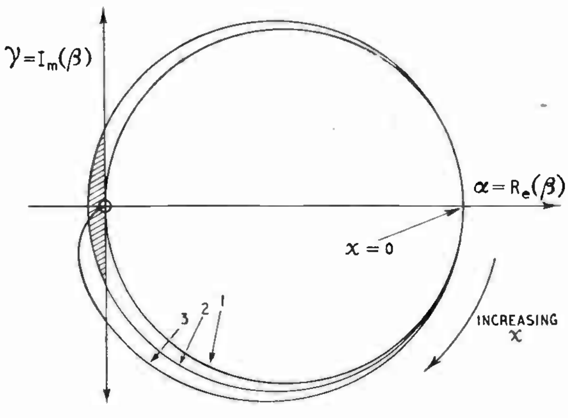

Fig. 2. *Twin-T transmission;* 1, $a = 1$; 2, $a = 0.5$; 3, $a = 0$.

2.2.1 *Antiphase Transmission*

The quadratic in x^2 in the numerator of equation 4(d) is equal to zero, so that

$$x_\pi^2 = \frac{-(a-a^2)+\sqrt{(a-a^2)^2+8(a^2+a)}}{2(a^2+a)} \quad (5)$$

From equations 4(a) and 4(b)

$$\alpha_\pi = \frac{1-ax_\pi^2}{1-x_\pi^2(3a+2)} \quad \ldots \quad \ldots \quad (6)$$

α_π is, of course, inherently negative.

2.2.2 *Quadrature Transmission*

The quadratic in a in the denominator of equation 4(d) is equal to zero, so that

$$a = \frac{x^2}{x^4+x^2+1} + \frac{\sqrt{3x^6+x^4+x^2-1}}{x(x^4+x^2+1)} \quad \ldots \quad (7)$$

where $x = x_{\pm\pi/2}$.

From equations 4(a) and 4(b)

$$\gamma_{\pm\pi/2} = \frac{ax(1-x^2)}{1-x^2(3a+2)} \quad \ldots \quad \ldots \quad (8)$$

where $x = x_{\pm\pi/2}$, and a is the corresponding value from equation (7).

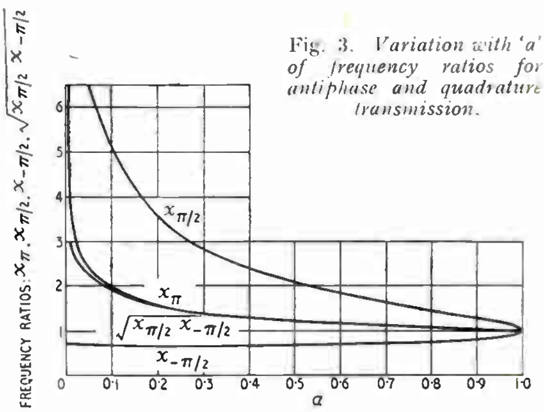

Fig. 3. *Variation with 'a' of frequency ratios for antiphase and quadrature transmission.*

2.2.3 *Conclusions*

The variations of x_π and $x_{\pm\pi/2}$ with a are shown in Fig. 3. The frequency of antiphase transmission, $x_\pi f_0$, is seen to rise with decreasing a. The geometric mean of $x_{\pi/2}$ and $x_{-\pi/2}$ is also shown in the figure, and is seen to be coincident with x_π, except at small values of a. The variation of α_π and of $\gamma_{\pm\pi/2}$ with a are shown in Fig. 4.

Consider the range $0{\cdot}5 < a < 1$. From Fig. 4 it is seen that $\gamma_{\pi/2}$ and $-\gamma_{-\pi/2}$ are virtually equal and $(\gamma_{\pm\pi/2})^2$ is approximately equal to $-\alpha_\pi$ which is a property of intersecting chords of a circle. Thus in this range, the locus of β is approximately a circle with centre on the real axis. For $a < 0{\cdot}5$, the part of the locus above the real axis becomes flatter and in the limiting case where $a = 0$, the locus lies wholly below the real axis.

3. Feedback from Anode to Grid of a Single-Stage Amplifier

Two possible arrangements, in which the feedback voltage is either in shunt or in series with the valve input voltage, are shown in Fig. 5.

3.1 *Shunt Feedback*

Both the input voltage and the feedback voltage are applied to the grid. The effect of feedback is to vary the impedance Z_{in}, presented by the amplifier to the source, which is taken to include the grid leak, and has open-circuit voltage e and internal resistance R_o. The gain is,

$$G = A\frac{Z_{in}}{Z_{in}+R_o} = \frac{A}{1+(1+A\beta)R_o/Z} \qquad (9)$$

and
$$Z_{in} = \frac{Z}{1+A\beta} \qquad (10)$$

Here,

$A = -v/v_g$ is the stage gain of the amplifier without feedback, but with the anode loaded by the filter.

$Z =$ is the impedance presented at the grid, by the filter loaded by the valve output impedance.

$G = -v/e$ is the overall gain.

β is the filter transmission.

A detailed analysis of the gain of the amplifier in terms of x and a is complicated by the fact that Z, as well as β, is complex and a function of x and a. For the case of negative feedback through a balanced filter ($a = 1$), a case often met in practice, a simple analysis is possible. This embodies the assumptions that $Z = R/(1 + j)$, which is the input impedance of a balanced twin-T filter at its null frequency with the output terminals *either* open or short-circuited, and furthermore that Z does not vary with frequency. It is shown in an appendix that the frequency of maximum response and Q are given by

$$f = f_0(1 + R/AR_o) \qquad (11)$$

$$Q = \frac{A + R/R_o}{2(2+R/R_o)} \qquad (12)$$

Q will have a maximum limiting value of $A/4$ when R_o is infinite. If R_o is made too large, however, the overall gain G [equation (9)] is reduced. If R is made small so that the ratio R/R_o becomes small, although the limiting value for Q, of $A/4$, is again approached, A itself is now reduced, because of the increased loading of the anode by the filter. For these reasons, the Q obtained in practice, with this type of circuit, is disappointingly low.

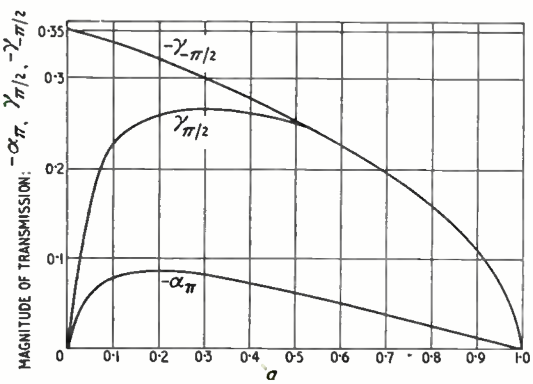

Fig. 4. *Variation with 'a' of magnitude of antiphase and quadrature transmission.*

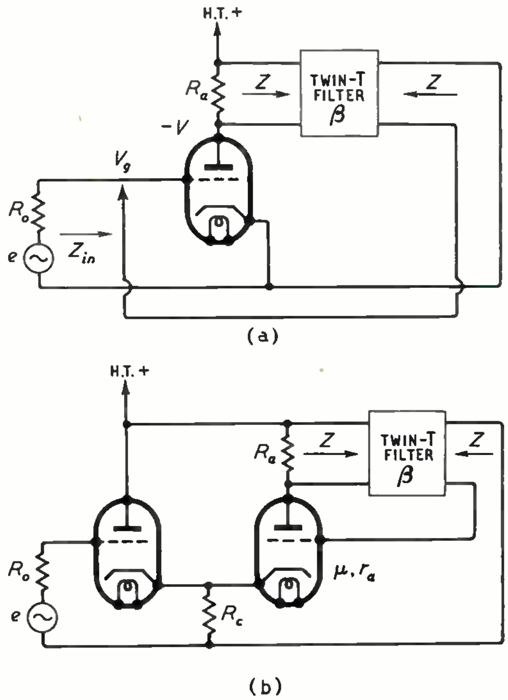

Fig. 5. (*a*) *Shunt feedback;* (*b*) *series feedback.*

3.2 *Series Feedback*

The input signal to the amplifier is injected into the cathode circuit from a cathode follower, which presents a high impedance to the source [Fig. 5(b)]. The feedback voltage is applied to the grid and is therefore in series with the input signal. The overall gain is now virtually independent of the source resistance R_o, as the cathode-follower input impedance is high. The series resistance arm of the filter, R, may be made as large as is consistent with the capacitance values of the filter being large compared with stray valve and circuit capacitances. This minimizes the loading effect of the filter on the anode circuit. With the grid leak of the amplifier on the anode side of the filter, although the stage gain is somewhat reduced by the leak being in parallel with the anode load, the voltage fed back to the grid is fully β times the anode voltage. The gain from cathode to anode of the amplifier is,

$$G = \frac{\mu+1}{\mu}\cdot\frac{A}{1+A\beta} \approx \frac{A}{1+A\beta} \qquad (13)$$

where $A = \mu R_a'/(R_a' + r_a)$ is the stage gain from grid to anode, with the anode loaded by the filter and grid leak, resulting in a composite anode load R_a'. r_a = a.c. anode resistance of the valve.

3.2.1 *Series Feedback with a Balanced Filter*

This case, when β is given by equation (1), has been treated by Fleisher,[2] who shows that Q is approximately $A/4$ when A is large.

3.2.2 *Series Feedback with an Unbalanced Filter*

In the range $0{\cdot}5 < a < 1$, it has been shown that the locus of β is approximately circular and that points on the locus, symmetrically disposed about the negative real axis, correspond to frequencies which have geometric symmetry about the resonance frequency. Thus, near resonance it may be assumed that the gain is represented by,

$$G = \frac{\hat{G}}{1 + jQu} \qquad (14)$$

where, $\hat{G}$ is the maximum value of gain, occurring at frequency $\hat{f} = x_\pi f_0$

G is the gain at frequency f

u is $f/\hat{f} - \hat{f}/f$

Putting β in the form $\alpha + j\gamma$, equation (13) may be rewritten as

$$G = \frac{A}{1 + A\alpha + jA\gamma} \qquad (15)$$

The maximum value of gain occurs when $\alpha = \alpha_\pi$ ($\gamma = 0$, $x = x_\pi$).

$$\hat{G} = \frac{A}{1 + A\alpha_\pi} \qquad (16)$$

The general expression for gain may, therefore, be written,

$$G = \frac{\hat{G}}{\dfrac{1 + A\alpha}{1 + A\alpha_\pi} + j\dfrac{A\gamma}{1 + A\alpha_\pi}} \qquad (17)$$

The presence of the factor $(1 + A\alpha)/(1 + A\alpha_\pi)$, in the denominator of equation (17), shows that equation (14) cannot accurately represent the response of the feedback amplifier and that Q does not have an exact meaning. The representation is, however, a good approximation near resonance when $(1 + A\alpha)/(1 + A\alpha_\pi) \approx 1$, and also when $|A\gamma| > |1 + A\alpha|$ (α is negative in the region of positive feedback), so that the imaginary term in the denominators of equations (14) and (17) is the dominant term. The latter condition is fulfilled at the quadrature points. A comparison of equations (14) and (17) shows that

$$Qu \approx \frac{A\gamma}{1 + A\alpha_\pi} \qquad (18)$$

Q may be derived from this expression, using the corresponding values of u and γ, calculated in Section 2.2.2, for quadrature transmission. The resulting variation of Q with a, with A as parameter, is shown in Fig. 6. It is noted that for each value of A, oscillation occurs at that value of a for which $\alpha_\pi = -1/A$. This value may be obtained from Fig. 4.

3.3 *Experimental Results*

Results are given for feedback through different twin-T filters having resonant frequencies, with the shunt arm balanced, of approximately 0·25 c/s, 0·5 c/s and 360 c/s. The circuit arrangement used is that of Fig. 5(b), with the amplifier grid leak connected on the anode side of the filter.

The variation of Q, $\hat{f}$, the frequency of maximum response, and x_π, the frequency shift ratio $\hat{f}/f_0$, are studied in terms of a. It will be noted that the largest experimentally-determined value of a at which oscillation occurs affords a measure of A, the loaded stage gain of the amplifier without feedback. In the condition where oscillation is incipient $A = -1/\alpha_\pi$ [equation (16)], the value of α_π, corresponding to the measured value of a, being given in Fig. 4.

Example 1

The valves used with two very-low frequency filters are type EF37A pentodes, the cathode-follower being triode connected. The amplifier anode load and cathode-follower anode voltage are adjusted to give measured values of 45 and 30 for A. Results for each filter are given in Tables 1 and 2.

TABLE 1

(a) *Variation of Q and frequency-shift ratio, x_π, with 'a'. Filter components:*—R = 670,000 ohms; C = 0·97 μF; f_0 = 0·247 c/s.

a	$A = 45$		$A = 30$	
	Q	$\hat{f}/f_0$	Q	$\hat{f}/f_0$
1·00	12·0	1·000	7·8	1·000
0·95	15·8	1·015	10·0	1·018
0·92	18·2	1·026	11·5	1·026
0·89	25·6	1·034	12·2	1·034
0·86	40·0	1·050	15·5	1·050
0·82	Oscillation		20·2	1·063
0·745	—	—	Oscillation	

(b) *Check on measured value of A, from $A = -1/\alpha_\pi$ for incipient oscillation.*

A (measured)	a	α_π (Fig. 4)	$A = -1/\alpha_\pi$
45	0·82	−0·023	43·5
30	0·745	−0·0325	30·8

TABLE 2

(a) *Variation of Q and frequency shift ratio, x_π, with 'a'. Filter components:—R = 670,000 ohms, C = 0·495 μF; f_0 = 0·48 c/s.*

a	$A = 45$		$A = 30$	
	Q	$\hat{f}/f_0$	Q	$\hat{f}/f_0$
1·00	11·0	1·000	8·2	1·000
0·95	16·0	1·010	10·5	1·011
0·92	20·7	1·015	11·5	1·021
0·89	24·6	1·025	14·0	1·030
0·86	48	1·037	16·0	1·040
0·83	Oscillation		21·0	1·045
0·75	—	—	Oscillation	

(b) *Check on measured value of A, from $A = -1/x_\pi$ for incipient oscillation.*

A (measured)	a	α_π (Fig. 4)	$A = -1/\alpha_\pi$
45	0·83	− 0·022	45·5
30	0·75	− 0·032	31·2

Example 2

A type ECC32 double triode valve, with operating conditions giving a measured value of A of 18, is used with a filter having f_0 = 363 c/s. Results are given in Table 3.

TABLE 3

(a) *Variation of Q and frequency shift ratio, x_π, with 'a'. Filter components:—R = 92,000 ohms, C = 0·477 μF; f_0 = 363 c/s.*

Loaded amplifier stage gain A = 18

a	Q	$\hat{f}/f_0$
1·0	4·6	1·000
0·776	9·6	1·075
0·71	14·7	1·095
0·601	30·0	1·16
0·54	Oscillation	

(b) *Check on measured value of A, from $A = -1/\alpha_\pi$ for incipient oscillation.*

A (measured)	a	α_π (Fig. 4)	$A = -1/\alpha_\pi$
18	0·54	− 0·059	17

The experimental points are found to show good agreement with the theoretically-determined curves relating Q and x_π to a (Figs. 6 and 3 respectively).

4. Conclusions

It is seen that accurate matching of components of a twin-T filter, to give a sharp null in the transmission is unnecessary, since it will not result in the maximum possible selectivity, when the filter is used as a feedback network, in a single-stage amplifier. The selective positive feedback provided by the filter, when unbalanced by reduction of its shunt resistance arm, enables working values of Q up to 20 to be readily obtained. The increase in Q is accompanied by a rise in the frequency of maximum response. If a certain value of Q is required at a particular central frequency, with a given value of A, the twin-T components may be determined, using first Fig. 6, to give the required value of a, and then Fig. 3 to give the corresponding value of x_π, from which f_0 may be determined. R and C are then calculated from $f_0 = 1/2\pi RC$.

The value of Q obtainable, for a single-stage amplifier, with series negative feedback through a balanced twin-T filter, is usually quoted as $A/4$ when A is large. It is seen that this is an upper limiting value in the case of shunt feedback.

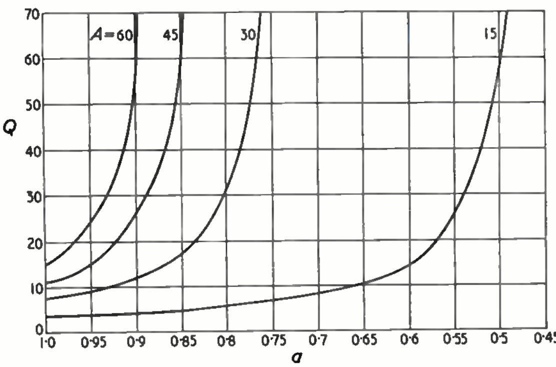

Fig. 6. *Variation of Q with 'a'.*

Acknowledgments

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APPENDIX

SHUNT NEGATIVE FEEDBACK THROUGH A BALANCED FILTER

The overall gain is,

$$G = \frac{A}{1 + (1 + A\beta)\, R_0/Z} \quad \text{(i)}$$

DISTORTION IN BEAT-FREQUENCY SOURCES

Effect of 'Pulling' on Waveform

By C. G. Mayo, M.A., B.Sc., M.I.E.E.

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SUMMARY.—If in a beat-frequency tone source there is stray coupling between the two oscillators there is a tendency for the oscillators to lock into step at very low beat frequencies. This is generally termed 'pulling'. At higher beat frequencies than the frequency at which the oscillators lock there is distortion of the beat-frequency waveform. It will be shown that there is a simple relation between the second-harmonic distortion at any frequency $\omega/2\pi$ and the highest beat-frequency setting $\omega_0/2\pi$ at which the oscillators lock. The relation is: Percentage second harmonic distortion $= 100\ \omega_0/2\omega$.

General

THE general solution of oscillators and their interaction has been treated by van der Pol.[1] His method of attack is from the differential equation

$$v'' - \alpha(1 - v^2)v' + \omega^2 v = 0 \qquad \left(\frac{\alpha}{\omega} \ll 1\right)$$

(If $\alpha/\omega \ll 1$ the solution is approximately sinusoidal whereas if it is large the solution is more like that of a multivibrator.)

In this non-linear equation the so-called damping term is $-\alpha(1 - v^2)\,v'$ and it is due to this term that a stable oscillation at a definite level is possible. In general approximate terms, if $v^2 > 1$ this term becomes positive and the resultant oscillation tends to die away, whereas if $v^2 < 1$ the oscillation tends to build up. This term can be taken as due to specific non-linearity in the characteristic of the driving valve or as the effect of grid current when the voltage exceeds a certain value.

In certain cases there may be in addition an automatic regulator of a different type. For example, the bias of the driving valve may be controlled by the output in such a way as to reduce the mutual conductance of the valve when the output exceeds a definite value. This control may involve time delay and serve to set the operating level, but in general the instantaneous stability is due to the van der Pol term $-\alpha(1 - v^2)v'$.

This non-linear equation is solved by van der Pol for various cases including that of oscillators with two degrees of freedom, and also that of an oscillator with an external drive of fixed frequency. In the latter case the external drive tends to suppress the oscillator's own frequency to an increasing extent as the frequencies are more nearly equal.

In the present case an approximate solution is obtained, not from the differential equation in which the regulating effect is explicit in the term $-\alpha(1 - v^2)v'$, but by taking account of the regulator effect at a later stage of the analysis.

MS accepted by the Editor, June 1951

Analysis

Suppose the oscillator voltages are $e_1 \cos st$ and $e_2 \cos (s - \omega)t$ for oscillators A and B respectively where $\omega \ll s$ and $\omega/2\pi$ is the beat-frequency.

Suppose the coupling between the oscillators to be represented by mutual admittances

$$g + jb = ye^{j\phi}$$

and

$$g' + jb' = y'e^{j\phi'}$$

where y represents the magnitude of the current flowing in the tuned circuit of oscillator A due to unit voltage on oscillator B and ϕ is the phase angle of such current relative to the phase of the unit voltage. y' and ϕ' are the corresponding quantities for the effect of oscillator A on oscillator B.

It is now assumed that oscillator B can be considered as oscillating at a frequency $s/2\pi$ with a small rate of change of phase $-\omega$.

This premise is justifiable on the grounds that the principal effect of the small difference frequency $\omega/2\pi$ is in fact the cumulative change of phase due to it. Owing to the fact that $\omega \ll s$ impedances in terms of s or $s - \omega$ differ only by a very small quantity. On this basis the pulling current in oscillator A due to oscillator B is

$$i_p = e_2 y e^{j\phi} e^{-j\omega t}$$

$$= e_1 \frac{e_2}{e_1} y e^{j\phi} e^{-j\omega t}$$

i.e., as if due to a self admittance $\frac{e_2}{e_1} y e^{j\phi} e^{-j\omega t}$ across the oscillator tuned circuit. This admittance behaves as a positive or negative conductance, or positive or negative susceptance in accordance with the value of t.

The inphase component of i_p tends to increase or decrease e_1 without affecting the frequency and the tendency will be counterbalanced by the non-linear elements in the oscillator. The quadrature components of i_p will tend to increase or decrease the frequency of oscillator A. This is the effect known as 'pulling'.

Now $\frac{e_2}{e_1} y e^{j\phi} e^{-j\omega t}$

$$= De^{j(\theta - \pi/2)} = D \sin \theta + jD \cos \theta$$

where $D = \frac{e_2}{e_1} y$ and $\theta = \phi - \omega t + \frac{\pi}{2}$

that is, θ differs from the actual phase angle between the oscillators only by a constant.

The quadrature component of i_p is thus

$$D \cos \theta$$

Hence the change of frequency of oscillator A due to the mutual coupling is

$$\delta_1(s) = \delta_1(d\theta/dt) = K_1 \cos \theta$$

Similarly the effect of oscillator A on oscillator B is to cause a change of the frequency of oscillator B and thus of $d\theta/dt$

$$\delta_2(d\theta/dt) = K_2 \cos \theta$$

Thus the total change of $d\theta/dt$ due to both couplings can be simply written

$$\delta(d\theta/dt) = K \cos \theta$$

or $$d\theta/dt = \omega + K \cos \theta$$

That is, $(d\theta/dt)$, the actual rate of change of θ or the so-called instantaneous frequency differs from the beat-frequency ω in the absence of pulling by a term $K \cos \theta$ where K depends on the magnitude of the coupling between the oscillators.

Now suppose $\omega < K$

Then for some value of θ, $\omega + K \cos \theta_1 = 0$ so that for values of θ near θ_1, say $\theta = \theta_1 + \delta\theta_1$

$$\frac{d\theta}{dt} \approx - K \sin \theta_1 \, \delta\theta$$

Thus if $\sin \theta_1$ is positive the condition is stable, since small deviations θ are concomitant with restoring forces. If $\sin \theta_1$ is negative the condition is unstable. The stable condition means that the oscillators lock at a phase difference θ_1. Although there are two values of θ_1 which make $\omega + K \cos \theta = \theta$ only one is a stable locking phase. As $\omega \to K$ the two roots of $\omega + K \cos \theta$ tend towards $\theta_1 = \pi$ and the condition is one of neutral equilibrium when $\omega = K$.

Thus if ω_0 is the highest value of ω (i.e., the highest difference frequency at which the oscillators lock),

$$d\theta/dt = \omega + \omega_0 \cos \theta$$

The solution of this differential equation gives the required result, but before going on to the solution it is useful to derive the equation in a different manner.

The oscillator circuit may be considered as a tuned circuit with loss, together with a maintaining amplifier which exactly supplies the loss at some specific amplitude. It may equally well be considered as a loss-free tuned circuit with a regulating amplifier which supplies positive or negative loss if the amplitude varies from a datum.

Such a loss-free impedance may be written for oscillator A as

$$Z = Z_0 ps/(p^2 + s^2)$$

where $p = j\omega$

This represents a loss-free tuned circuit of angular frequency s. Z_0 is a constant.

Suppose the voltage without stray coupling is

$$e = e_1 \sin st$$

Suppose the current in this circuit due to oscillator B is

$$i_1 = \frac{2\lambda}{Z_0} e_1' \sin \{(s - \omega)t + \theta\}$$

where $2\lambda/Z_0$ is the numerical value of the mutual admittance from oscillator B to oscillator A and θ is the effective relative phase.

Using well-known methods the total resultant voltage on oscillator A is

$$e_2 = e_1 \sin st + \frac{2\lambda}{\omega} e_1' \Big[\sin \{(s - \omega)\, t + \theta\} - \sin (st + \theta) \Big]$$

$$= e_1 \Big[\sin st + \frac{\mu}{\omega} \Big\{ \sin \{(s - \omega)\, t + \theta\} - \sin (st + \theta) \Big\} \Big]$$

where μ is a coupling constant.

It is seen that the stray coupling produces no immediate change of either phase or amplitude since when $t = 0$, $e_2 = e_1 \sin \omega t$.

Now e_2 may be written

$$e_2 = R \sin (st + \psi)$$

where $$R^2 = e_1^2 \left[1 + \frac{\mu}{\omega} \cos (\omega t - \theta) - \frac{\mu}{\omega} \cos \theta \right]^2 + e_1^2 \left[\frac{\mu}{\omega} \sin (\omega t - \theta) + \frac{\mu}{\omega} \sin \theta \right]^2$$

and $$\tan \psi = - \frac{\frac{\mu}{\omega} [\sin (\omega t - \theta) + \sin \theta]}{1 + \frac{\mu}{\omega} \cos (\omega t - \theta) - \frac{\mu}{\omega} \cos \theta}$$

In the absence of the regulator this would be the solution valid for all time. As there is no non-linearity there is no distortion as yet. The voltage e_2 is, in fact, merely the sum of two voltages of different frequencies and may be considered as a carrier and a single sideband. If ψ were constant e_2 would be an amplitude-modulated carrier of angular frequency s and if R were constant e_2 would be phase modulated.

The effect of the regulator is, however, to maintain R constant without effect on ψ_1 since the operation of the regulator is in effect to insert positive or negative conductance into the tuned circuit as required to maintain R constant.

Since the regulator has no direct effect on ψ the formula

$$\tan\psi = -\frac{\frac{\mu}{\omega}[\sin(\omega t - \theta) + \sin\theta]}{1 + \frac{\mu}{\omega}\cos(\omega t - \theta) - \frac{\mu}{\omega}\cos\theta}$$

gives the correct value of ψ and $d\psi/dt$ at time $t = 0$. At any later times the régime will be different owing to the effect of the regulator.

From this equation

$$\left.\frac{d\psi}{dt}\right]_{t=0} = -\mu\cos\theta$$

i.e., the increment of the rate of change of θ due to the effect of oscillator B on oscillator A is

$$\delta_1\frac{d\theta}{dt} = -\mu\cos\theta \text{ as before.}$$

Thus the differential equation to be solved is

$$\frac{d\theta}{dt} = \omega + \omega_0\cos\theta$$

and $$dt = \frac{d\theta}{\omega + \omega_0\cos\theta}$$

or $$\tan\frac{\theta}{2} = \sqrt{\left(\frac{\omega - \omega_0}{\omega + \omega_0}\right)}\tan\sqrt{(\omega^2 - \omega_0^2)}\,t/2$$

and $$\sin\theta = \frac{2\tan\theta/2}{1 + \tan^2\theta/2}$$

therefore $$\sin\theta = \sqrt{\left(1 - \frac{\omega_0^2}{\omega^2}\right)}\frac{\sin\sqrt{(\omega^2 - \omega_0^2)}t}{1 - \frac{\omega_0}{\omega}\cos\sqrt{(\omega^2 - \omega_0^2)}t}$$

$$= \sqrt{1 - \frac{\omega_0^2}{\omega^2}}\left[\sin\omega t\sqrt{1 - \frac{\omega_0^2}{\omega^2}} + \frac{\omega_0}{2\omega}\sin 2\omega t\sqrt{1 - \frac{\omega_0^2}{\omega^2}} + \ldots\right]$$

Thus (1) The beat frequency is reduced by pulling-in the ratio $\sqrt{(1 - \omega_0^2/\omega^2)}$ and locking occurs when $\omega \approx \omega_0$.

(2) The ratio of second harmonic to fundamental is $\omega_0/2\omega$.

(3) Higher harmonics are of higher order in ω_0/ω; i.e., if the second harmonic is 1%, the third harmonic will be about 0·01%. Thus the distortion at 50 c/s due to a locking frequency of 1 c/s is 1%.

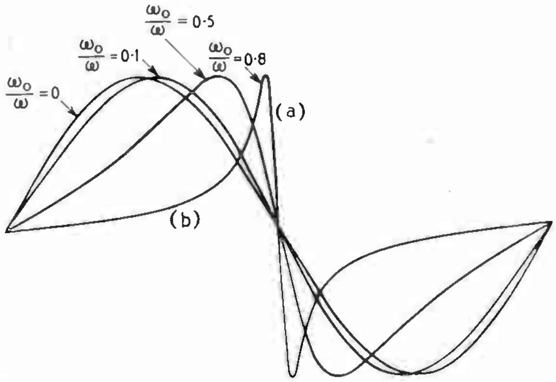

Fig. 1. Effect of pulling for various values of ω_0/ω. In each case one complete cycle of the waveform is shown. Over the part of the cycle marked (a) the stray coupling tends to increase the relative frequency of the oscillators, whereas over the part marked (b) it tends to decrease it.

In Fig. 1 are drawn the actual waveforms for various values of ω_0/ω. It should be observed that this non-linear distortion is introduced by the non-linear nature of the oscillator regulators which have here been assumed to maintain the oscillator voltages constant. There are two general types of regulator, one of which operates like grid current, directly to increase the oscillator circuit loss. The second type operates by changing the régime over a comparatively long period of time. Such regulators may for example operate by 'thermistor' action or by change of bias voltage on a variable-μ valve, etc. If the time constant of such regulators is sufficiently long to prevent the regulator from maintaining oscillator peak voltages constant, then the distortion introduced is less.

Acknowledgments

The foregoing analysis was undertaken in reference to a portable tone source developed for use in room acoustics by the B.B.C. Engineering Research Department. An account of this tone source has already been published.[2]

The author is indebted to the Chief Engineer of the B.B.C. for permission to publish this article.

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RESONANT CIRCUIT WITH PERIODICALLY-VARYING PARAMETERS

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SUMMARY.—The circuit with periodically-varying resistance is studied both experimentally and theoretically.

Experimentally the periodic variation of the resistance is achieved by applying a voltage of a given frequency to the grid of a dynatron. This varies periodically the slope of the $i_a - v_a$ characteristic, which determines the magnitude of the negative resistance which the dynatron presents to the oscillatory circuit.

Two separate cases are considered: (1) the steady-state response of the circuit to an applied alternating e.m.f. of approximately the resonant frequency; (2) the oscillatory regime, when oscillations are sustained in the circuit by the periodic variation of the resistance.

The steady-state response exhibits a phenomenon of multiple resonance, provided the frequency of the voltage applied to the grid is much smaller than the resonant frequency of the circuit. A multiple resonance shows several maxima, instead of a single one as observed for constant parameters. Each maximum corresponds to detuning equal to an integral multiple of the frequency of the grid voltage; i.e., the frequency of the resistance variation.

The resistive case of parametrically-excited oscillations completes the means of generating these oscillations, since inductive and capacitive oscillations have been achieved previously.

As in the inductive and capacitive case the excitation frequency at which the oscillations are most easily excited is double the resonant frequency of the circuit. If the alternating voltage applied to the grid is increased from zero, a certain critical value is reached at which the circuit bursts into oscillations, the frequency of which is exactly half of the frequency of the grid voltage, even if the circuit is detuned.

Theoretically the problem consists of the solution of Mathieu's and Hill's equations. In Part 1 the steady state solution of Mathieu's equation is obtained by means of integral equations, while in Part 2 a solution of an extended Hill's equation is obtained by a method similar to that of E. L. Ince.

Part 1. Multiple-Resonance Phenomena

SYMBOLS

R_o = time constant resistance of the circuit.
R_1 = amplitude of the time variant resistance.
R_n = amplitude of the nth harmonic of the time variant resistance.
$\chi_o = \dfrac{R_o}{2L}$ = natural decrement of the circuit.
$\delta_1 = \dfrac{R_1}{2\omega_1 L}$ = time-variant decrement of the circuit.
$\chi(t) = \chi_o t + \delta_1(t)$ = decrement of the circuit.
ω_o = resonant angular frequency of the circuit.
ω_1 = angular frequency of the resistance variation; i.e., of the voltage applied to the grid of the dynatron.
ω = angular frequency of the driving e.m.f.
$p = \omega - \omega_o$ = angular frequency detuning.
E = amplitude of the driving e.m.f.
V_g = voltage applied to the grid.
V_c = voltage across the capacitor of the resonant circuit.
V_n = amplitude of the nth harmonic of the envelope waveform of V_c.
A = mean amplitude of V_c.
A_o = amplitude of V_c for $R_1 = 0$.
$G = \left|\dfrac{A}{A_o}\right|$ = amplification factor due to resistance variation.

Introduction

A SIMPLE resonant circuit consists of three parameters: inductance, capacitance, and resistance.

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The first two are the energy-storing elements and the third a dissipative element, but its dissipative character is changed if the resistance of the circuit becomes negative during parts of a cycle.

The behaviour of the circuit when its parameters are constant is well known. Recently more attention has been paid to the behaviour of the circuit when its parameters are functions of time, especially when they vary periodically. Frequency modulation, for instance, utilizes variation of capacitance or, to a lesser degree, inductance. Introduction of super-regeneration stimulated an interest in periodic variation of resistance, while in the parametric excitation of oscillations, such oscillations are excited simply by varying one of the parameters periodically at a frequency approximately twice the resonant frequency of the circuit.

The theoretical basis of these studies rests on the differential equations of Mathieu and Hill:

$$\frac{d^2y}{dt^2} + (a - 2q\cos 2z)\,y = F(z)$$

$$\frac{d^2y}{dt^2} + [a - 2q\psi(2z)]\,y = F(z)$$

The Hill's equation is a more general one, since it contains Mathieu's equation as a particular case, viz., $\psi(2z) = \cos 2z$. $\psi(2z)$ is frequently written in a series form $\sum_n \theta_n \cos 2nz$.

The steady-state forced oscillations correspond to a particular integral solution, which gives an expression for the multiple-resonance phenomenon observed with super-regenerative circuits.

A solution for the oscillatory regimes (parametric oscillation) is obtained for $F(z) = 0$ (complementary function), and an unstable solution is sought; i.e., a solution for which $y \to \pm\infty$ as $z \to \pm\infty$.

2. Derivation of the Differential Equation

The case of a periodically-varying resistance will be considered, but a similar procedure can be applied to the cases where L or C vary.

Considering the series-resonant circuit of Fig. 1 we have

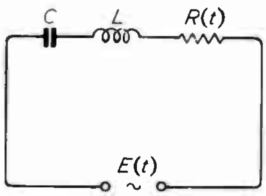

Fig. 1. *A series resonant circuit.*

$$L\frac{di}{dt} + iR(t) + \frac{1}{C}\int^t i\,dt = E(t) \quad .. \quad .. \quad (1)$$

where $E(t) = E\sin\omega t$ and $R(t) = R_0 + R_1\cos\omega_1 t$ and

$$\frac{d^2q}{dt^2} + \frac{R(t)}{L}\frac{dq}{dt} + \frac{q}{LC} = \frac{1}{L}E(t) \quad .. \quad .. \quad (2)$$

Putting $q = e^{-\chi(t)}u(t)$, where $\chi(t) = \int_0^t \frac{R(\tau)}{2L}d\tau = \chi_0 t + \delta_1(t)$ and since the $\frac{du(t)}{dt}$ terms cancel out, we have

$$\frac{d^2u}{dt^2} + \left[\frac{1}{LC} - \frac{R^2(t)}{4L^2} - \frac{1}{2L}\frac{dR(t)}{dt}\right]u = \frac{1}{L}E(t)\,e^{\chi(t)} \quad .. \quad .. \quad .. \quad (3)$$

This holds for any variation of R, provided $\chi(t)$ converges.

3. Steady-State Solution

Equation (3) can be written

$$\frac{d^2u}{dt^2} + \omega_0^2 u = 2k\psi(\omega_1 t)u + f(t) \quad .. \quad .. \quad (4)$$

where

$$\omega_0^2 = \frac{1}{LC} - \frac{R_0^2}{4L^2},$$

$$2k\psi(\omega_1 t) = \frac{R^2(t) - R_0^2}{4L^2} + \frac{1}{2L}\frac{dR(t)}{dt}$$

and $f(t) = \frac{1}{L}E(t)\,e^{\chi(t)}$

Multiplying both sides of (4) by $\sin\omega_0 t$ and integrating by parts we get

$$\frac{du(t)}{dt}\sin\omega_0 t - \omega_0 u(t)\cos\omega_0 t = -\omega_0 u(0) + 2k\int_0^t \psi(\omega_1\tau)\sin\omega_0\tau\, u(\tau)d\tau + \int_0^t f(\tau)\sin\omega_0\tau\, d\tau \quad .. \quad (5)$$

And similarly for $\cos\omega_0 t$ we get

$$\frac{du(t)}{dt}\cos\omega_0 t + \omega_0 u(t)\sin\omega_0 t = \frac{du}{dt}(0) + 2k\int_0^t \psi(\omega_1\tau)\cos\omega_0\tau\, u(\tau)d\tau + \int_0^\tau f(\tau)\cos\omega_0\tau\, d\tau \quad .. \quad (6)$$

Multiplying (5) by $\frac{1}{\omega_0}\cos\omega_0 t$, (6) by $\frac{1}{\omega_0}\sin\omega_0 t$ and subtracting the first from the second yields the integral equation

$$u(t) = u(o)\cos(\omega_0 t) + \frac{1}{\omega_0}\frac{du(0)}{dt}\sin\omega_0 t + \frac{2k}{\omega_0}\int_0^t \psi(\omega_1\tau)\sin\omega_0(t-\tau)\,u(\tau)d\tau + \frac{1}{\omega_0}\int_0^t f(\tau)\sin\omega_0(t-\tau)\,d\tau \quad .. \quad (7)$$

The first two terms are the transient oscillations of the circuit at its natural frequency f_0. Since we are only interested in the forced oscillations due to $E(t)$, these can be omitted.

As a first approximation the first integral term will be neglected, and we are left with

$$u(t) = \frac{1}{\omega_0}\int_0^t f(\tau)\sin\omega_0(t-\tau)\,d\tau \quad .. \quad (8)$$

Having evaluated this integral, we can substitute the expression for $u(\tau)$ into

$$\frac{2k}{\omega_0}\int_0^t \psi(\omega_1\tau)\sin\omega_0(t-\tau)\,u(\tau)\,d\tau$$

to obtain a second approximation, and so on to any desired degree of accuracy.

The solution will consist of the first approximate expression for $u(t)$ [Equation (8)] plus a series of terms in ascending powers of k/ω_0^2.

Thus, provided $\omega_0^2 \gg k$ (i.e., $\omega_0^2 \gg \omega_1^2$), the first approximation will give sufficient accuracy.

The integral of equation (8) will now be evaluated.

$$f(t) = \frac{E}{L} e^{\chi(t)} \sin \omega t;$$

$$\text{let } e^{+\chi(t)} = \sum_{n=-\infty}^{\infty} \alpha_n e^{in\omega_1 t} + e^{\chi_0 t}$$
$$= e^{\delta_1(t)} + e^{\chi_0 t}$$

$$\text{and } e^{-\chi(t)} = \sum_{n=-\infty}^{\infty} \alpha'_n e^{in\omega_1 t} + e^{-\chi_0 t}$$

Substituting into (8) and expressing the circular functions as exponentials we get

$$q(t) = e^{-\chi(t)} u(t)$$
$$= \frac{E}{\omega_0 L} e^{-\delta_1 t} \int_0^t e^{-\chi_0 (t-\tau)} \sum_{n=-\infty}^{\infty} \alpha_n e^{in\omega_1 \tau}$$
$$[-\exp i\{\omega_0 t + (\omega - \omega_0)\tau\} - \exp - i\{\omega_0 t + (\omega - \omega_0)\tau\} + \exp - i\{\omega_0 t - (\omega + \omega_0)\tau\} + \exp i\{\omega_0 t - (\omega + \omega_0)\tau\}]\, d\tau \quad .. \quad (9)$$

If the series is uniformly convergent, integration can be carried out term by term. This holds for $R(t)$ periodic in time.

In practice we are usually interested in frequencies near the resonant frequency f_0 of the circuit.

Therefore $\omega \approx \omega_0$ and $|\omega - \omega_0| \ll \omega + \omega_0$. We can, therefore, omit integrating terms of frequency $\omega + \omega_0$ in τ, since they would, on integrating, have $\omega + \omega_0$ in the denominator, thus making them negligible compared with the terms of frequency $\omega - \omega_0$.

Terms with $e^{-\chi_0 t}$ as a factor do not enter into the steady-state solution and are omitted.

Under these conditions we have

$$q(t) = -\frac{E}{4\omega_0 L}\left\{\sum_{n=-\infty}^{\infty} \alpha'_n e^{in\omega_1 t}\right\}$$
$$\left\{e^{i\omega t} \sum_{n=-\infty}^{\infty} \frac{\alpha_n e^{in\omega_1 t}}{\chi_0 + i(n\omega_1 + p)} + e^{-i\omega t} \sum_{n=-\infty}^{\infty} \frac{\alpha_n e^{in\omega_1 t}}{\chi_0 + i(n\omega_1 - p)}\right\}$$
$$= -\frac{E}{2\omega_0 L} \operatorname{Re}\left\{\sum_{n=-\infty}^{\infty} \alpha'_n e^{in\omega_1 t}\right\}$$
$$\left\{\sum_{n=-\infty}^{\infty} \frac{\alpha_n e^{in\omega_1 t}}{\chi_0 + i(n\omega_1 + p)}\right\} e^{i\omega t} \quad .. \quad (10)$$

Where $p = \omega - \omega_0$. Equation (10) can be written as

$$q(t) = -\frac{E}{2\omega_0 L} \operatorname{Re} \sum_{n=-\infty}^{\infty} \beta_n e^{in\omega_1 t} \cdot e^{i\omega t} \quad (11)$$

$$\text{where } \beta_n = \sum_{\mu=-\infty}^{\infty} \frac{\alpha_\mu \alpha'_{n-\mu}}{\chi_0 + i(\mu\omega_1 + p)}$$

We shall now evaluate α_n and α'_n.

$$\chi(t) = \int^t \frac{1}{2L}[R_0 + R_1 \cos \omega_1 \tau]\, d\tau = \frac{R_0}{2L} t + \frac{R_1}{2\omega_1 L} \sin \omega_1 t = \chi_0 t + \delta_1 \sin \omega_1 t$$

$$e^{\delta(t)} = e^{\delta_1 \sin \omega_1 t} = \sum_{n=-\infty}^{\infty} \alpha_n e^{in\omega_1 t}$$
$$= \sum_{n=-\infty}^{\infty} (-i)^n J_n(i\delta_1) e^{in\omega_1 t}$$
$$\therefore \alpha_n = (-i)^n J_n(i\delta_1) \quad .. \quad .. \quad (12)$$

$$e^{-\delta(t)} = e^{-\delta_1 \sin \omega_1 t} = \sum_{n=-\infty}^{\infty} \alpha'_n e^{in\omega_1 t}$$
$$= \sum_{n=-\infty}^{\infty} i^n J_n(i\delta_1) e^{in\omega_1 t}$$
$$\therefore \alpha'_n = i^n J_n(i\delta_1) \quad .. \quad .. \quad (13)$$

and

$$\beta_n = i^n \sum_{\mu=+\infty}^{\infty} \frac{(-1)^\mu J_\mu(i\delta_1) J_{n-\mu}(i\delta_1)}{\chi_0 + i(\mu\omega_1 + p)} \quad (14)$$

4. Multiple-Resonance Phenomenon

The voltage across the capacitor of the resonant circuit is

$$V_c = \frac{q}{C} = -\frac{E}{2\omega_0 LC} \operatorname{Re.} \sum_n \beta_n e^{in\omega_1 t} \cdot e^{i\omega t} \quad (15)$$

This expression shows that the amplitude of the forced oscillations of frequency f is not constant, but varies periodically with time at frequency f_1; i.e., the frequency of the resistance variation.

The amplitude envelope consists of the fundamental of frequency f_1 and an infinite series of harmonics. The magnitude of the nth harmonic of the amplitude envelope is given by:

$$V_n = \frac{E}{2\omega_0 LC} \beta_n$$
$$= \frac{E}{2\omega_0 LC} i^n \sum_{\mu=-\infty}^{\infty} \frac{(-1)^\mu J_\mu(i\delta_1) J_{n-\mu}(i\delta_1)}{\chi_0 + i(\mu\omega_1 + p)} \quad .. \quad .. \quad .. \quad (16)$$

The amplitude envelope is not the same as the resistance waveform. In this case sinusoidal resistance variation produces an envelope which contains an infinite series of harmonics.

The Bessel coefficients decrease rapidly with their order (δ_1 is of the order of 1), hence the amplitudes of the higher harmonics will be very small.

In practical applications one is usually interested in the mean value of the amplitude, since it is this value that determines the amplification due to the circuit.

The mean amplitude A is given by

$$A = \frac{\omega_1}{\pi}\int_0^{\pi/\omega_1} \frac{E}{2\omega_o LC}\sum_{n=-\infty}^{\infty}\beta_n e^{in\omega_1 t}\,dt = \frac{E}{2\omega_o LC}\beta_o$$

Substituting for β_o

$$A = \frac{E}{2\omega_o LC}\sum_{n=-\infty}^{\infty}\frac{J_n^2(i\delta_1)}{\chi_0 + i(n\omega_1 + p)} \quad .. \quad (17)$$

When $\chi_0 = 0$ (i.e., the mean resistance $R_o = 0$) the amplitude will tend to become infinite whenever $n\omega_1 + p = 0$.

If we restrict the meaning of resonance to this tendency of the amplitude to grow very large at a certain frequency of the applied e.m.f. (or at a certain capacitance C of the circuit, for a given driving frequency), then we are faced with the phenomenon of *multiple resonance.*

This occurs at $n\omega_1 + p = 0$, $p = \omega - \omega_o$

$$\therefore \omega = \omega_1 \pm n\omega, \text{ or } f = f_o \pm nf_1 \quad .. \quad (18)$$

For finite R_o, the amplitude of the forced oscillations has its maxima whenever the difference between the natural frequency of the circuit f_o and the frequency of the applied e.m.f. is an integral multiple of the frequency of the resistance variation f_1. This is strictly true only for $\chi_0 = 0$; for finite values of the decrement the resonances are slightly displaced from $p = \pm n\omega$, except for the main resonance, which always occurs at $f = f_o$ ($n = 0$).

5. Amplification

The modulus of A is

$$|A| = \frac{E}{2\omega_o LC}\left|\sum_{n=-\infty}^{\infty}\frac{J_n^2(i\delta_1)}{\chi_0 + i(n\omega_1 + p)}\right|$$

$$= \frac{E}{2\omega_o LC}\sqrt{\left[\left\{\sum_{n=-\infty}^{\infty}\frac{\chi_0 J_n^2(i\delta_1)}{\chi_0^2 + (n\omega_1 + p)^2}\right\}^2 + \left\{\sum_{n=-\infty}^{\infty}\frac{(n\omega_1 + p)J_n^2(i\delta_1)}{\chi_0^2 + (n\omega_1 + p)^2}\right\}^2\right]} \quad (19)$$

The phase angle ϕ is given by

$$\phi = \tan^{-1}\frac{\displaystyle\sum_{n=-\infty}^{\infty}\frac{(n\omega_1 + p)J_n^2(i\delta_1)}{\chi_0^2 + (n\omega_1 + p)^2}}{\displaystyle\chi_o\sum_{n=-\infty}^{\infty}\frac{J_n^2(i\delta_1)}{\chi_0^2 + (n\omega_1 + p)^2}} \quad (20)$$

For $p = 0$ (i.e., the main resonance $f = f_o$) $\phi = 0$. For all the remaining resonances ϕ is finite. Thus, the second condition usually associated with resonance (viz., the voltage in phase with current) holds only for the main resonance.

For constant resistance, $\delta_1 = 0$, we have

$$A_o = \frac{E}{2\omega_o LC}\frac{1}{\sqrt{\chi_0^2 + p^2}}$$

Defining amplification due to the periodic resistance variation as $G = \left|\frac{A}{A_o}\right|$, we get

$$G^2 = \left[\sum_{n=-\infty}^{\infty}\frac{\delta_o\sqrt{\delta_o^2 + (p/\omega_1)^2}}{\delta_o^2 + (n + p/\omega_1)^2}J_n^2(i\delta_1)\right]^2 + \left[\sum_{n=-\infty}^{\infty}\frac{(n + p/\omega_1)\sqrt{\delta_o^2 + (p/\omega_1)^2}}{\delta_o^2 + (n + p/\omega_1)^2}J_n^2(i\delta_1)\right]^2 \quad .. \quad (21)$$

where $\delta_o = \dfrac{\chi_0}{\omega_1} = \dfrac{R_o}{2\omega_1 L}$

For $p = 0$ (i.e., $f = f_o$) the main resonance

$$G = \sum_{n=-\infty}^{\infty}\frac{\delta_o^2 J_n^2(i\delta_1)}{\delta_o^2 + n^2} \quad .. \quad .. \quad (22)$$

This is greater than 1 for any finite value of δ_1. For example:—

If $R_o = 8{\cdot}9\ \Omega$, $L = 0{\cdot}0851$H, $f_1 = 50$ c/s

$\chi_0 = \dfrac{R_o}{2L} = 51{\cdot}9$ (as determined by V_g and valve characteristics)

$$\delta_o = \frac{\chi_0}{\omega_1} = \frac{51{\cdot}9}{100\pi} = 0{\cdot}1654,\ \delta_o^2 \approx 0{\cdot}027,\ \delta_1 = 1{\cdot}2$$

$$\therefore G = 0{\cdot}027\left[\frac{J_o^2(i\delta_1)}{\delta_o^2} + \frac{2J_1^2(i\delta_1)}{1 + \delta_o^2} + \frac{2J_2^2(i\delta_1)}{4 + \delta_o^2} + \frac{2J_3^2(i\delta_1)}{9 + \delta_o^2} + \ldots\right]$$

$$= 1{\cdot}9421 - \frac{0{\cdot}027}{1{\cdot}027} \times 0{\cdot}5108 \times 2 + \frac{0{\cdot}054}{4{\cdot}027} \times 0{\cdot}04105 - \frac{0{\cdot}054}{9{\cdot}027} \times 0{\cdot}00155 + \ldots$$

$$= 1{\cdot}9421 - 0{\cdot}0269 + 0{\cdot}0055 \ldots = 1{\cdot}921$$

G increases rapidly with δ_1; and is much greater for values of δ_1 used in super-regenerative circuits.

6. Physical Interpretation of Multiple Resonance

This can be deduced from equation (3). Compared with $\omega_o^2 = 1/LC$, the terms $R^2(t)/4L^2$ and $R(t)/2L$ can be neglected. This assumption is justified as follows:—

$$\omega_o^2 = \frac{1}{LC} = 4\pi^2 \times (3200)^2 \approx 20100^2 = 4{\cdot}04 \times 10^8$$

$$\frac{\dot{R}}{2L} = \frac{\omega_1 R}{2L} = \omega_1^2\delta_1 = (100\pi)^2\delta_1 = 10^4\pi^2 \times 1{\cdot}2 = 1{\cdot}184 \times 10^5$$

$$\frac{\omega_o^2}{\dot{R}/2L} = \frac{4{\cdot}04}{1{\cdot}184} \times 10^3 = 3420\,;$$

$$\frac{R^2}{4L^2} = \omega^2\delta_1^2 = 10^4\pi^2 \times 1{\cdot}44 = 1{\cdot}42 \times 10^5\,;$$

$$\frac{\omega_o^2}{R^2/4L^2} = \frac{4{\cdot}04}{1.42} \times 10^3 = 2850.$$

Making this approximation, we have

$$\ddot{u} + \omega_o^2 u = \frac{E(t)}{L}\, e^{\chi(t)}$$

This differs from the equation for the circuit with constant resistance in having a periodically-varying decrement $\chi(t)$ instead of a constant one χ_0.

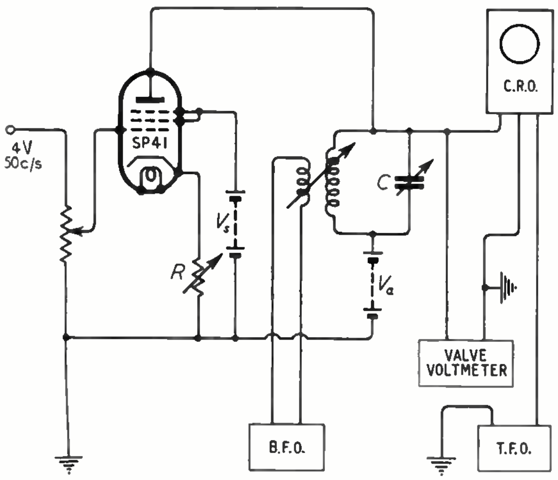

Fig. 2. *Circuit for measuring multiple resonance. (T.F.O.—Tuning fork oscillator).*

But $e^{\chi(t)} = e^{\chi_o t + \delta_1 \sin \omega_1 t}$

$$= e^{\chi_o t} \sum_{n=-\infty}^{\infty} (-i)^n J_n(i\delta_1)\, e^{in\omega_1 t}$$

$$= e^{\chi_o t}\Big[J_o(i\delta_1) + 2\Big\{J_2(i\delta_1)\cos 2\omega_1 t + J_4(i\delta_1)\cos 4\,\omega_1 t + \ldots\Big\} - 2i\Big\{J_1(i\delta_1)\sin \omega_1 t + J_3(i\delta_1)\sin 3\omega_1 t + \ldots\Big\}\Big]$$

The product of this series with $E(t) = E \sin \omega t$, gives another series of e.m.fs of frequencies $f \pm nf_1$ (n — an integer).

Thus, periodic variation of resistance replaces the driving e.m.f. by a series of e.m.fs of frequencies differing by an integral multiple of frequency of the resistance variation f_1.

Since the system is linear we can apply the superposition principle and obtain the resultant response curve by adding the separate responses of the circuit to each e.m.f. of the series acting alone.

The separate responses have a normal resonance peak at $f = f_o \pm nf_1$, and their sum will, therefore, contain all of these peaks.

It can be seen from expression (16) that the nth harmonic of the amplitude envelope has its maximum at $f = f_o \pm (n-1)f_1$. Hence, the shape of the envelope will change as the circuit or the driving frequency are detuned from the resonance at $f = f_o$.

Thus at $f = f_o \pm f_1$ the envelope was predominantly of fundamental and second harmonic frequency; at $f = f_o \pm 2f_1$ second and third harmonic, etc.

Experimental Results

The circuit used in measuring the multiple resonance response is shown in Fig. 2.

The effective resistance of the resonant circuit was varied by applying an alternating voltage to the grid of the dynatron. The resonant circuit was energized by means of a loose coupling from a beat-frequency oscillator, the frequency of which was checked against the tuning-fork oscillator (t.f.o.), using a c.r. oscilloscope for viewing the two waveforms.

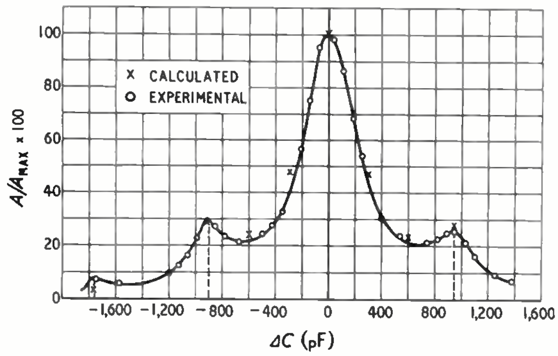

Fig. 3. *Multiple resonance response for* $\delta_1 = 1{\cdot}2$. *Dotted ordinates correspond to* $\Delta f = \pm 50$ *c/s.*

To simplify the experimental arrangement the mains frequency of 50 c/s was used as the frequency of resistance variation f_1. The resonant circuit was tuned to $f = 3200$ c/s, thus giving $(f_1/f)^2 = 0{\cdot}000244 \ll 1$, as required by the theory.

The voltage across C was measured with a peak-reading valve voltmeter. In order to measure the mean voltage a rectifier voltmeter was also used. It was connected to the circuit through the Y-amplifier of the oscilloscope, acting as a buffer stage, to prevent damping of the resonant circuit by a direct connection.

In order to increase the accuracy of the measurements the driving frequency f was kept constant at 3200 c/s, and the response was measured for different values of detuning of the circuit capacitance from its resonant value C_o.

The results are plotted in Figs. 3 and 4.

Fig. 3 shows the mean voltage response A/A_{max} for $\delta_1 = \dfrac{R_1}{2\omega_1 L} = 1{\cdot}2$, $V_g = 1{\cdot}4$ volts peak $= 1{\cdot}0$ V r.m.s., 50 c/s.

In this curve three maxima are clearly distinguishable. They occur at values of ΔC corresponding almost exactly to $\Delta f = \pm 50$ c/s. The calculated points are shown on the curve. They are in good agreement with the experimental results.

Fig. 4 shows the peak voltage response of the circuit, as measured by the valve voltmeter. For this curve the grid alternating voltage was increased, giving a larger resistance swing; i.e., a larger value of δ_1. As a result, six resonance maxima are discernible. If the resistance swing is increased still further, more resonances appear. As many as thirteen were observed.

Besides increasing the number of separate maxima, an increased resistance swing causes the main resonance at $\Delta C = 0$ to become sharper. This is equivalent to the increase of the selectivity or Q of the circuit in the region of the main resonance.

Since the voltage across the resonant circuit is proportional to its Q, the amplification obtained through periodic variation of the resistance can be attributed to this increase in Q.

As the resistance swing is increased the asymmetry in the response becomes more noticeable; the maxima corresponding to $C_o + \Delta C$ are less pronounced than those occurring at $C_o - \Delta C$. The reason for this is twofold: 1. V_c is inversely proportional to C, therefore the response tends to be larger for values of C smaller than C_o; 2. the correct circuit for deriving the differential equation is with the dynatron represented as a varying resistance in parallel with L and C (not as a series one as assumed). This circuit gives the equivalent series resistance introduced by the dynatron as inversely proportional to the capacitance of the circuit. Thus the effective resistance swing is greater for smaller value of the capacitance.

Similar curves were obtained for the frequency detuning (keeping $C = C_o$ constant).

If the frequency of the resistance variation f_1 is increased the resonances move away from the central one. At the same time they become less pronounced due to the fact that the argument of the Bessel coefficients $\delta_1 = \dfrac{R_1}{2\omega_1 L}$ is inversely proportional to f_1.

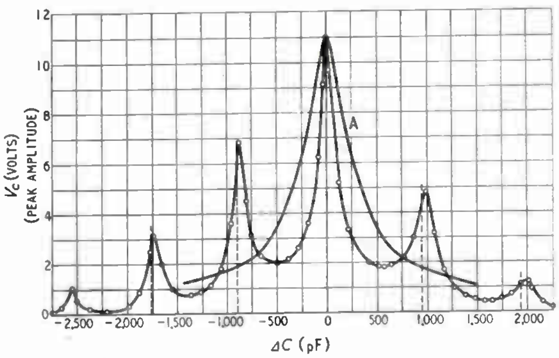

Fig. 4. *Multiple resonance response for $\delta_1 > 2$. Curve A shows the response of a circuit of the same decrement reduced to the same maximum, but in the absence of any time variation of the circuit resistance. Dotted ordinates correspond to detuning $\pm$ 50n c/s.*

The difference between successive maxima and minima of the response is more pronounced due to greater variation in the value of the denominator $\chi_0 + i\,(n\omega_1 + p)$ in the expression (17) for the mean amplitude.

As f_1 is decreased the resonances move closer together and at the same time the difference between maxima and minima becomes less. Thus, finally, as f_1 becomes very small as compared with f_o, the separate resonance peaks merge into a normal smooth response curve.

This is the reason why the multiple resonance cannot be observed if the ratio f_1/f_o is either too large or too small. The other two factors which determine the phenomenon are the natural decrement of the circuit χ_0 and the magnitude of the resistance swing.

Besides being already the basis of superregeneration, the circuits with periodically varying resistance may, in view of the multiple-resonance response, find further application in special filter devices.

(To be concluded in May issue)

MAGNETOSTRICTION TRANSDUCER MEASUREMENTS

By H. J. Round

THE name transducer is now given to certain reversible devices which transform electrical energy into mechanical energy or vice versa. Sometimes the term is localized to piezo-electric and magnetostriction instruments, but, used in a wider sense, the Bell telephone, moving-coil instruments, electrostatic telephones, etc., come under the same heading.

Magnetostriction action, which will be particularly studied here, is in fact very similar to the Bell receiver action. In the telephone receiver an alternating current passes through the winding, sets the diaphragm vibrating and, due to the change of air gap, a change of reluctance takes place. Owing to the presence of the permanent magnet this change of reluctance produces a change of flux which in turn produces across the input winding a back e.m.f.

All magnetostriction devices are polarized with a steady field in some way or other. An alternating exciting field is superimposed on the steady field. For those not familiar with magnetostriction instruments, Fig. 1 gives a sketch of a simple magnetostriction-rod oscillator which can easily be set up and with which all the measurements afterwards described can be made.

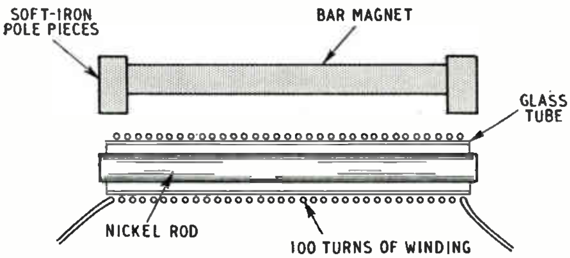

Fig. 1. *Simple magnetostriction oscillator.*

A core of laminated nickel stampings 1 cm^2 cross-section and 12 cm long is a useful size and it will have a fundamental resonance of about 20,000 c/s. The stampings (or cuttings from a nickel sheet) must be well annealed at about 800 C for a few minutes. Then they should be tied together not too tightly in a bundle, impregnated with a varnish such as shellac varnish and baked until a solid rod is formed. Any nickel thickness from 0·004 in. to 0·012 in. will do, but the thinner material will give less electrical losses.

The core should now be fitted loosely into a glass tube which can be the same length as the rod. About a hundred turns of wire should now be wound round the glass tube from end to end.

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To polarize the rod, a straight bar magnet 10 cm long fitted with two iron pole-pieces can be brought near the rod. The distance from the rod for best action will depend on magnet strength, and a rough criterion can be obtained as follows:—

When the apparatus is set up for impedance measurements, measure the impedance with the magnet far away from the rod at a frequency well below the resonant frequency. Now bring the magnet up to the rod until this impedance falls to about two-thirds of its initial value. Magnetostrictive action will then be very strong. Actually there is quite a large range of field strength which gives strong action. With normal nickel the best flux density is about 5,000 gauss.

As in the Bell receiver, alternating current passing through the exciting winding sets up mechanical changes of dimensions, this change of dimension varies the reluctance of the material and owing to the presence of the polarizing m.m.f. an additional change of flux is produced which then appears as a back e.m.f.

Both telephone and magnetostriction instruments are reversible, for vibration applied to the diaphragm or to the magnetostrictive material will give an e.m.f. in the respective windings.

Kennelly's Research

Kennelly was the first to study the telephone receiver, chiefly near its resonant points, and he evolved methods of measurement and analysis which are now classical.

His method consisted in measuring the impedance of the telephone at a large number of frequencies, first, with the diaphragm clamped so that it could not move and then with the diaphragm allowed to move freely. In these measurements the reactance and effective resistance were determined at each impedance measurement so that the vectorial impedance was known. He then vectorially subtracted the two sets of measurements and showed that the resulting figures near any single resonance point would give a circle diagram when plotted with the reactance as ordinates and the resistance as abscissa.

He called this a motional-impedance circle and from its dimensions he reasoned out certain constants of the telephone. The circle was identical with that which would be given by a parallel

LCR circuit but was actually due to the reaction of the mass, elasticity and mechanical resistance of the diaphragm.

One further experiment determining the effective mass of the diaphragm by the method of adding a small known mass to the diaphragm and determining the change of period enabled him to get the energy relation between the electrical constants of the circle and the mechanical system which really produced it. For instance he could say that the mass he measured should be represented by the inductance of his circle diagram and then from the energy relation $\frac{1}{2}mv^2 = \frac{1}{2}LI^2$ he would get a relation between the diaphragm velocity and the exciting current. Thus the diaphragm velocity or displacement was at once known for any value of current.

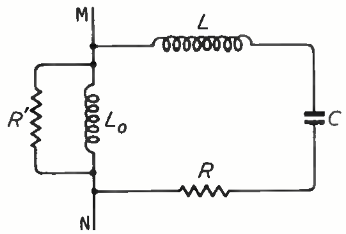

Fig. 2. *Equivalent circuit of magnetostriction oscillator.*

Kennelly's circle diagram is still used for magnetostriction measurements, the only difference being that a clamped impedance curve is not available since it is impossible to clamp the device except in cases where the amount of material is very small. An approximate result can be obtained by interpolation from the unclamped curve, however, if sufficient readings are taken well away from the resonant point. The actual effective mass of the magnetostriction device can be obtained either by calculation in the case of straight rods, or generally by using Kennelly's added mass method.

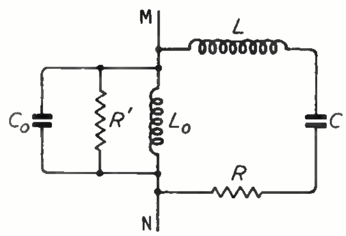

Fig. 3. *Circuit of Fig. 2 with additional shunt capacitance C_o.*

In this way fairly accurate estimates of the velocities of the device can be obtained and power outputs into air or water can be calculated.

As is well known, magnetostriction is greatly used for the production of high-frequency waves in water for signalling purposes and the behaviour of the devices has been studied mainly by Kennelly's methods.

Actual mechanical movements are quite impossible to measure with any degree of accuracy owing to their extreme smallness.

Equivalent Circuits

A step forward in the examination of transducers was made by the introduction of equivalent circuits. The idea is that all these reversible devices are similar in action to a leaky transformer and that an electrical circuit can be constructed which at its input terminals will behave like the transducer. The equivalent can then be used for calculation or in the form of an electrical dummy to make measurements which would be difficult to do in the actual transducer. One of the most convenient of these equivalent circuits is illustrated in Fig. 2 where L_o represents the inductance of the exciting winding and LCR represents the motional action. A resistance R' in shunt with L_o is found to represent the magnetic losses fairly well and R represents the mechanical losses.

It is found that the diagram simulates very fairly the action of the magnetostriction device near its resonant point.

If such a circuit is set up and the constants chosen correctly, the impedance behaviour at the points MN over a small frequency band near the resonant point will almost exactly duplicate the impedance behaviour of a magnetostriction instrument such as that of Fig. 1. From Kennelly measurements, the exact constants in the equivalent circuit can be obtained.

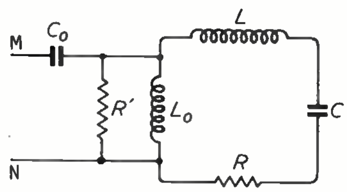

Fig. 4. *Circuit of Fig. 3 with additional capacitance C_o in series with the input.*

Simplified Method of Obtaining the Constants

The Kennelly method is rather laborious and an attempt to simplify it has been made which has been successful, except in a limited number of cases which will be considered later.

It will be noted that if a capacitor C_o (Fig. 3) is added to Fig. 2 in shunt with L_o and R', a type of coupled circuit is produced and if L_oC_o is made resonant at the same frequency as LC, an arrangement with unity power factor is obtained across the terminals MN.

In consequence, this is a useful combination for the insertion of power from a generator. C_o may alternatively be added in series at the points MN as in Fig. 4 in which case C_oL_o will be made to tune with $(L_o + L)C$ and R to give unity power factor.

This latter arrangement gives the usually accepted tuned coupled-circuit arrangement and the coupling coefficient is given by $\sqrt{\frac{L_0}{L + L_0}}$.

By the use of either of these tuning methods a more rapid method of determining the equivalent circuit can be obtained.

For certain reasons the shunt tuning arrangement of Fig. 3 is more convenient and in the following discussion the liberty has been taken of calling $k = \sqrt{\frac{L_0}{L}}$. The accepted value of k can easily be calculated if required.

Overall Impedance

If Y is the admittance of L_0C_0R' at the terminals MN of Fig. 3 and Z the impedance of LCR, the total impedance at the terminals will be $\frac{Z}{YZ + 1}$.

If Q_0 and Q refer to L_0C_0R' and LCR respectively, f_0 the frequency to which both are tuned and Δ is a small frequency difference from f_0 then it can easily be shown that

$$Y = \frac{1}{\omega_0 L_0}\left[\frac{1}{Q_0} + j\frac{2\Delta}{f_0}\right]$$

$$Z = \omega_0 L\left[\frac{1}{Q} + j\frac{2\Delta}{f_0}\right]$$

ω_0 has the usual meaning of $2\pi f_0$.

The impedance expression $\frac{Z}{YZ + 1}$ then becomes when rationalized

$$\omega_0 L_0 \frac{\left[\left\{\frac{1}{Q}\left[\frac{1}{Q_0Q} - \left(\frac{2\Delta}{f_0}\right)^2 + k^2\right] + \left(\frac{2\Delta}{f_0}\right)^2\left(\frac{1}{Q_0}+\frac{1}{Q}\right)\right\} + j\frac{2\Delta}{f_0}\left[k^2 - \left(\frac{2\Delta}{f_0}\right)^2 - \frac{1}{Q^2}\right]\right]}{\left[\frac{1}{Q_0Q} - \left(\frac{2\Delta}{f_0}\right)^2 + k^2\right]^2 + \left[\frac{2\Delta}{f_0}\right]^2\left[\frac{1}{Q_0}+\frac{1}{Q}\right]^2}$$

This expression is purely resistive when the j term is zero. This happens when Δ is zero and also when

$$\left(\frac{2\Delta}{f_0}\right)^2 = k^2 - \frac{1}{Q^2} \quad \cdots \quad \cdots \quad \cdots \quad (1)$$

This latter, of course, refers to two points on the impedance curve equi-distant above and below f_0. Calling the impedance [resistance] when Δ is zero R_B, then from the full equation

$$R_B = \omega_0 L_0 \left[\frac{1/Q}{k^2 + 1/QQ_0}\right] \quad \cdots \quad \cdots \quad (2)$$

and calling the impedance at the two equidistant points R_{AC} we get

$$R_{AC} = \omega_0 L_0 \frac{1}{\frac{1}{Q_0} + \frac{1}{Q}} \quad \cdots \quad \cdots \quad \cdots \quad (3)$$

If the values of $\omega_0 L_0$, f_0, Δ, R_B and R_{AC} are measured, there will be sufficient data to determine the equivalent circuit.

Experimental Layout

The experimental arrangement for determining the required constants is shown in Fig. 5. A beat-frequency oscillator is connected through the usual high resistance to give constant current to the transducer T or to the resistance box RB via the switch S.*

A c.r. oscilloscope is connected as shown, the b.f.o. voltage being put on the X axis, and the voltage across the transducer on the Y-amplifier axis. If the voltages applied to X and Y are in phase the oscilloscope figure will be a sloping straight line; and if they are not in phase an ellipse will be produced.

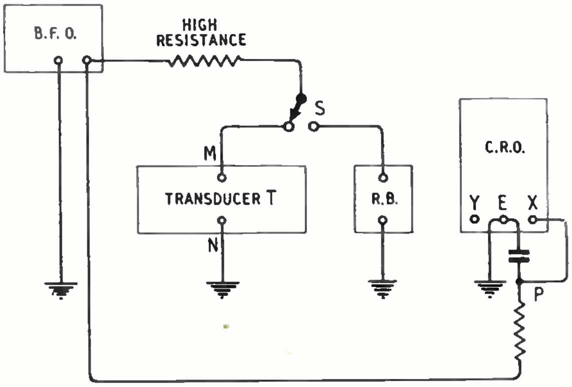

Fig. 5. *Arrangement of apparatus for transducer measurements*

A small RC network P is usually necessary initially to set equality of phase when the switch is thrown to RB. This network P can be fixed for quite a large range of frequency and gain control on some oscilloscopes.

With the transducer and its tuning capacitor C_0 connected, a run through the frequency range on the b.f.o. will give the well-known M curve shown in Fig. 6, the impedances at any point being determined by comparison with the resistance box. The apparatus will also show the phase changes taking place.

At positions A, B and C it will be noted that the transducer is resistive. The tuning capacitor

*H. J. Round, "Impedance Measurements," *Wireless Engineer*, May 1950 p. 154.

should be adjusted until the three resistive points are equally spaced in frequency, these frequencies noted and the resistances at A, B and C measured.

The resistances R_{AC} at A and C should of course be equal but, if not very different, an average value can be assumed. If they are very different a check should be made on the frequency spacing between A, B and C.

The determination of the constants of the equivalent circuit from these measurements is now comparatively easy and in certain cases very easy.

Calculation of Constants

The capacitance C_o, which can be measured by comparing it with a resistance at the frequency f_o, will give the value of $\omega_o L_o$ at once.

Then from equations (1), (2) and (3) we have

$$\frac{1}{Q} = \frac{1}{\omega_o L_o}\left(\frac{2\Delta}{f_o}\right)^2 \left[\frac{R_B R_{AC}}{R_{AC} - R_B}\right] \quad \cdots \quad (4)$$

$$\frac{1}{Q_o} = \frac{\omega_o L_o}{R_{AC}} - \frac{1}{Q} \quad \cdots \quad \cdots \quad \cdots \quad (5)$$

and $$k^2 = \left(\frac{2\Delta}{f_o}\right)^2 + \frac{1}{Q^2} \quad \cdots \quad \cdots \quad \cdots \quad (6)$$

We also have

$$\frac{1}{Q} = \frac{R}{\omega_o L} = \frac{Rk^2}{\omega_o L_o} \text{ or } R = \frac{\omega_o L_o}{Qk^2}$$

$$k^2 = \frac{\omega_o L_o}{\omega_o L} \text{ or } \omega_o L = \frac{\omega_o L_o}{k^2}$$

$$\frac{1}{Q_o} = \frac{\omega_o L_o}{R'} \text{ or } R' = \omega_o L_o Q_o$$

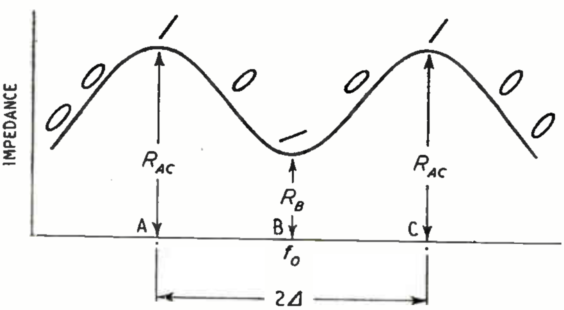

Fig. 6. *Form of resonance curve obtained, together with an indication of the type of oscilloscope trace obtained.*

Simplification of Calculation

In certain cases, such as oscillators in air, Q is usually very high and neglecting $1/Q$ in (5) and (6) gives us

$$\frac{1}{Q_o} = \frac{\omega_o L_o}{R_{AC}} = \frac{\omega_o L_o}{R'}$$

so that, $R' = R_{AC}$

Also from (6)

$$k^2 = \left(\frac{2\Delta}{f_o}\right)^2 \text{ or } k = \frac{2\Delta}{f_o} \quad \cdots \quad \cdots \quad (7)$$

This equation (7) is very important when studying variations in the experimental apparatus such as change of permanent field, because it enables the value of k to be watched all the time without elaborate calculations.

Finally R can be found in this simplified case from (4), or more simply by noting that

$$\frac{1}{R} = \frac{1}{R_B} - \frac{1}{R_{AC}} \approx \frac{1}{R_B}$$

The curve under these conditions enables the required constants to be determined almost directly, as illustrated in Fig. 7.

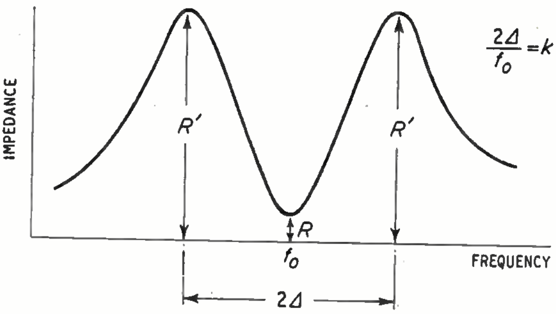

Fig. 7. *Form of resonance curve obtained when the Q is very high, indicating simplified calculations.*

Meaning of Constants

All the previous work is merely for the purpose of determining the equivalent circuit of a given transducer. Given these constants what is their meaning?

L_o and R' represent the magnetic part of the arrangement. If the tuning capacitor is put in shunt a certain Q_o value results. This Q_o can, of course, be decreased by additional shunt resistance, but can only be increased by the addition of low-loss inductance which, however, reduces k, unless the magnetic circuit is a bad one and it is possible to improve it. A case of this can be noted with Fig. 1.

There is, of course, a considerable air gap in a short straight-rod system. If on the opposite side of the bar to the polarizing magnet is placed a laminated U piece, or even a straight laminated bar, k will be increased. Of course, a new value of C_o must be found and possibly the magnet brought a little nearer. This increase of k will be at once noticed by an increase of Δ.

L can represent the mass of the oscillator or the compliance, whichever we like to choose, providing we determine the correct energy relation between the equivalent circuit and the real thing. R is a factor which is partly frictional resistance and partly radiation. They are not easy to separate.

Suppose we have a magnetostriction device designed to radiate in water.

Then first of all we determine its equivalent circuit in air when Q will be quite high if it is efficient. Q_o, k and $\omega_o L_o$ will simultaneously be determined.

If we now immerse the transducer in water, the only factor that will seriously alter is R. A new measurement will now determine the new R and by subtracting the air R, a rough value for the radiation resistance will be obtained. This is not entirely accurate because the immersion in water usually adds some more frictional resistance and more measurements of a different type are necessary to get the real radiation figure.

Difficult Cases

Examining the equation

$$\left(\frac{2\Delta}{f}\right)^2 = k^2 - \frac{1}{Q^2}$$

it can be seen that the method is inapplicable if k^2 is equal or less than $1/Q^2$, for in those cases there is only the central resistance point. This does not often occur in air oscillators. For instance if $Q = 100$ the coupling would have to be less than 1% for the 3 points to come together, that is for k^2 to be less than $1/Q^2$.

However, when oscillators are used in water it may occur and in that case, a measurement in air should be made, giving $\omega_o L_o$, k^2 and Q_o, and these can be assumed as the same when in water. Then from equation (2) the value of Q can be obtained.

It is sometimes preferable in these extreme cases to go back to the Kennelly method, although even then the analysis will be rather inaccurate as the motional circle will be small.

It should be noted that the leads between the capacitor C_o and the instrument should be as short as possible, otherwise the d.c. resistance of these leads and their inductance may result in the diagram being really as in Fig. 8 and the equations on which the method is founded no longer apply.

Kennelly's method does in a way enable d.c. resistance and any leakage inductance to be eliminated, but with magnetostriction oscillators the winding is usually a thick one of very low d.c. resistance and the linkage with the nickel is very close.

Results of analysis of oscillators by both methods give very similar results and the new method is much more rapid.

It is of particular value in watching the variations of k when changes, such as, of polarization-field values, are made. The factor k in magnetostriction instruments is directly related to the magnetostriction constant only when closed toroidal formations are used.

With the short straight rods or, in fact, any arrangement where an air gap must be reckoned with, k will be less than this value. No greater value is possible unless another material is used. k for good nickel in a toroidal form, used at the best polarization, is about 30%. Straight rods of dimensions such as in Fig. 1 are about 16% due to the air gap.

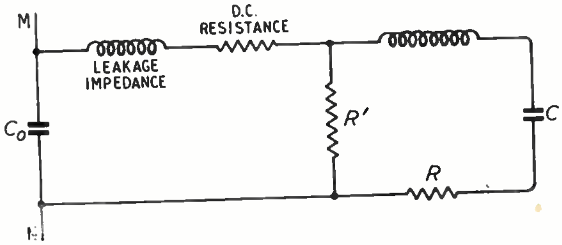

Fig. 8. *Form of circuit effectively obtained when long leads are used to the capacitor C_o.*

Fundamental Formula

It may be useful to put down here the magnetostriction definitions which are commonly used. The magnetostriction constant λ is defined by the formula.

$$dF = \lambda dB$$

where dF is the magnetostriction force produced by a flux change dB.

If, on a unit length of magnetostrictive material away from the mechanical resonance point, a change of flux dB is produced by electrical means, there will be a total storage of potential energy E. Of this energy E_1 will be in the form of mechanical stored energy.

The coupling coefficient is then related to these two energies by the formula

$$K^2 = E_1/E$$

and this results in an overall formula

$$K = \lambda\sqrt{\frac{4\pi\mu}{E}}$$

where μ is the material permeability and E the Youngs modulus.

It must be noted that this K is not quite the same as the k used in the previous work.

Referring to the equivalent circuit diagrams the k used there is equal to $\sqrt{L_o/L}$ whereas this K, the true coupling coefficient, is equal to $\sqrt{L_o/(L + L_o)}$ The relation between the two is obviously

$$K^2 = \frac{k^2}{1 + k^2}$$

Well-annealed nickel operated at a static H value of 15 has a permeability of about 30 and the value of E is 20×10^{11} so that, with the above figure of $k = 30\%$, the consequent value of K is 28·8% and the magnetostriction constant

$$\lambda = K\sqrt{\frac{E}{4\pi\mu}} = 21{,}000.$$

TCHEBYSHEV FILTERS AND AMPLIFIER NETWORKS

By Vitold Belevitch

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SUMMARY.—The insertion-loss method of filter synthesis, described by Darlington,[1] is applied to the design of simple low-pass filters composed of alternate series coils and shunt capacitors; i.e., to structures similar to conventional multi-section constant-k filters. Both the case of two resistive terminations and of one open-circuit termination are examined. When filters of this type are used as amplifier input or output networks, the prescribed value of the terminal shunt capacitance imposes a known physical limitation on the gain-bandwidth product of the stage. This has been studied by Bode[2] but additional precision results from the present work. More general filters with one open-circuit termination are also mentioned and a new method of design is outlined. Only low-pass filters are considered but the results can be extended to other types of filters by frequency transformations.

SYMBOLS

f_c = cut-off frequency
ω_c = $2\pi f_c$
Ω = f/f_c normalized frequency
α = $\cos^{-1}\Omega$, transformed frequency in the pass-band
p = $j\Omega$
R = value of terminal resistance or resistances
L_o = R/ω_c reference inductance
C_o = $1/R\omega_c$ reference capacitance
$f_1 f_2 \ldots$ = All filter components are normalized; i.e., inductances are divided by L_o and capacitances by C_o. Normalized components are denoted by $f_1\, f_2\, f_3 \ldots$ irrespective of their nature
A = insertion loss in nepers (1 neper = 8·68 db)
A_m = maximum insertion loss in the pass-band
H = $(e^{2A_m} - 1)^{\frac{1}{2}}$
β = $\log_e \coth (A_m/2) = 2 \sinh^{-1} H^{-1}$
A_e = return loss in nepers
n = number of filter sections (integer or half-integer)
k = transformer turns ratio

1. Insertion Loss Characteristics

FIG. 1 shows three typical forms of characteristic which can be obtained for a low-pass filter according to the method used for the design.

In the *conventional method* all normalized filter components have the value 2, except the two-terminal components which are halved, and the resulting insertion loss, for a filter composed of n complete sections and working between two resistances R, is

$$A = \tfrac{1}{2} \log_e (1 + \sin^2 2n\alpha \cos^4\alpha/4 \sin^2\alpha) \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

in the pass-band. The trigonometric expression in (1) is a polynomial in $\Omega = \cos\alpha$ and, in that form, holds in the attenuation range. A similar compact expression is obtained for the attenuation range by replacing circular by hyperbolic functions. The characteristic (1) has $n - 1$ maxima of amplitude increasing to cut-off [see Fig. 1 (a)] and, at the theoretical cut-off frequency, the insertion loss has the large value $\frac{1}{2}\log_e (1 + n^2)$. These are known drawbacks of the conventional design.

By altering the component values a more convenient behaviour of the insertion loss can be secured. When a maximum attenuation A_m is imposed in the pass-band, a steadily increasing characteristic according to Fig. 1 (b) is obtained from

$$A = \tfrac{1}{2} \log_e (1 + H^2\Omega^{4n+2}) \quad \ldots \quad \ldots \quad (2)$$

and corresponds to the so-called *maximally-flat filters* studied by various authors.

It is well known, however, that the largest loss in the attenuation range, compatible with a given A_m in the pass-band, is obtained from the *Tchebyshev characteristic* [Fig. 1 (c)].

$$A = \tfrac{1}{2} \log_e \left[1 + H^2 \cos^2(2n + 1)\alpha\right] \quad \ldots \quad (3)$$

As $\cos (2n + 1)\alpha$ is a polynomial in $\Omega = \cos\alpha$, with $2^{2n}\Omega^{2n+1}$ as highest term, the asymptotic behaviour at high frequencies is

$$A = \log_e H + 2n \log_e 2 + (2n + 1) \log_e \Omega \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

and differs by $2n \log_e 2$ from the corresponding expression for the maximally-flat filter. Thus, although both characteristics give the same maximum distortion in the pass-band, the Tchebyshev behaviour gives an additional 12-db loss per section at high frequencies. For small

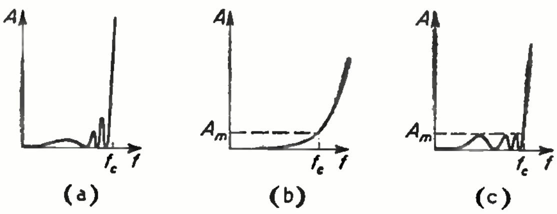


Fig. 1. *Insertion loss characteristics of (a) conventional filter, (b) maximally-flat filter, and (c) Tchebyshev filter.*

MS accepted by the Editor, March 1951

values of A_m, $\log_e H$ is approximately $\frac{1}{2}\log_e 2A_m$ and, thus, halving A_m corresponds to a 3-db decrease in high-frequency loss.

2. Design Formulae for Tchebyshev Filters

Filters having the insertion loss (3) and working between two resistances equal to R are designed by Darlington's method. As the final formulae seem to have never appeared in the literature, they will first be stated without proof; an outline of the derivation is given in next section.

For integral values of n the filter is geometrically *symmetrical* and contains $2n + 1$ elements, but only $n + 1$ elements need be calculated. By considering normalized values, the same formulae apply to both dual configurations of Fig. 2 (a) and (b).

Introducing the auxiliary symbols

$$\gamma = \sinh \beta/2\,(2n + 1) \qquad (5)$$

$$\left.\begin{aligned} a_i &= \sin\,(2i + 1)\pi/2\,(2n + 1) \\ b_i &= \cos\,(2i + 1)\,\pi/2\,(2n + 1) \end{aligned}\right\} \qquad (6)$$

where $i = 0, 1, 2 \ldots$ The design formulae for the $n + 1$ distinct elements $f_1, f_2 \ldots$ are:

$$\left.\begin{aligned} f_1 &= 2a_0/\gamma \\ f_2 &= \frac{2a_1\gamma}{\gamma^2 + b^2_{n-1}} \\ f_3 &= \frac{2a_2\,(\gamma^2 + b^2_{n-1})}{\gamma\,(\gamma^2 + b^2_{n-2})} \\ f_4 &= \frac{2a_3\,\gamma(\gamma^2 + b^2_{n-2})}{(\gamma^2 + b^2_{n-1})\,(\gamma + b^2_{n-3})} \end{aligned}\right\} \qquad (7)$$

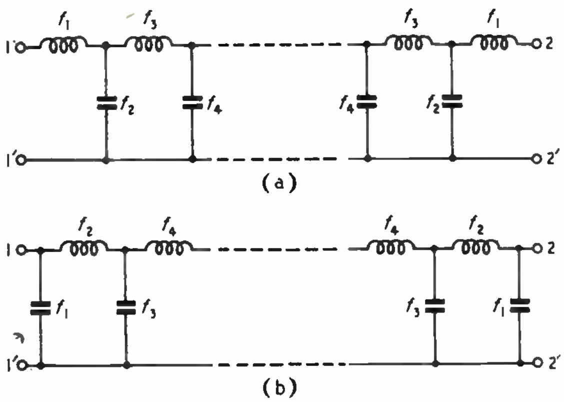


Fig. 2. *Dual forms of symmetrical low-pass filter.*

Half integral values of n lead to *antisymmetrical filters*. Expression (3) then shows that the insertion loss takes the value A_m at zero-frequency and, for a low-pass filter, this can only be obtained by mismatching the terminations. For design purposes it is simpler to keep equal terminal resistances and to incorporate an ideal transformer in the filter. If the transformer is inserted in the centre of the filter the structural antisymmetry is preserved and it is again sufficient to calculate half of the elements (Fig. 3). The transformer ratio is deduced from the mismatch loss A_m and is

$$k = \tanh \beta/4 \qquad (8)$$

This ratio is always smaller than 1 (with the notation of Fig. 3), the transformer presenting its low-impedance side to the mid-shunt termination side of the filter. The mismatch is thus in the same direction as in the conventional design. The $n + 1/2$ distinct elements $f_1, f_2 \ldots$ are still given by formulae (7).

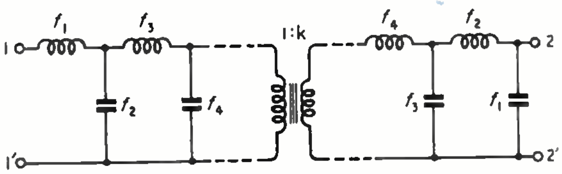


Fig. 3. *Antisymmetrical low-pass filter.*

3. Derivation of the Formulae

We briefly indicate the derivation of formulae (7) for symmetrical filters, according to Darlington's method. Consider the expression

$$\begin{aligned} e^{2A} &= 1 + H^2\cos^2\,(2n + 1)\alpha \\ &= |\,1 + jH\cos\,(2n + 1)\alpha\,|^2 \end{aligned}$$

The roots of

$$1 + jH\cos\,(2n + 1)\alpha = 0$$

are

$$\alpha_i = [(1 + 2i)\pi + j\,(-1)^i\beta]/2\,(2n + 1)\,,$$

where $i = 0, 1 \ldots 2n$, and the corresponding complex frequency roots are

$$p_i = j\cos\alpha_i = (-1)^i\gamma a_i + j\,(1 + \gamma^2)^{\frac{1}{2}} b_i$$

By separating roots $p_0, p_2 \ldots$ with positive real parts from roots $p_1, p_3 \ldots$ with negative real parts, two Hurwitz polynomials

$$\begin{aligned} A_1 + pB_1 &= (p - p_1)\,(p - p_3) \ldots \\ A_2 + pB_2 &= (p + p_0)\,(p + p_2) \ldots \end{aligned}$$

are obtained, where A_1, A_2, B_1, B_2 are polynomials in p^2. The normalized lattice impedances of the symmetrical filter are then

$$Z_1 = pB_1/A_1\,;\; Z_2 = A_2/pB_2$$

By Bartlett's bisection theorem, a continued fraction expansion of either Z_1 or Z_2 yields all the elements $f_1, f_2 \ldots$ except the central element f_{n+1} which can only be obtained from one of these expansions. By identifying successive coefficients of the expansions of Z_1 and Z_2 a number of relations are obtained, which simplify the calculation of further coefficients. As an example, the first coefficient of the expansion of Z_1 is the ratio of the second to the first highest coefficient in $A_1 + pB_1$ and this is the negative sum of the roots $-(p_1 + p_3 + \ldots)$; similarly the first coefficient of Z_2 is $p_0 + p_2 \ldots$. Both coefficients have the common value

$$\tfrac{1}{2}\sum_{i=0}^{2n} (-1)^i p_i = \tfrac{1}{2}\gamma \sum_{i=0}^{n} a_i = \gamma/2 \sin \pi/2(2n+1)$$

and this yields the reciprocal of the first filter component f_1 as given by (7). The derivation of the design formulae for further elements and for the antisymmetrical case involves tedious trigonometric calculations and will be omitted.

4. Application to Amplifier Circuits

For application to amplifier input or output circuits the value of the terminal shunt capacitance C is of particular interest, as it is usually identified with the valve and transformer parasitic capacitance. Its normalized value

$$x = CR\omega_c \qquad \ldots \quad \ldots \quad \ldots \quad (9)$$

which is the parameter considered by Bode, is f_1 in our notation, and the first formula (7) gives

$$x = 2[\sin \pi/2(2n+1)]/[\sinh \beta/2(2n+1)] \quad (10)$$

which tends to $2\pi/\beta$ for large values of n. In Fig. 4(a) x is plotted versus A_m for various values of n and it appears that the convergence to the limiting value $2\pi/\beta$ $(n = \infty)$ is quite rapid.

According to Bode, the terminal shunt capacitance C of a reactive network

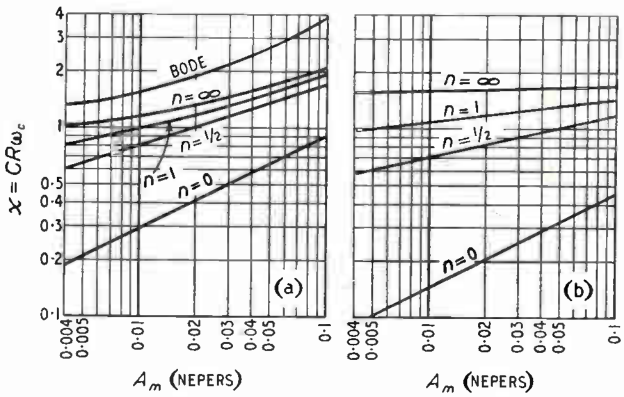


Fig. 4. *Normalized value $x = CR\omega_c$ of the last element of a Tchebyshev filter as a function of the distortion A_m in the pass-band. Filter working between resistances (a) and open-circuit filter (b).*

working between resistances R is related to the frequency integral of its return loss A_e by

$$\int_0^\infty A_e d\omega = \pi/CR \qquad \ldots \quad \ldots \quad \ldots \quad (11)$$

Using the known relation

$$e^{-2A} + e^{-2A_e} = 1 \qquad \ldots \quad \ldots \quad \ldots \quad (12)$$

and introducing the normalized frequency, (11) becomes

$$\int_0^\infty \log_e (1 - e^{-2A})^{-1} d\Omega = 2\pi/x \qquad \ldots \quad (13)$$

Bode considers the ideal case where the insertion loss is infinite in the attenuation range and has the constant value A_m in the pass-band and so obtains for x the limiting value

$$x = 2\pi/\log_e (1 - e^{-2A_m})^{-1} \qquad \ldots \quad \ldots \quad (14)$$

which is also represented in Fig. 4(a).

Bode's relation yields another method of establishing formula (10), by calculating the integral (13), where A has the value (3). The integral is

$$\int_0^\infty \log_e[1 + 1/H^2\phi^2(\Omega)]d\Omega =$$
$$\pi \sinh [\sinh^{-1} 1/2H(2n+1)]/\sin \pi/2(2n+1) \qquad \ldots \quad \ldots \quad (15)$$

where $\phi(\Omega)$ denotes the polynomial

$$\phi(\Omega) = \cos [(2n+1) \cos^{-1}\Omega] \qquad \ldots \quad (16)$$

It is interesting to note that our result for $n = \infty$ does not converge to Bode's limiting value. In our case, the insertion loss is actually infinite in the attenuation range, but its oscillatory behaviour in the pass-band accounts for the difference; this is essentially due to the Tchebyshev behaviour.*

5. Open-Circuit Filters

Filters working between a generator of given internal resistance R at the input and an open-circuit at the output are also studied in Darlington's paper. Their insertion loss and phase are defined by

$$A + jB = \log_e (V_2/E)$$

where V_2 is the open-circuit output voltage and E the generator e.m.f. Denoting by Z_{11}, Z_{12}, Z_{22} the elements of the normalized impedance matrix of the network, the following relation holds

$$e^{A+jB} = (Z_{11} + 1)/Z_{12} \qquad \ldots \quad \ldots \quad (17)$$

and the design is based on the research of the roots of (17). It is interesting to note that the design is reduced to *rational operations* if the elements of the filter having the same insertion loss between two finite resistances are known.

* After the completion of the manuscript, the author's attention has been drawn to a paper by R. M. Fano[3] where a sequence of filters is given which converge to Bode's limiting value in the terminated case.

Since this is the case for all conventional filters, it is possible to design, by rational operations only, an open-circuit filter having the same insertion loss as any prescribed terminated conventional filter. In particular, open-circuit filters having the loss characteristic (1) are easily designed, as will be shown in Section 6.

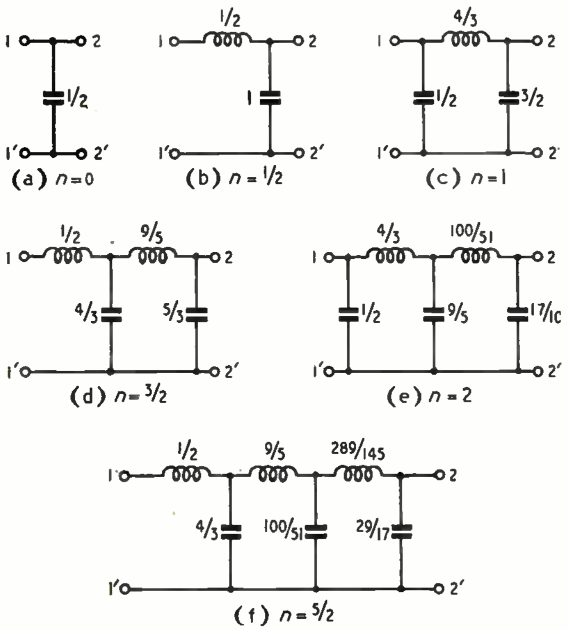


Fig. 5. *Open-circuit filters with conventional behaviour (normalized elements, 22′ is the open-circuited terminal pair.*

Consider a terminated filter of normalized-impedance matrix elements Z'_{11}, Z'_{12}, Z'_{22}, having the same insertion characteristics $A + jB$ as the open-circuit filter considered in (17). The relation corresponding to (17) for a terminated filter is

$$e^{A+jB} = [(1 + Z'_{11})(1 + Z'_{22}) - Z'^2_{12}]/2Z'_{12} \quad (18)$$

Identifying (17) and (18) and separating real and imaginary parts gives two relations which enable one to calculate the elements Z_{11} and Z_{12} of the open-circuit filter to be designed, in terms of the known elements of the terminated filter. In particular the value of Z_{11} is

$$Z_{11} = (Z'_{11} + Z'_{22})/(1 + Z'_{11}Z'_{22} - Z'^2_{12}) \quad (19)$$

and the knowledge of Z_{11} is usually sufficient to calculate all the filter elements by means of a continued fraction expansion. If the terminated filter is symmetrical, relation (19) is further simplified by introducing the normalized lattice impedances Z'_1 and Z'_2:

$$Z_{11} = (Z'_1 + Z'_2)/(1 + Z'_1 Z'_2) \quad .. \quad (20)$$

6. Open-Circuit Filters with Conventional Behaviour

Relation (20) will be applied to the design of open-circuit filters having a loss characteristic given by (1). From elementary filter theory, the equivalent lattice impedances of a symmetrical n-section constant-k low-pass filter terminated at mid-series are

$$Z'_1 = -j \tan n\alpha/\sin \alpha; \; Z'_2 = j \cot n\alpha/\sin \alpha$$

The open-circuit input impedance of the open-circuit filter is, by (20),

$$Z_{11} = -j(1 + \sin^2\alpha) \tan 2n\alpha/2 \sin \alpha \quad .. \quad (21)$$

This is always a rational fraction in Ω; for instance, for $n = 1$:

$$Z_{11} = j\Omega(\Omega^2 - 2)/(2\Omega^2 - 1) \quad .. \quad .. \quad (22)$$

By expanding (22) in a continued fraction, the filter of Fig. 5(a) is obtained. It is interesting to compare this with the terminated filter with normalized elements 1, 2, 1; this shows that the total shunt capacitance is unchanged.

The method just described holds for any value of n; it is sufficient to expand (21) in a continued fraction. A similar design based on (19) enables one to synthesize open-circuit filters equivalent to conventional antisymmetrical filters. The results are summarized in Fig. 5, and it appears that this family of filters is characterized by the sequence

$$f_1 = 1/2, \; f_2 = 4/3, \; f_3 = 9/5, \; f_4 = 100/51 \, .. \quad (23)$$

which gives all the elements, except the last one, of any filter of the family. Considering in addition the sequence of last elements

$$g_1 = 1, \; g_2 = 3/2, \; g_3 = 5/3, \; g_4 = 17/10 \, .. \quad (24)$$

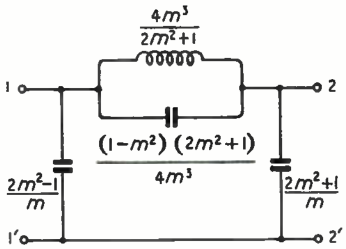


Fig. 6. *m-derived open-circuit filter.*

the recurrence relations

$$g_{i+2} = 2 - f_{i+2} + g_i$$
$$f_i = [g_i(2 - g_i)]^{-1} \quad .. \quad .. \quad (25)$$

where, putting $i = 1, 2 \ldots$ enables one to calculate further elements in both sequences. The detailed proof of (25) is too long and will be omitted. The sequence $f_1, f_2 \ldots$ quickly converges to 2 and the sequence $g_1, g_2 \ldots$ converges to $1 + \sqrt{2}/2$. In addition, equations (25) show that the total shunt capacitance of any filter of the family has the same value as the total shunt capacitance of the corresponding terminated filter.

A similar method of design holds for m-derived filters but the expansion in continued fractions must be made by taking into account the known attenuation poles. As an example the normalized elements of a single-section open-circuit filter are shown in Fig. 6, where the parameter m has the conventional meaning.

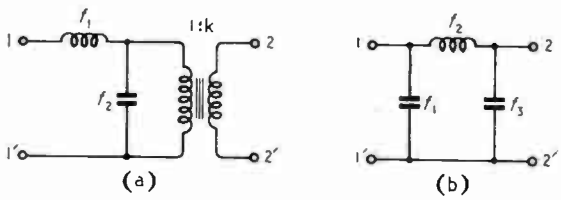


Fig. 7. *Open-circuit filter with Tchebyshev behaviour. Half-section* (a) [*see equ.* (26)] *and entire section* (b) [*see equ.* (27)].

7. Open-Circuit Filters with Tchebyshev Behaviour

The method developed in Section 5 can also be applied to non-conventional filters and enables one to carry out the design, using only rational operations, if the elements of the corresponding terminated filter are known. This is easily applied to filters having constant-k structure and Tchebyshev behaviour. The following design formulae are obtained for the simplest filters

(A) $n = 1/2$ [Fig. 7(a)]

$$f_1 = 1/\sqrt{2}\gamma,\ f_2 = \sqrt{2}\gamma/(\gamma^2 + 1/2),\quad k = e^{-A_m} \qquad (26)$$

(B) $n = 1$ [Fig. 7(b)]

$$f_1 = 1/2\gamma,\ f_2 = 4\gamma/3(\gamma^2 + 1/4),\quad f_3 = (3\gamma^2 + 3/4)/2\gamma(\gamma^2 + 3/4) \quad .. \quad (27)$$

where γ is given by (5).

No simple formulae have been found for the general case, except for the last shunt capacitance C which is important in amplifier circuits. In this case another relation due to Bode permits us to express the normalized value $x = CR\omega_c$ in terms of the frequency integral of the power output, thus

$$\pi/2x = \int_0^\infty e^{-2A} d\Omega \quad .. \quad .. \quad .. \quad (28)$$

For the Tchebyshev characteristic (3), this gives

$$\pi/2x = \int_0^\infty [1 + H^2\phi^2(\Omega)]^{-1} d\Omega \quad .. \quad (29)$$

where ϕ is the polynomial defined by (16). The integral in (29) is easily deduced from (15) by differentiating both members with respect to H and this finally gives

$$x = \frac{(2n+1)\coth \beta/2 \sin \pi/2(2n+1)}{\cosh \beta/2(2n+1)} \quad .. (30)$$

For large values of n, this tends to

$$x = \frac{\pi}{2}\coth \beta/2 = \frac{\pi}{2} e^{A_m} \quad .. \quad .. \quad (31)$$

which is approximately $\pi/2$ for small values of A_m and thus coincides with Bode's limiting value for open-circuit networks.

In Fig. 4(b) x is plotted versus A_m for various values of n. It should be noted that x gives the normalized shunt capacitance at the open-circuited filter end, so that, when comparing with the value of f_2 given in (26), a correction must be applied for the transformer ratio. When comparing the curves of Fig. 4(a) and (b), the 3-db gain due to the open-circuit instead of the matched termination must be taken into account by doubling the values of x in Fig. 4(b) and the curves for $n = 0$ then coincide. For larger values of n the open-circuit networks give a larger x than the terminated filters but have a very poor impedance.

REFERENCES

[1] Darlington, S. "Synthesis of Reactance 4-Poles." *J. of Mathematics and Physics*, September 1939, Vol. 18, pp. 2573–53.
[2] Bode, H. W. "Network Analysis and Feedback Amplifier Design." 1st Edition, Van Nostrand, New York, 1945, Ch. 16.
[3] Fano, R. M. "Theoretical Limitations of the Broadband Matching of Arbitrary Impedances." *J. Franklin Inst.* Jan. and Feb. 1950.

EXHIBITIONS

The Northern Radio Show is being opened at 12 noon on 23rd April. It is being held at the City Hall, Manchester, and will be open until 3rd May.

The 19th National Radio Show will be held at Earls Court, London, from 26th August to 6th September. Admission will be restricted on the opening day.

I.E.E. MEETINGS

9th April. Symposium of papers on microwave links:—"Microwave Radio Links", A. T. Starr, M.A., Ph.D., and T. H. Walker, B.Sc.Tech.; "Circuit Technique in Frequency-Modulated Microwave Links", H. Grayson, B.Sc.Tech., T. S. McLeod, M.A., R. A. G. Dunkley, B.Sc., and G. Dawson, B.Sc.; "Microwave Technique for Communication Links", G. King, L. Lewin, J. Lipinski and J. B. Setchfield.

24th April. Kelvin Lecture. "Iron Atoms in the Service of the Electrical Engineer", Sir Charles Goodeve, O.B.E., D.Sc., F.R.S.

25th April. Discussion on "The Teaching of Transients and the Use of Operational Methods", opened by Instr.-Cdr. D. K. McCleery, M.Sc., R.N., and H. Tropper, Ph.D.

These meetings will be held at the Institution, Savoy Place, London, W.C.2, and will commence at 5.30.

BRIT. I.R.E. MEETING

16th April. "Current Radio Interference Problems", by E. M. Lee, to be held at the London School of Hygiene and Tropical Medicine, Keppel St., Gower St., London, W.C.1, at 6.30.

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.

R.F. Time-Delay Measurements

SIR,—In *Wireless Engineer* for September 1951, Messrs. Naismith and Bramley, in an article entitled "Time-Delay Measurements on Radio Frequencies," describe the results of Loran-type measurements made during 1943-1944 in the U.K. In Fig. 5 of their article is a curve of the most probable values of the transmission delay for night-time E-layer (as used in the Loran system) based on their measurements. It is rather surprising that Messrs. Naismith and Bramley neglected to include a reference from the Radiation Laboratory volume,[1] "Loran," in which the delay curve for measurements taken in the U.S.A. from 1941 to 1944 is given both in graphical form and in an empirical formula. In the accompanying Fig. 1 we have plotted the curve from the Loran book and that of Naismith and Bramley, so that your readers can see how remarkably consistent the transmission delay is, regardless of latitude and season.

We should like to remark here that the delay curve given in the Loran book was used to compile the 'sky-wave corrections' in the charts and tables of the North Atlantic Loran chain; these were placed in service[2] early in 1943. It was also used in the compilation of charts for the S. S. Loran chain which went into operation in 1944 as a further navigational aid for elements of R.A.F. Bomber Command.

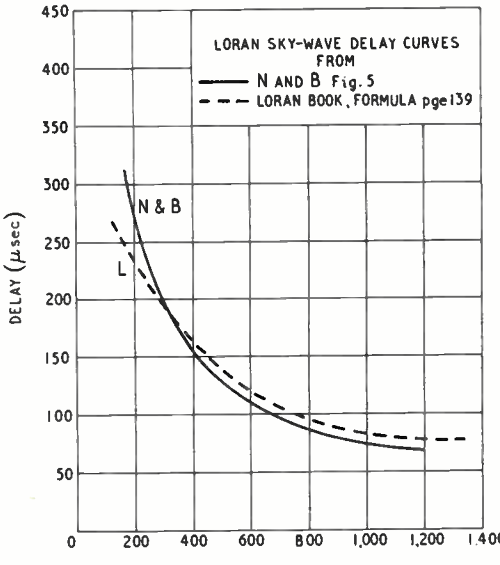


Fig. 1.

In the accompanying Fig. 2 are reproduced the Naismith and Bramley standard deviation curve for a single delay (their Fig. 7) together with the version from the Loran volume. Here the curves are quite different, particularly at short distances. While sky-wave corrections are not tabulated in the tables and charts for short ranges (where ground-wave service is available from both stations of a pair), there may arise occasions when one ground wave may be so weak (owing to a path partially over land) that a navigator may be tempted to match the ground wave from one station against the sky-wave of the other, taking into account the transmission delay. In that case, the standard deviation is a measure of the confidence he can have in his reading. Since there is disagreement on the standard deviation at short distances, it would appear that more measurements might be made to resolve this difference.

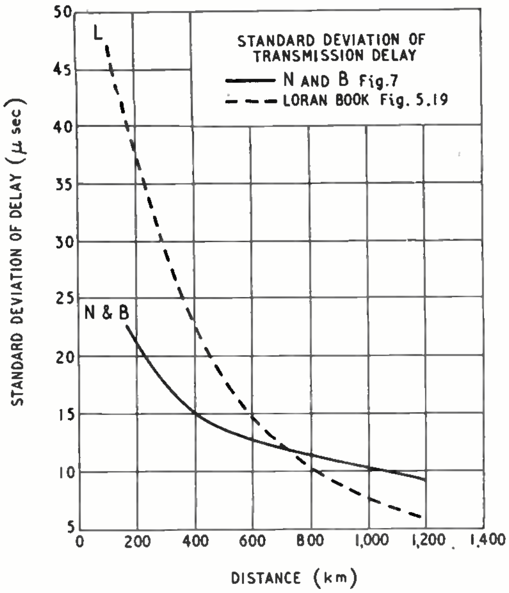


Fig. 2.

During the observations of 16th-24th March 1944, listed by Naismith and Bramley, the writer was at Port Errol assisting in S. S. Loran synchronization tests with the slave station at Bizerte, North Africa. At that time, observations were made at Port Errol of the transmission time to and from North Africa, and it would have been interesting to show them here, for they would have given the asymptotic value of the delay. Unfortunately, the records of these observations were swallowed up in the war-time files of either 60 Group, R.A.F., or the British Branch Radiation Laboratory.

It might also have been profitable had Naismith and Bramley given the relevant magnetic indices (K-values) from Lerwick for their periods of observation. As the Loran book shows in Chapter 5, the delays are affected during high K-values if the reflection point is near or in the auroral belt.

It is a tribute to Naismith and Bramley and their observers that they were able to obtain a fine accumulation of data in spite of often erratic performance of the transmitting stations during some of the trials of 1943 and 1944.

D. DAVIDSON.

Cruft Laboratory,
Harvard University,
Cambridge, Massachusetts, U.S.A.
7th February, 1952.

REFERENCES

[1] Pierce, McKenzie, Woodward. "Loran," Radiation Laboratory Ser., Vol. 4, McGraw-Hill, New York (1948). See Chap. 5, esp. pp. 133-148.
[2] *Ibid.* See pp. 30-34 and Appendix, p. 405.

SIR,—We are grateful to Mr. Davidson for drawing attention to the substantial agreement between the Loran sky-wave delay times as measured in the U.S.A. and in the U.K. In connection with the discrepancy at distances less than 300 km shown in Fig. 1 it is stated in the Loran book that 250 miles (400 km) is the minimum safe distance at which the American delay curve may be used.

The inadequate cross referencing in both British and American published work undoubtedly arose from the incomplete interchange of information under wartime conditions and we, for our part, regret the omission.

One of the more important conclusions reached in the course of our 1943-44 measurements to improve the sky-wave corrections to be used in the Loran charts on this side of the Atlantic was the existence of two effective reflecting regions (A and B in Fig. 6 of our article in the *Wireless Engineer*). We think this feature is of some importance in assessing the merits of this useful navigational aid.

It is difficult to comment on the discrepancy between the figures quoted for the standard deviation of the delay times without more information about the actual technique of observation in the American experiments. It may be noted that the two results quoted on page 140 of the Loran book, which are stated to represent less than the normal scatter, are in good agreement with our values.

R. NAISMITH.
E. N. BRAMLEY.

Radio Research Station,
Slough.
1st March 1952.

Power Meter and Mismatch Indicator

SIR,—In the Appendix to this paper (*Wireless Engineer*, September 1951), the sensitivity of the pick-up loop is expressed in terms of the integral:

$$f(K) = \frac{4}{\pi}\int_0^{\pi/2}[\cos^2\theta/(1 - K\sin\theta)]d\theta$$

The author expands the term $1/(1 - K\sin\theta)$ by the binomial theorem and integrates term by term, obtaining $f(K)$ as a series in powers of K.

This expansion is satisfactory if K is not too large. Thus, for $K = 0{\cdot}5$, the first five terms, written out in the paper, give 1·304, the correct value being 1·309. However, when K approaches 1 the convergence becomes very slow. When $K = 0{\cdot}9$, for example, the first five terms give 1·790, as against a correct value of 2·040. Even using 13 terms the sum reaches only 1·996. In the case of $K = 1$ (when $F = 3{\cdot}273$), the sum of the first 13 terms is only 2·420 (of which the last eight terms contribute only 0·451). Thus, for the higher K values shown in the author's Fig. 8, very many terms of increasingly cumbersome form must have been used, unless the graph was constructed by another method.

However, these questions of convergence can be avoided, since an expression for the integral exists in closed form. It is[1]

$$f(K) = \frac{4}{\pi}\left[\frac{1}{K} + \frac{1}{K^2}\left\{\pi/2 - (\pi/2 + \sin^{-1}K)\sqrt{1 - K^2}\right\}\right]$$

This formula may be conveniently used for all values of K in the range.

R. G. MEDHURST.
J. A. KNUDSEN.

London, S.W.10.
5th March, 1952.

[1] Joseph Edwards. "A Treatise on the Integral Calculus," Vol. 1, Macmillan, 2nd edition, pp. 181 and 195.

NEW BOOKS

Les Tubes Electroniques A Commande Par Modulation de Vitesse

By R. WARNECKE and P. GUENARD. Pp. 776, with 473 illustrations. Gauthier-Villars, 55, Quai des Grands Augustins, Paris. Price 7000 francs.

In a work of seven parts and 37 chapters with a bibliography of 385 references, the authors have set out to present all that is known about velocity-modulation valves, from the point of view both of theory of performance and design and of practical operation and application.

The authors, whose repute in this field is international, base their treatment on the work carried out continuously since 1939 in the laboratories of the Compagnie Générale de T.S.F., but they have drawn freely on all external information accessible to them. The reader is, however, reminded in the Preface that "pour aucun des types de tube à modulation de vitesse, il n'y trouvera une théorie d'ensemble a la fois complète et précise: de telles théories n'existent pas," a statement that will not come as a surprise to those readers who have themselves been occupied with research and development on microwave valves. The treatment adopted is best summarized by the authors' statement that "En s'appuyant continuellement sur des connaissances classiques et en partent de considérations simples pour atteindre progressivement les plus compliquées, les auteurs se sont efforcés de présenter leur exposé d'une manière rationnelle, telle que l'ingenieur et le physician trouvent commodément les informations détaillées dont ils peuvent avoir besoin."

The subject matter of this work is essentially that of the 'classical' velocity-modulation valve where the processes of velocity modulation (and demodulation) and bunching in a drift space take place at differing times and places and are not combined as, for instance, in the travelling-wave tube.

After a historical introduction, the first Part is concerned quite generally with the interaction between an electron and a rapidly-changing field, and with its simplest application in the velocity-modulation valve. A second Part deals with hollow resonant cavities used as circuit elements and a third goes on to consider the elementary theory of operation in the principal types of velocity-modulation valve (two- and three-cavity amplifier, frequency multiplier, reflex oscillator, two-cavity self-oscillator).

The performance characteristics and forms of construction are dealt with in a fourth Part together with the elements of the technology of construction.

A fifth Part is concerned with a more detailed analytical approach to various problems encountered in actual valves, the exchange of energy with non-uniform h.f. fields, effect of space charge in the drift space, the phenomena of hysteresis and noise, etc. It is completed by typical calculations of the design of different types of velocity-modulation valve.

Part six is concerned with problems arising with the use of actual velocity-modulation valves, such as wide-band matching, modulation, the provision of suitable power supplies and includes a section on various methods of measurement.

The final Part surveys the performance realized up to the present time with velocity-modulation valves of the various types and deals with the possibility of further advances in the future.

The book is profusely illustrated and is extremely readable for anyone with moderate experience with the reading of French technical literature in this field. The splitting of the various Parts into Chapters, averaging five per Part, with further sub-division of Chapters into as many as 12 Sections will aid particularly in the use of this work as a book of reference.

The authors are to be congratulated on producing this very comprehensive book on a subject which has grown so very rapidly during the last 20 years, not only for the quality of the material presented, but also on account of the enormous effort which must have been necessary to produce such a work.

W. E. W.

Introduction to Electronic Circuits

By R. FEINBERG, Dr. Ing., M.Sc. Pp. 163 + xiv. Longmans, Green & Co. Ltd., 6 & 7 Clifford Street, London, W.1. Price 18s.

The arrangement of the material of this book follows an unusual pattern and the book is clearly intended for the student with little or no prior knowledge of the subject. In his preface, the author explains his unusual approach and also specifies the background which the student must have. He says, "The intellectual equipment required by the student for penetration into the subject within the scope of this book are: (1) a general conception of electrons, atoms and ions, of the atomic structure of an electric conductor, and of the mechanisms of electron emission from an electric conductor, and (2) familiarity with the method of calculation of a.c. problems with complex quantities, and familiarity with Taylor's and Fourier's series expansion theorems and with simple differential equations of the second order."

He also says that "the subject cannot be fully understood by anyone who has not carried out a course of experimental work with valves." The treatment, therefore, consists of a description of the physical structure of the object with an explanation of how it works, followed by a translation of the mechanism of action into mathematical terms. After this, directions for experiment are given and followed by numerical problems and notes on literature for further reading.

There are eight chapters in the book. The first covers the diode and triode, while the second deals with tetrode, pentode, the basic magnetron and the cathode-ray tube. Chapter III is entitled "Fundamentals of Alternating-Current Amplification" and Chapter IV covers non-linear effects under the sub-sections of diode, triode and hexode. Sinusoidal and relaxation oscillators are discussed in separate chapters, Chapter VII deals with gas-filled valves and Chapter VIII with cold-cathode valves, photo-valves and mercury-pool valves.

The treatment is, on the whole, satisfactory although some will find it rather compressed. The book is only an introduction to the subject and a great deal is covered in its 163 pages. Inevitably in such a book one is conscious of things which are missing and of wishing that the author had gone rather further. One would still feel the same if he had, however.

There are a few peculiar things in the book which will perhaps not hinder the real beginner but which make things unnecessarily difficult for the reader who does have some prior knowledge. One is the strange fashion of drawing anode-volts—anode-current characteristics with the horizontal scale for current and the vertical for voltage. The curves have a very odd appearance. The author does this quite a bit, but not always.

A convention which, again, is confusing is the use of arrows to indicate the polarity of voltages. The arrow is drawn alongside the source of e.m.f., or the potential difference, with its head pointing to the negative end. Arrows are also used for current with the result that a circuit diagram carries quite a bewildering array of arrows and the reader has to sort out which are which. It would have been much clearer if the author had adopted the usual convention of confining arrows to the indication of the direction of current flow, and of indicating polarity by + and — signs.

These are, however, minor points in an excellent book.

W. T. C.

The Measurement of Radio Isotopes

By DENIS TAYLOR, M.Sc., Ph.D., M.I.E.E., F.Inst.P. Pp. 118 + viii with 40 illustrations. Methuen & Co., Ltd., 36 Essex St., London, W.C.2. Price 6s. 6d.

This little book contains three chapters covering apparatus for measuring radioactivity and the methods employed for counting. In addition there are chapters on source geometry, statistics, correction factors and health hazards.

Principles of Alternating Currents

By W. SLUCKIN, B.Sc.(Lond.), B.Sc.Eng.(Lond.). Pp. 320 with 165 illustrations. Cleaver-Hume Press, Ltd., 42a South Audley St., London, W.1. Price 10s. 6d.

Electrical Instruments and Measurements

By W. ALEXANDER, M.Sc., B.Sc.(Eng.), M.I.E.E. Pp. 352 with 112 illustrations. Cleaver-Hume Press, Ltd., 42a South Audley St., London, W.1. Price 12s. 6d.

Bases Techniques de la Télévision

By H. DELABY. Pp. 340 with 116 illustrations. Les Editions Eyrolles, 61 boulevard Saint-Germain, Paris Ve. Price 2,200 francs.

Cour de Radioélectricité Générale

Book 3, Vol. 1. L'Emission. By ROGER RIGAL. Pp. 139 with 164 illustrations. Price 950 francs.

Book 3, Vol. 2. La Réception. By PIERRE DAVID. Pp. 355 with 156 illustrations. Price 2,200 francs.

Les Editions Eyrolles, 61 boulevard Saint-Germain, Paris Ve.

Exercises de Radioélectricité (Lignes-Antennes-Hyperfréquences)

By S. ALBAGLI. Pp. 76. Gauthier-Villars, 55 Quai des Grands-Augustins, Paris. Price 550 francs.

Semi-Conducting Materials

Edited by H. K. HENISCH. Pp. 281. Butterworth Scientific Publications, Ltd., Bell Yard, Temple Bar, London, W.C.2. Price 40s.

This book forms the Proceedings of a Conference held at the University of Reading under the auspices of the International Union of Pure and Applied Physics in co-operation with the Royal Society. It comprises 25 papers with which are provided summaries in French.

Tables of n! and $\Gamma(n + \frac{1}{2})$ for the First Thousand Values of n

By HERBERT E. SALZER. National Bureau of Standards Applied Mathematics Series 16, iii. Pp. 10. National Bureau of Standards, U.S. Department of Commerce, Washington 25, D.C., U.S.A. Price 15 cents (postage 5 cents).

In this booklet values of $n!$ are given to 16 significant figures and of $\Gamma(n + \frac{1}{2})$ to eight significant figures.

Materials Technology for Electron Tubes

By WALTER H. KOHL. Pp. 493 + xv. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 80s.

This is an American book and has 19 chapters. The first six are devoted to some aspect of glass—physics, annealing, strain analysis, electrical conduction, radiation fields and glass-to-metal seals. Chapter 7 deals with electrons, atoms, crystals and solids. Six further chapters cover the conductors, tungsten, molybdenum, tantalum, nickel, copper, carbon and graphite.

The joining of metals by soldering and brazing has a chapter and others cover ceramics and mica and ceramic-to-metal seals. Chapters on high-vacuum technique and thermionic emission conclude the book.

The book is intended for the "tube designer, development engineer and technician alike."

The Oxide-Coated Cathode

By Dr. Ing. G. Hermann and Dr. Phil. S. Wagener. Translated by Dr. Phil. S. Wagener. Vol. 1—Manufacture. Pp. 148 + viii with 78 illustrations. Price 21s. Vol. 2—Physics. Pp. 311 + xiv with 154 illustrations. Price 42s. Chapman & Hall, Ltd., 36 Essex Street, London, W.C.2.

In Vol. 1 there is an historical review of the oxide cathode and a discussion of its different forms and their applications. The manufacturing processes and cathode characteristics are then discussed. Special types for gas-filled valves are included.

Vol. 2 deals with the physics of the cathode and starts with the thermal emissions of electrons from metals, and the measurement of work function. Phenomena in ionic solids and the mechanism of the emission from an activated oxide coating in equilibrium are then treated. Concluding chapters cover various forms of oxide coatings and the effect of variations in the equilibrium of the coatings.

TECHNICAL PUBLICATIONS

Units and Standards of Measurement: 1. Length, Mass, Time.

Pp. 12. Published for the Department of Scientific and Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 9d. (25 cents, U.S.A.).

The booklet deals with the fundamental units, length, mass and time, as well as the derived units, volume, density, specific gravity, gravity, force and pressure. It describes the units and standards of measurement used at the National Physical Laboratory.

In order to avoid confusion between the units of force and mass the symbols Lb and Kg are used to denote the units of force which are the weights of one pound and one kilogramme respectively. For the units of mass the usual symbols lb and kg are used.

British Standard for Dimensions and Nominal Voltages of Batteries for Valve-Type Hearing Aids (B.S.966:1951)

Revision of standard first published in 1949. British Standards Institution, Sales Dept., 24 Victoria St., London, S.W.1. Price 2s. (including postage).

Proceedings of a Conference on Centimetre Aerials for Marine Navigational Radar

Held 15th-16th June 1950, in London. Pp. 150 + vi. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 15s.

Thirteen papers are included with the discussions on them and cover the three sessions of the Conference—Shipborne Aerials for Navigational Radar, Shore-Based Aerials and Aerial Measurement Techniques.

Radio Direction Finding and Navigational Aids

Radio Research Special Report No. 21. Pp. 92 + iv. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 3s. 6d.

Nine papers are included which are translations of certain German reports captured in 1945. All but two formed part of the proceedings of an official German conference on radio navigational aids held at Landsberg in March 1944. The remaining two are additional and later German reports.

A Single Frequency Instrument for the Measurement of Interference with Television Reception due to Ignition Systems

Technical Report M/T102. Pp. 5. The British Electrical and Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 3s. (postage 3d.).

Report of the Radio Research Board 1950

Pp. 51 + iv. H.M. Stationery Office, Kingsway, London, W.C.2. Price 1s. 9d. (U.S.A. 45 cents).

This report describes the work carried out during 1950 by the Department of Scientific & Industrial Research on problems concerning radio propagation through the upper and the lower atmosphere, direction finding, navigational aids and radio noise. It covers also investigations into the electrical properties of germanium and into v.h.f. measuring techniques.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for February, 1952

Date 1952 February	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1029-1130 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1*	− 0·4	+ 2	− 30·2
2	− 0·3	+ 2	− 31·0
3	− 0·3	+ 2	− 32·0
4	− 0·3	+ 1	− 33·8
5	− 0·6	+ 1	− 35·6
6	− 0·3	+ 2	− 36·5
7*	− 0·2	+ 2	− 37·2
8*	− 0·2	+ 3	N.M.
9	0·0	+ 2	− 39·2
10	0·0	+ 2	N.M.
11**	—	+ 3	—
12*	0·0	+ 3	− 44·0
13*	+ 0·1	+ 4	− 44·8
14*	+ 0·2	+ 4	− 46·7
15*	+ 0·4	+ 4	− 46·9
16	+ 0·3	+ 4	− 47·3
17	+ 0·3	+ 4	− 48·5
18	+ 0·3	+ 4	− 49·8
19*	+ 0·4	+ 4	− 50·9
20	+ 0·4	+ 4	− 51·6
21	+ 0·4	+ 5	− 52·3
22*	+ 0·5	+ 5	− 53·7
23	+ 0·4	+ 5	N.M.
24	+ 0·4	N.M.	− 53·2
25	+ 0·5	+ 6	N.M.
26*	+ 0·6	− 4	− 53·3
27*	+ 0·6	− 4	− 53·7
28**	—	− 3	—
29**	—	− 3	—

The transmitter employed for the MSF 60-kc/s signal is sometimes required for another service.

N.M. = Not measured.

* = No MSF transmission at 1029 G.M.T. Results for 1429-1530 G.M.T.

** = No MSF transmission at 1029 G.M.T. or at 1429 G.M.T.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research and published by arrangement with that Department.

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 the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

ACOUSTICS AND AUDIO FREQUENCIES

519.272.11 : [534.874.1 + 621.396.62 **882**

The Correlation Function in the Analysis of Directive Wave Propagation.—H. Nodtvedt. (*Phil. Mag.*, Sept. 1951, Vol. 42, No. 332, pp. 1022–1031.) Correlation-function theory is useful for investigating directivity in the reception of signals with a broad spectrum. Conditions are analysed for systems comprising (*a*) two non-directional receivers, and (*b*) two directional receivers. Results are shown graphically and compared with those obtained by the phase-comparison method. Increase of signal bandwidth leads to reduction of side lobes, but the amount of the increase required for a substantial effect is too high for the method to be of practical importance in h.f. problems, though it is of interest in sound reception.

534 + 621.395.61/.62](083.71) **883**

I.R.E. Electroacoustics Standards: Genesis of the Glossary of Acoustical Definitions.—(*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, p. 1567.) Corrections are given to I.R.E. Standards on Electroacoustics (2310 of 1951).

534.012 **884**

Longitudinal Modes of Elastic Waves in Isotropic Cylinders and Slabs.—A. N. Holden. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 956–969.) The general properties of the longitudinal vibration modes are derived, making use of the close formal analogy between the dispersion equations for slabs and cylinders.

534.213 **885**

The Propagation of Sound Waves through a Medium with very Small Random Variations in Refractive Index.—T. H. Ellison. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 14–21.) "Mathematical techniques, developed for the theory of turbulence, are applied to the study of the propagation of waves in a medium in which the refractive index varies slightly with position in a random manner. Both a diffraction and a ray theory are used to relate the statistical properties of the wave to those of the refractive-index field, and the conditions for the validity of the ray theory are exposed. A particular result is, that, for a wave travelling in a statistically homogeneous medium where the scale of the variations in refractive index is suitably large compared with the wavelength but small compared with the path length, the variations in intensity produced by the medium are proportional to the cube of the path length for short paths, but directly proportional to it for long paths."

534.3 : 786.6 **886**

Organ Acoustics.—F. Trendelenburg. (*Ricerca sci.*, March 1951, Vol. 21, No. 3, pp. 339–352.) A review of German research work covering (*a*) the various types of organ pipe and their characteristic frequency spectra, (*b*) the differences between the sound quality of organs of the Baroque period and of modern organs, (*c*) initial transients and their acoustic effect, (*d*) the relation between the acoustics of churches and organ music style.

534.321.9.012 : [539.32 : 546.74 **887**

Frequency Dependence of Elastic Constants and Losses in Nickel.—R. M. Bozorth, W. P. Mason & H. J. McSkimin. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 970–989.) An ultrasonic pulse method was used to measure the elastic constants of single crystals of Ni and their variation with magnetic field. Measurements were also made of the velocity and attenuation of elastic vibrations in well-annealed polycrystalline Ni rods over the frequency range 5–150 kc/s. The elastic constants vary with domain distribution in the case of demagnetized crystals. The small variation with magnetic field observed at the frequency used (10 Mc/s) is due to a relaxation in the domain-wall motion due to micro-eddy-current damping. The average domain size was found to be about 0·04 mm.

534.85 : 681.84 **888**

New Lightweight Pickup and Tone Arm.—L. J. Anderson & C. R. Johnson. (*Broadcast News*, May/June 1951, No. 64, pp. 8–14.) Illustrated description of R.C.A. Type-MI-11874 pickup and Type-MI-11885 arm for use at broadcasting stations and suitable for disk speeds of 33⅓ or 78 r.p.m. (ordinary groove) and 33⅓ or 45 r.p.m. (fine groove).

534.862.4 **889**

The Reproduction of Magnetically Recorded Signals.—R. L. Wallace, Jr. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1145–1173.) "It has been found experimentally and theoretically that introducing a spacing of *d* inches between the reproducing head and the recording medium decreases the reproduced voltage by 54·6 (d/λ) decibels when the recorded wavelength is λ inches. For short wavelengths this loss is many decibels even when the effective spacing is only a few ten-thousandths of an inch. On this basis it is argued that imperfect magnetic contact between reproducing head and recording medium may account for much of the high-frequency loss which is experimentally observed." Mathematical appendices deal with (*a*) the field due to a flat sinusoidally magnetized medium, (*b*) the field due to a round wire.

621.395.61/.62 **890**

Crystal-Transducer Response.—B. J. Shelley. (*Electronic Engng*, Sept. 1951, Vol. 23, No. 283, pp. 353–354.) Over the a.f. range crystal transducers can have identical response curves for either direction of operation. By using two such crystals coupled to a common cavity the frequency response curves for the individual crystals can be obtained by combining the measured sum and difference response curves.

621.395.61 : 621.396.645.029.3 **891**

Construction of A.F. Amplifiers for Studio Equipment.—Schiesser & Gathmann. (See 950.)

621.395.61 : 621.396.645 **892**

Equipment for Acoustic Measurements: Part 1—A Portable General-Purpose Microphone Amplifier using Miniature Valves.—Shorter & Beadle. (See 946.)

621.395.616 **893**

New Capacitor Microphones for Broadcasting Studios.—H. Grosskopf. (*Fernmeldetech. Z.*, Sept. 1951, Vol. 4, No. 9, pp. 398–402.) The bearing of acoustic and transmission conditions on the design of microphones is discussed. Two new capacitor types developed by the North-West German Broadcasting Organization are described, viz., the M49 gradient type with continuously variable directivity, and the M50 pressure type.

621.395.616 : 534.612.4 **894**

Condenser Microphone Sensitivity Measurement by Reactance-Tube Null Method.—H. E. von Gierke & W. W. von Wittern. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, p. 1534.) Correction to paper abstracted in 2624 of 1951.

621.395.623.7 : 621.396.645 **895**

Cathode-Follower Loudspeaker Coupling.—Fletcher & Cooke. (See 947.)

621.395.623.74 : 537.58 **896**

The Ionic Loudspeaker.—S. Klein. (*TSF pour Tous*, Sept. 1951, Vol. 27, No. 275, pp. 278–281. Erratum, *ibid.*, No. 276, p. 342.) Description of a loudspeaker with no moving parts, developed from the thermionic cell previously described (593 of 1947 and 288 of January). The ion source consists of a Pt wire inserted in a small tube of refractory material, such as quartz or porcelain, covered with a positive-ion emitting layer formed by a mixture of precipitated Pt, $AlPO_4$, Ir and graphite. This tube is fixed in the throat of an exponential horn of fused quartz with double walls, the space between the walls being evacuated. Surrounding the whole is a grounded cylindrical electrode. When a h.f. h.v. source (400 kc/s; 10 kV) is connected to the Pt wire, after about a minute the wire is maintained at a temperature of about 1 000°C, with consequent positive-ion emission from the coating on the surrounding tube and intense ionization of the air molecules in the throat of the horn. Modulation of the h.f. voltage thus results in modulation of the ionization of the air, and sound waves are emitted. The frequency/response curve given shows relatively small variations of level from 30 c/s to over 10 kc/s. The device can also be used as a microphone. Circuit diagrams are given for the loudspeaker and microphone arrangements, with component details for the h.f. h.v. generator.

621.395.623.74 : 537.58 **897**

The Physical Principles of the Ionic Loudspeaker and Microphone.—S. Klein. (*TSF pour Tous*, Oct. 1951, Vol. 27, No. 276, pp. 340–342.) Discussion, with numerical calculations, of (*a*) the effect of the high temperature on the air in the throat of the horn (896 above), (*b*) the effect of the accelerating voltage on the positive ions, (*c*) ionic mobility in the rarefied air in the horn throat, (*d*) the effect of frequency.

621.395.623.74 : 537.58 **898**

The Ionophone.—M. Bonhomme. (*Toute la Radio*, Oct. 1951, No. 159, pp. 251–255.) General description of Klein's ionic loudspeaker (896 above), with diagrams and photographs. See also *Wireless World*, Jan. 1952, Vol. 58, No. 1, pp. 2–3 (F.L.D.) and *Radio-Electronics*, Nov. 1951, Vol. 23, No. 2, pp. 44–45 (Aisberg & Bonhomme).

621.395.625.2 **899**

A Disk Reproducing Desk.—R. D. Petrie. (*B.B.C. Quart.*, Summer 1951, Vol. 6, No. 2, pp. 105–112.) The basic requirements for this equipment are outlined and the electrical and mechanical design adopted to meet them are described. The output from the sapphire stylus pickup, after equalization to suit the frequency characteristic of the disk being played, is amplified and feeds a star mixer network. The output amplitude/frequency characteristic when playing a B.B.C. record is flat to with ± 1 db from 50 c/s to 10 kc/s. The aluminium turntable is rim driven, speeds of 78 and 33⅓ r.p.m. being available. Several mechanical features are incorporated to prevent damage when changing disks.

621.395.625.3 **900**

An Editing Machine for Magnetic Tape Recording.—D. C. Yarnes. (*Broadcast News*, May/June 1951, No. 64, pp. 66–69.) Description of a professional-quality machine with facilities for the rapid marking, cutting and splicing of the tape.

621.395.625.3 **901**

Magnetic Processes in Tapes and Magnetic Heads of Magnetic Sound-Recording Apparatus.—W. Puhlmann. (*Funk u. Ton*, Feb. 1951, Vol. 5, No. 2, pp. 65–75.) The h.f. bias technique for measurement of the static remanence characteristic of tape materials is described. Advantages of the powder-coated tape, and the selection of suitable magnetic materials are discussed. The 'internal' demagnetization effected in the coated tape reduces the loss of sensitivity at high frequencies normally occurring after wiping. Losses due to the air gap in the magnetic head are calculated and noise is discussed with reference to the irregular distribution of magnetic particles.

621.395.665.1 **902**

Automatic Control of [volume] **Contrast.**—Balz. (See 940.)

621.395.92.001.4 **903**

Measurements on Hearing Aids.—F. Müller. (*Funk u. Ton*, July 1951, Vol. 5, No. 7, pp. 361–368.) An account of the fundamentals of measurement technique regarding frequency range, amplification, sensitivity and distortion.

621.395.92.001.4 **904**

Testing of Hearing Aids.—F. Müller. (*Funk u. Ton*, Aug. 1951, Vol. 5, No. 8, pp. 400–410.) An outline of all the measurements required to determine the characteristics of hearing aids, with illustrative examples. See also 903 above.

789.983† : 621.396.615.029.3 **905**

Recent Design Developments in Electronic-Organ Tone Generators.—S. L. Krauss & C. J. Tennes. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 344–347.) Pentode and triode circuits are described which generate simultaneously a sinusoidal and a pulse wave, this combination having been found particularly effective for the production of the tone qualities required for the different organ voices.

534.232 **906**

Die elektroakustischen Wandler (Electroacoustic Transducers). [Book Review]—W. Furrer. Publishers: J. A. Barth, Leipzig, 2nd edn 1951, 221 pp., 22.30 Swiss francs. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Sept. 1951, Vol. 29, No. 9, p. 358.) Additions and corrections to the first edition, published in 1940, are included; the Giorgi system of units is used. Both sonic and ultrasonic devices are considered. The book will be useful for both students and practising engineers.

AERIALS AND TRANSMISSION LINES

621.392 : 621-231.221 **907**

Microwave Rotating Joints.—J. P. Grantham. (*Electronic Engng*, Sept. & Oct. 1951, Vol. 23, Nos. 283 & 284, pp. 332–335 & 377–381.) Most of the known types of rotating joint for circular-section waveguides and coaxial lines are described. Data are furnished which, together with information available in the works noted in the bibliography, should enable a rotating joint for a specific purpose to be designed with little experimental work. Coaxial-line joints are the most compact and easiest to design of all the types described; they are free from resonance effects, can cover very wide bands, and have adequate power-handling capacity. The 'doorknob' transformer is very suitable for rotating-joint applications; a truncated-cone type of 'doorknob' used with a semicircular piston has good wide-band properties without the use of matching irises. A form of circular-waveguide joint using resonant-ring H_{11}-mode suppressors and cavities is free from resonances, but is less compact than the coaxial-line joint, requires closer manufacturing tolerances, and is more susceptible to voltage breakdown.

621.392.09 **908**

Advances in the Theory of Waves on Wires.—H. Kaden. (*Arch. elekt. Übertragung*, Sept. 1951, Vol. 5, No. 9, pp. 399–414.) The theory of Harms and Goubau for surface waves on wires is extended to apply to magnetically loaded wires produced either by coating the wire with ferrite or by inserting a ferrite core in the coiled wire. By using these methods of concentrating the field, the upper wavelength limit can be raised from 1 to 10 m. The field structure is characterized by a parameter called 'critical radius', which increases with wavelength. Characteristic impedance is proportional to the logarithm of critical-radius/wire-radius. The concept of 'equivalent coaxial line', i.e., that having the same characteristic impedance and attenuation constant as the wire, facilitates investigation of the coupling between parallel lines. Values of parameters are tabulated for particular lines at various wavelengths, and their performance is compared with that of coaxial lines.

621.392.21 + 621.392.26† **909**

Simple Branch Connections of Lines and Waveguides.—H. H. Meinke. (*Fernmeldetech. Z.*, Sept. 1951, Vol. 4, No. 9, pp. 385–388.) Whereas at frequencies below 100 Mc/s circuit branching presents no special problems, at decimetre wavelengths certain unwanted reactances become appreciable, and at centimetre wavelengths the difficulties are even greater. The problem is surveyed for three-line junctions, series and parallel arrangements being distinguished; a selected bibliography is given.

621.392.26† **910**

The Offset Wave-Guide Junction as a Reactive Element.—L. D. Smullin & W. G. Glass. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1124–1127.) These junctions are characterized by rugged mechanical construction and ease of adjustment. Simple analytical expressions are derived for some of the more important properties. The capacitive junction allows the practical construction of capacitively coupled resonant cavities; these have a more constant bandwidth over their tuning range than can be obtained with inductively coupled cavities.

621.392.26† **911**

Theory of Space-Charge Waves in Cylindrical Waveguides with Many Beams.—P. Parzen. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 217–219.) The effect on gain and bandwidth due to the separation of the beams and the presence of the surrounding guide is derived approximately by an integral-equation method. The mechanism of energy transport is such that the input and output circuits required would be similar to those of a klystron rather than a travelling-wave valve. Cases given special attention are: (*a*) two thin annular beams within a circular guide, (*b*) a homogeneous mixture of two beams filling the guide.

621.392.26† : 538.61 **912**

Magneto-optics of an Electron Gas with Guided Microwaves.—L. Goldstein, M. A. Lampert & J. F. Heney. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 233–234.) Reprint. See 2911 of 1951.

621.392.26† : 538.61 **913**

Magneto-optics of an Electron Gas with Guided Microwaves.—L. Goldstein, M. Lampert & J. Heney. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, p. 1053.) Correction to paper abstracted in 2911 of 1951.

621.392.26† : 538.61 **914**

Magneto-optics of an Electron Gas for Guided Microwaves: Propagation in Rectangular Wave-Guide.—L. Goldstein, M. Lampert & J. Heney. (*Phys. Rev.*, 15th Sept. 1951, Vol. 83, No. 6, p. 1255.) The experimental arrangement was similar to that described in 2911 of 1951 for a circular waveguide. In the present case the guide without the electron gas supports only the TE_{10} mode at the frequency used, hence no polarization transformation can occur. Significant effects were observed at values of magnetic field such that electron gyrofrequency and signal frequency are nearly equal.

621.392.26† : 621.392.09 **915**

The Propagation of Waves in Cylindrical Waveguides and the Hertzian Solution as Special Cases of the Propagation of Waves in Horns.—H. Kleinwächter. (*Arch. elekt. Übertragung*, Sept. 1951, Vol. 5, No. 9, p. 439.) Explanatory note on 320 of February.

621.396.67 **916**

An Aerial for V.H.F. Broadcasting.—G. D. Monteath. (*B.B.C. Quart.*, Summer 1951, Vol. 6, No. 2, pp. 122–128.) Illustrations and description of the aerial system installed at Wrotham. See also 2340 and 2643 of 1951 (Gillam).

621.396.67 **917**

The Measurement of Current Distributions along Coupled Antennas and Folded Dipoles.—T. Morita &

C. E. Faflick. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1561–1565.) The theoretical analysis of coupled aerials is facilitated by representing the current in each element as a superposition of components in opposite directions, the components in one direction constituting the radiating currents which maintain the far-zone field, while those in the other direction are essentially non-radiating currents. Experiments on closely coupled aerials and folded dipoles are described which confirm the correctness of the above representation.

621.396.67 **918**

Theory of V-Antennas.—R. King. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1111–1121.) An integral equation for the current in an apex-driven symmetrical V-aerial is derived and solved by successive approximations. General formulae for the distribution of current and the impedance are obtained. The zero-order impedance is evaluated for a leg length $h = \lambda_0/4$, as a function of the angle between the legs; the results are compared with measured values, and experimental impedance curves for various angles of lateral and forward tilt are discussed.

621.396.67.029.62/.63 **919**

Omnidirectional Aerials with Horizontal Polarization for Wide Frequency Bands in the Metre- and Decimetre-Wave Region.—K. Lamberts. (*Frequenz*, July 1951, Vol. 5, No. 7, pp. 177–185.) Experimental investigation of two types of aerial consisting of four flat circular disks arranged (*a*) as two horizontal dipole elements crossed at right angles, (*b*) with the disks vertical and forming a square with length of side effectively $\lambda/2$. In case (*a*), the total input impedance is relatively constant but the radiation pattern is poor; the arrangement is not particularly suitable for wide-band operation. In case (*b*), the input impedance varies with frequency but the radiation characteristic is nearly circular for frequencies within $\pm$ 20% of the centre frequency.

621.396.671 **920**

On Zuhrt's Theory of Dipole Aerials.—R. King: H. Zuhrt. (*Frequenz*, Aug. 1951, Vol. 5, No. 8, pp. 219–223.) King gives a critical analysis of Zuhrt's method of calculating the current distribution and input impedance of a symmetrical tubular radiator (2977 of 1950). It is pointed out that the current at the open end of the aerial does not vanish, but only changes direction. Evaluation of the Fourier series to only the third term limits the application of the method to aerials sufficiently thin and short or of length very near that for series resonance. Measurements substantiate these conclusions.

Zuhrt states that a correction for end effects is already included in the manuscript of a forthcoming book, and suggests a possible anomaly in the measurements mentioned by King.

621.396.676 **921**

Glide-Path Cavity Antenna for Jet Fighter Aircraft.—L. E. Raburn. (*Tele-Tech*, Sept. & Oct. 1951, Vol. 10, Nos. 9 & 10, pp. 32–33, 76 & 50–51, .91.) Radiation-pattern measurements on a one-tenth scale model were confirmed by results obtained with an experimental mock-up of a cavity installed in the lip of the air intake. The aerial is horizontally polarized, receives signals of frequency 329–335 Mc/s from any forward direction, and has a voltage s.w.r. of over 5 : 1.

621.396.677 **922**

Corrugated End-Fire Antennas.—D. K. Reynolds & W. S. Lucke. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 16–28.) The effect of the surface parameters in modifying the guided-wave propagation over a corrugated surface is discussed in relation to the attainment of the optimum end-fire radiation pattern. The effects on field pattern and frequency bandwidth of various methods of driving such aerials are compared. Practical forms of corrugated aerial are described and theoretical and measured field patterns are compared.

621.396.677.3† **923**

General Formulation for Calculation of Shaped-Beam Antennas.—A. S. Dunbar. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, p. 1217.) Outline of a method based on Chu's synthesis of a shaped beam by the controlled variation of phase in an aperture having a known amplitude distribution. Using the method of stationary phase the far-field intensity is formally derived and illustrated by the case of a line source with cylindrical reflector.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.3† **924**

Noise Figure of the Magnetic Amplifier.—N. R. Castellini. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 52–58.) An expression for the noise figure is derived by using an equivalent circuit which takes into account winding and leakage inductances and stray capacitance. The effect of the Barkhausen and thermal-noise currents is considered. The noise figure varies inversely as the amplification factor and operating frequency. The calculated mean output current at threshold agrees well with experimental values.

621.314.3† **925**

Magnetic Amplifiers with Orthonol-Tape Cores.—W. A. Geyger. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 59–67.) Performance figures for these amplifiers, which use multi-layer toroidal windings on Orthonol-tape cores, show that far superior characteristics and greater power output are obtained when this rectangular-hysteresis-loop core material is used.

621.314.3† **926**

Transient Response of Magnetic Amplifiers.—L. A. Finzi, D. P. Chandler & D. C. Beaumariage. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, p. 809.) Summary of 1951 A.I.E.E. Great Lakes District Meeting paper.

621.314.3† **927**

The Time Constant of a Magnetic Amplifier.—H. Goldstein. (*Funk u. Ton*, Feb. 1951, Vol. 5, No. 2, pp. 76–78.) An approximate value of the time delay, τ, for build-up in the d.c. windings, is given by $\tau = \beta.V/\omega$ where β is the ratio of the currents in the a.c. circuit with and without d.c. premagnetization, and V the amplification factor.

621.318.572 **928**

The Counting of Random Pulses.—E. H. Cooke-Yarborough. (*J. Brit. Instn Radio Engrs*, Sept. 1951, Vol. 11, No. 9, pp. 367–380. Discussion, *ibid.*, Dec. 1951, Vol. 11, No. 12, p. 594.) 1951 Radio Convention (London) paper, surveying electronic and electro-mechanical methods of counting pulses. Numerous references are given to published circuits.

621.318.572 : 621.385.38 **929**

The Thyratron as a Close-Differential Relay.—J. J. Baruch. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 87–93.)

621.318.572 : 621.396.615.17 **930**

Linear Analysis of Electronic Switching.—R. Ahmed. (*Indian J. Phys.*, July 1950, Vol. 24, No. 7, pp. 281–290.) The multivibrator circuit is considered; assumptions are made leading to a simple equivalent circuit which can be treated by linear analysis, enabling the switching time to be calculated with fair accuracy. Criteria are established for designing fast switching circuits.

621.318.572 : 621.396.615.17 **931**

A Study of the Switching Action in a Multivibrator Circuit: Part 1.—B. M. Banerjee. (*Indian J. Phys.*, Aug. 1950, Vol. 24, No. 8, pp. 361–370.) Description of c.r.o. equipment for displaying the rapid changes in valve electrode voltages which occur during conduction transfer from one valve to the other of a multivibrator. Oscillograms obtained with different values of the circuit parameters are reproduced.

621.319.4 **932**

Electrical Charge Storage in Polystyrene Capacitors.—L. A. Matheson & V. J. Caldecourt. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1176–1178.) Measurements on charge storage in polystyrene-film capacitors show that charge 'soakage' is very small and that charges may be retained for periods of the order of 100 years. The specific resistance of the insulation approaches 10^{22} Ω.cm over a period of months.

621.319.45 **933**

Electrolytic Capacitors.—G. W. A. Dummer. (*Wireless World*, Dec. 1951, Vol. 57, No. 12, pp. 510–512.) Principles of operation are stated, and the advantage of the high capacitance volume ratio is stressed. Present-day types are classified according as they use a plain-foil, etched-foil or sprayed-gauze anode. Methods of increasing anode surface area and the use of new materials, especially tantalum, have been investigated.

621.392.5 **934**

A Note on the Initial Excitation of Linear Systems.—L. A. Zadeh. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1216–1217.) The formula for the system response, previously given for the case where the initial values of input and output voltages and their derivatives are specified at the instant immediately following the application of the input voltage (2668 of 1951), is modified so as to apply to the case where the initial values are specified at the instant immediately preceding the application of the input voltage.

621.392.5 **935**

Graphical Study of Quadripoles at Constant Frequency.—J. Schérer. (*Rev. gén. Élect.*, Sept. 1951, Vol. 60, No. 9, pp. 359–370.) A relatively simple method is developed from the geometrical properties of the nomographic transformation relating the input impedance of a quadripole to the load impedance, making use of the basic open-circuit and short-circuit impedances or admittances and of the constructions of elementary plane geometry. Of the numerous possible applications of the method, the determination of the impedance of an iterative network and the method of transforming a quadripole are discussed in particular.

621.392.5 : 512.831 **936**

Quadripoles and Matrices.—H. G. Möller. (*Elektrotechnik, Berlin*, Sept. 1951, Vol. 5, No. 9, pp. 426–430.) The use of matrix methods in circuit analysis is illustrated with reference to transformers, Lecher-wire systems, filter networks, delay circuits, loudspeakers, amplifier valves and detectors.

621.392.52 **937**

The Reduction of Branched Circuits and its Application to Ladder Arrangements of Quadripoles.—F. M. Pelz. (*Funk u. Ton*, Feb. 1951, Vol. 5, No. 2, pp. 57–64.) Klein's method of determinantal transformation (2999 of 1950) is used to derive a network transformation of star-delta type. Simplification of the reduced circuit is achieved by introducing a given relation between the values of the elements of the original circuit. Application of the method to the design of a band-pass filter is illustrated.

621.392.52 **938**

Determination of Input Impedance and Other Properties of Networks.—E. Green. (*Marconi Rev.*, 4th Quarter 1951, Vol. 14, No. 103, pp. 141–155.) Parameters of many classes of network can be expressed in terms of frequency response where this is known exactly. Formulae given by Dishal (3369 of 1949) in relation to band-pass networks are used for this purpose.

621.392.52 **939**

The Transfer Impedance of Recurrent Π and T Networks: Part 2.—J. B. Rudd. (*A.W.A. tech. Rev.*, Sept. 1951, Vol. 9, No. 2, pp. 67–72.) An expression is derived for the transfer impedance of an N-section chain of symmetrical Π or T sections with equal resistance terminations. The impedance can be expressed in terms of Tchebycheff polynomials of the first and second kinds. Expressions are also derived for the insertion loss and insertion phase shift of ladders containing N sections with purely reactive elements. Part 1: 657 of 1947.

621.395.665.1 **940**

Automatic Control of [volume] Contrast.—G. Balz. (*Funk u. Ton*, Feb. 1951, Vol. 5, No. 2, pp. 79–90.) The theory of the basic hexode volume-expansion circuit is developed, with particular attention to the time constants and the characteristic of the control circuit. In an automatic compression or expansion circuit designed to simulate the operation of a manual system, an additional control voltage depending on the programme volume is derived from an auxiliary slow-acting control circuit and applied to the regulator valve in series with the usual control voltage. In the case of recorded sound the additional control is derived from an auxiliary pickup operating in advance of the main pickup.

621.396.6.015.7 **941**

Speed of Response of the Cathode-Coupled Clipper.—P. F. Ordung & H. L. Krauss. (*TV Engng, N.Y.*, April & June 1951, Vol. 2, Nos. 4 & 6, pp. 22–24, 32 & 21, 27.) When essentially rectangular pulses were applied to a cathode-coupled clipper circuit, the build-up time of the output voltage was about 50 mμs, either with or without regeneration. Analysis of the effect of variation of the circuit constants indicates that the limiting speed is determined primarily by the time constant of the output circuit. With a sinusoidal input voltage this type of clipper circuit should operate reasonably well at frequencies up to 5 Mc/s.

621.396.611.1 **942**

Response of a Circuit to a Linear-Frequency-Sweep Voltage.—J. Marique. (*Onde élect.*, July 1951, Vol. 31, No. 292, pp. 313–315.) The universal response curves of Hok (671 of 1949) are transformed into a set of curves which can be applied more readily in practice. These are based on circuit bandwidth corresponding to a 3-db fall in response, and show response as a function of sweep velocity. As this increases the current maximum decreases and occurs later, while the pass band (3 db below the maximum) increases.

621.396.611.1 : 681.142 **943**

The Study of Oscillator Circuits by Analog Computer Methods.—Han Chang, R. C. Lathrop & V. C. Rideout. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 286–294.)

621.396.615.18 **944**

Modified Locked-Oscillator Frequency Dividers.—P. G. Sulzer. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1535–1537.) The circuit described consists simply of an LC oscillator with a resistor inserted between the oscillatory circuit and the valve. It is capable of operating over a 7-to-1 range of anode voltage when dividing by a factor of 30.

621.396.645 **945**

Cascading Cathode Followers to provide High Impedance-Transformation Ratios.—S. E. Smith & W. J. Kessler. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 129–135.) The input resistance and impedance-transformation ratio of a cathode-follower circuit can be increased considerably by connecting a second circuit in series so that the load resistor for the first stage is the grid resistor for the second stage. The effective load resistance of the first stage is then considerably greater than the d.c. circuit resistance. The derivation of the amplification equations is outlined and examples of the application of the circuit are described.

621.396.645 : 621.395.61 **946**

Equipment for Acoustic Measurements: Part 1 — A Portable General-Purpose Microphone Amplifier using Miniature Valves.—D. E. L. Shorter & D. G. Beadle. (*Electronic Engng*, Sept. 1951, Vol. 23, No. 283, pp. 326–331.) The amplifier is designed to have the highest possible signal/noise ratio with minimum distortion at the maximum output of 400 mW. Details of the circuits, the mechanical layout and the performance are given.

621.396.645 : 621.395.623.7 **947**

Cathode-Follower Loudspeaker Coupling.—E. W. Fletcher & S. F. Cooke. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 118–121.) Excellent square-wave response from 10 c/s to 40 kc/s is obtained from a balanced push-pull cathode-follower stage with four Type-6AS7G double triodes per side (all valves in parallel), with the low-frequency and high-frequency loudspeakers in a dividing network connected directly between the cathodes. The audio power to the speakers is approximately 6 W and the power gain is 37 db. With ten 6AS7G valves per side the audio power is increased to about 33 W.

621.396.645.018.424† **948**

Design of Wide-Band Amplifiers.—H. Behling. (*Frequenz*, Aug. & Sept. 1951, Vol. 5, Nos. 8 & 9, pp. 209–217 & 246–249.) Overall amplification V as a function of the number of stages *n* is compared for three types of amplifier. The relation is expressed in terms of a parameter *A* dependent on the slope and capacitance of the valves used and on the required total bandwidth. For a prescribed amplification and bandwidth, the stagger-tuned circuit requires the least number of stages. The V/*n* characteristics for the single-tuned amplifier are the least favourable since they show a pronounced maximum value for V. The double-tuned amplifier shows a similar tendency to a maximum value of V, depending on the value of *n*, but the attainable amplification is much greater than for the corresponding single-tuned amplifier and this type affords the best compromise between difficulties of practical adjustment and optimum design. Negative-feedback and stagger-damped circuits are also considered and a practical example is discussed.

621.396.645.029.3 **949**

A Symmetrical Audio Amplifier with Over-All Negative Feed-Back.—W. G. Whittleston. (*N.Z. J. Sci. Tech., B*, Sept. 1950, Vol. 32, No. 2, pp. 25–29.) Amplifiers are described incorporating phase-splitter circuits comprising cathode-coupled triode pairs, with signal input applied to the grid of one triode and negative-feedback voltage applied to the grid of the other. Stability is better than that of the Williamson amplifier (2715 of 1947).

621.396.645.029.3 : 621.395.61 **950**

Construction of A.F. Amplifiers for Studio Equipment.—H. Schiesser & H. Gathmann. (*Elektrotech. Z.*, 1st Sept. 1951, Vol. 72, No. 17, pp. 523–525.) A review of developments in the design of microphone amplifiers.

621.396.645.22 **951**

Medium-Frequency Amplifiers with Reduced Phase Distortion.—J. Laplume. (*Onde élect.*, Aug/Sept. 1951, Vol. 31, Nos. 293/294, pp. 357–362.) Analysis of tuned-transformer and stagger-tuned coupling circuits, showing phase- and amplitude-response diagrams. See also 681 and 1328 of 1949.

621.396.645.371 **952**

Note on Negative-Feedback Amplifiers.—U. Kirschner. (*Frequenz*, Aug. 1951, Vol. 5, No. 8, pp. 223–230.) The matrix method of representation of a valve circuit (2666 of 1951) is applied to determine the amplification, degree of stability, attenuation and phase distortion of a selective feedback amplifier. Such a circuit is represented as an 8-terminal network and methods of dealing with this are described.

621.396.645.372 : 621.385.5 **953**

Miniaturizing Pentode Amplifiers by Positive Feedback.—W. B. Anspacher. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 103–111.) Analysis of a 2-stage pentode amplifier circuit without bypass capacitors, the resulting degeneration being nullified by means of positive feedback between the two screen grids. An expression for the feedback resistance is developed in terms of the amplification factors obtained on removing in succession the screen-grid and cathode bypass capacitors.

621.316.8 **954**

Bauelemente der Nachrichtentechnik. Teil 2: Widerstände. [Book Review]—H. Nottebrock. Publishers: Schiele & Schön, Berlin, 1949, 216 pp., 7.50 DM. (*Arch. elekt. Übertragung*, May 1950, Vol. 4, No. 5, p. 188.) Gives practical information on the various types of resistors for telecommunication applications, including nonlinear types.

621.396.645 + 621.396.62 **955**

Verstärker und Empfänger (Amplifiers and Receivers). [Book Review]—M. J. O. Strutt. Publishers: Springer, Berlin, 2nd revised edn 1951, 422 pp., 53.55 Swiss francs. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, 1st Sept. 1951, Vol. 29, No. 9, p. 359.) A textbook covering all types of amplifiers and receivers, with thorough treatment of fundamental aspects.

GENERAL PHYSICS

537.525 **956**

High-Frequency Gas-Discharge Breakdown.—S. C. Brown. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1493–1501.) "High-frequency discharge breakdown is controlled by the process of electron diffusion and, besides the theory of its behaviour, the physical limitations of tube size, gas pressure, and frequency for this type of breakdown are given. The particular case of hydrogen is cited. The effects of superimposing a small d.c. field and a magnetic field on the a.c. field are also discussed."

537.525 **957**

The Mechanism of Positive-Ion Collection by a Spherical Probe in a Dense Gas.—R. L. F. Boyd. (*Proc. phys.*

Soc., 1st Sept. 1951, Vol. 64, No. 381B, pp. 795–804.) A theoretical treatment of the use of an electronegative spherical probe with dimensions greater than the ionic and electronic mean free paths shows this to be a possible means of determining ionic density in a dense gas discharge, where Langmuir's probe technique is inapplicable. The method depends on a knowledge of the radius of the space-charge sheath surrounding the probe. It is suggested that this is difficult to determine experimentally.

537.525 958
Form of Transient Currents in Townsend Discharges with Metastables.—J. P. Molnar. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 933–940.) The form of the current is calculated for a Townsend discharge stimulated by a pulsed light beam, with particular reference to the current component initiated by metastable effects.

537.525.5 : 621.385.1 959
Studies of Externally Heated Hot-Cathode Arcs: Part 1—Modes of the Discharge.—L. Malter, E. O. Johnson & W. M. Webster. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part I, pp. 415–435.) A study of the modes of discharge which take place as current increases in a gas-discharge cylindrical diode with inner hot cathode. The anode-glow, ball-of-fire, Langmuir and temperature-limited modes are considered and data are presented for plasma and floating potentials, electron temperature, plasma density and glow distribution for these forms of discharge.

537.533 : 537.525 960
Studies of γ-Processes of Electron Emission employing Pulsed Townsend Discharges on a Millisecond Time Scale.—J. P. Molnar. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 940–952.) An experimental investigation of the relative amounts of cathode electron emission due to ions, photons, and metastable atoms in a pulsed Townsend discharge in argon.

537.533.71 961
On the Reflection of Electrons by Metallic Crystals.—L. A. MacColl. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 888–906.) The reflection coefficient is calculated for electrons incident normally on a plane face of a metallic crystal, assuming that the potential energy of an electron is a sine function of distance inside the crystal and obeys the classical image-force law outside the crystal.

537.568 962
Concerning the Mechanism of Electron-Ion Recombination: Part 2.—M. A. Biondi. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 1078–1080.) Experiments on samples of He and Ne containing 0·1% Ar indicate that the recombination involves an electron and a molecular ion and that the mechanism is probably dissociative recombination. Part 1: 658 of March (Biondi & Holstein).

538.11 : 621.318.2 963
The Air-Gap Field of Permanent Magnets.—L. Kneissler. (*Elektrotech. u. Maschinenb.*, 1st Sept. 1951, Vol. 68, No. 17, pp. 393–398.) Theoretical investigation of the fundamental consequences following from the assumption that the field of a permanent magnet results from a system of elementary currents.

538.311 + 621.317.441 964
Axially Symmetric Systems for Generating and Measuring Magnetic Fields: Part 1.—M. W. Garrett. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1091–1107.) A systematic discussion of (*a*) axially symmetric magnetic fields, both central and remote from the origin, (*b*) search coils reporting the field and gradient at a single point, and (*c*) mutual inductors. Universal error-contour maps are derived for the central field or gradient in systems having errors of second, fourth or sixth order and for hybrid types. Source systems include circular filaments, cylindrical or plane circular current sheets, and thick solenoids of rectangular or notched section. Source constants derived for the particular source type are combined into a set of overall coefficients that express the field constants for a complete system. Rapid methods are given for computing the source constants and from them all the field derivatives. Tabular aids and reference formulae are given, and the rates of convergence of series for central and remote fields, and for mutual inductors, are discussed. Special systems are briefly described.

538.312 : 538.56 965
Theory of Multipole Radiations.—P. R. Wallace. (*Canad. J. Phys.*, Sept. 1951, Vol. 29, No. 5, pp. 393–402.) Analysis free from approximations is given for the radiation from a given oscillating system of charges and currents. The energy flux and density and the angular-momentum density are expressed in term of 'rotation operators' the expansion of which into eigenfunctions corresponds to separation into electric and magnetic multipoles of all orders.

538.56 966
Application of the Radiation from Fast Electron Beams.—H. Motz. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, p. 1219.) Correction to paper abstracted in 2411 of 1951.

538.566 967
Total Reflection of an Electromagnetic Wave with Metallic Reflection of the Evanescent Wave.—S. Gibellato. (*Nuovo Cim.*, 17th Sept. 1949, Vol. 6, No. 5, pp. 344–359.) Experiments using waves of length 3·43 cm are described in which a plane metal surface is arranged parallel to the surface of a paraffin prism at which total reflection occurs. The evanescent refracted wave is thus reflected and influences the polarization of the totally reflected wave. The variation of polarization with distance between the two reflecting surfaces is plotted.

538.566 : 535.42 968
Rigorous Theory of the Diffraction of Plane Electromagnetic Waves at a Perfectly Conducting Circular Disk and at a Circular Aperture in a Perfectly Conducting Plane Screen.—W. Andrejewski. (*Naturwissenschaften*, Sept. 1951, Vol. 38, No. 17, pp. 406–407.) Curves are given which have been calculated from the formulae previously derived by Meixner and the author (see 2767 of 1950). The values obtained by the Levine-Schwinger approximation method are included for comparison. The agreement is only satisfactory for values of $ka < 2$, where a is the radius of the disk and k the wave number.

538.566 : 535.42 969
On the Half-Plane Diffraction Problem.—F. G. Friedlander. (*Quart. J. Mech. appl. Math.*, Sept. 1951, Vol. 4, Part 3, pp. 344–357.) The diffraction of an arbitrary disturbance is considered. The solution, for all points of space, can be obtained by a modification of the method developed by Hadamard for Cauchy's problem. For the diffraction represented by Hadamard's 'elementary solution' a simple expression is found whose Laplace transform agrees with the Green's functions derived from Sommerfeld's two-valued solutions of the wave equation.

538.566 : 535.42 970
Study of Diffraction by Plane Screens and Application to Lenses for Hertzian Waves.—J. C. Simon. (*Ann. Radioélect.*, July 1951, Vol. 6, No. 25, pp. 205–243.) The

general problem of diffraction by a perfectly conducting plane screen with any form of aperture is solved by a perturbation method. Diffraction at a single aperture is studied by considering an auxiliary source in the aperture; approximate solutions are obtained for a circular hole (*a*) small compared with the wavelength, (*b*) of the order of the wavelength. The case of a plane wave incident normally on a screen with equal holes uniformly spaced is related to that of propagation in a rectangular waveguide with two walls of zero impedance and two of infinite impedance. Phase variations in the case of oblique incidence are related to ordinary-wave and extraordinary-wave propagation in crystal optics. An equivalent refractive index < 1 can be obtained by using several parallel screens. Measurements indicate that accurate phase correction may be made as in the Schmidt optical system. Application of these principles in the design of an aerial for 3·6 kMc/s is described.

538.566 : 537.525.6 **971**

Waves and Plasma.—W. O. Schumann. (*Elektron Wiss. Tech.*, Sept. 1951, Vol. 5, No. 9, pp. 279–288.) The mode of oscillation and characteristics of waves propagated in plasma layers are discussed. Different combinations of e.m. waves and stationary or moving plasma are considered in turn. 39 references.

539.153 : 546.28 **972**

Contribution to the Theory of Impurity Centers in Silicon.—W. Baltensperger. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 1055–1056.) Detailed analysis of the energy of an electron bound to a phosphorus impurity atom in silicon, with components due to the field of the ion, the dipole field of the polarized atoms, and the periodic field of the crystal. The result is in agreement with experiment and with the simple theory assuming a hydrogen-like atomic binding.

541.183.5 : 621.385.833 **973**

The Use of the Field Emission Electron Microscope in Adsorption Studies of W on W and Ba on W.—J. A. Becker. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 907–932.)

548.24 : 546.431.824-31 **974**

Detwinning Ferroelectric Crystals.—E. A. Wood. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 945–955.) "Unstrained single crystals of $BaTiO_3$ can be detwinned under the influence of an electric field at elevated temperature, but strained crystals cannot. It seems probable that this is also true of crystals in a polycrystalline body such as a ceramic."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 621.396.9 **975**

Variation of Meteor-Echo Rates with Radar-System Parameters.—McKinley. (See 995.)

523.72 **976**

Photon Counter Measurements of Solar X-Rays and Extreme Ultraviolet Light.—H. Friedman, S. W. Lichtman & E. T. Byram. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 1025–1030.) Results obtained in a V-2 rocket experiment indicate that solar soft X rays are important factors in E-layer ionization, that Hα radiation penetrates the atmosphere to well below the E layer, and that O_2 is rapidly changed to O at heights above 100 km.

523.72 : 621.396.822 **977**

Absence of Hydrogen Radiation of Wavelength 21 cm in the Sun.—C. de Jager, M. Minnaert & C. A. Muller. (*Nature, Lond.*, 1st Sept. 1951, Vol. 168, No. 4270, p. 391.) Using the same equipment as for investigations of the galaxy [982 below (Muller & Oort)] no indication of the hydrogen 1 420 Mc/s line could be detected in the solar r.f. radiation. An explanation of this is given.

523.85 : 621.396.822 **978**

An Accurate Determination of the Positions of Four Radio Stars.—F. G. Smith. (*Nature, Lond.*, 29th Sept. 1951, Vol. 168, No. 4274, pp. 555–556.) The results are given of recent determinations of the positions of four intense radio stars, using an interferometer previously described [121 of January (Ryle et al.)] and also a new interferometer at wavelengths of 3·7 and 1·4 metres. Declination was determined from the periodicity of the record, and by two new methods involving star transit times.

523.854 : 621.396.822 **979**

A Search for Long-Period Variations in the Intensity of Radio Stars.—M. Ryle & B. Elsmore. (*Nature, Lond.*, 29th Sept. 1951, Vol. 168, No. 4274, p. 555.) Interferometer observations were made on a wavelength of 3·7 metres on nearly every day for 18 months. It is concluded that none of the observed radio stars (approximately 100) varies by more than 0·1 magnitude with any period shorter than 1 000 days. See also 121 of January (Ryle et al.).

523.854 : 621.396.822 **980**

Galactic Radiation at 18·3 Mc/s.—C. A. Shain. (*Aust. J. sci. Res., Ser. A*, Sept. 1951, Vol. 4, No. 3, pp. 258–267.) The observations are expressed as equivalent-temperature contours and cover a zone centred on declination − 34°. The contours are similar in shape but higher in value than those found at 100 Mc/s. The ratio of maximum to minimum temperature is less than at 100 Mc/s even when the difference in aerial directivity is taken into account. Ionospheric absorption is higher and lower than expected near the F_1- and E-layer critical frequencies respectively. The collision frequency at the height of maximum ionization in the E layer is $< 10^4$/sec.

523.854 : 621.396.822 **981**

Radio from Galactic Hydrogen at 1 420 Mc/s.—H. I. Ewen & E. M. Purcell. (*Nature, Lond.*, 1st Sept. 1951, Vol. 168, No. 4270, p. 356.) Using a microwave radiometer which measured the apparent radio-temperature difference between two spectral bands 17 kc/s wide and 75 kc/s apart, the 1 420 Mc/s line was first detected in March 1951. The results of subsequent observations are briefly discussed.

523.854 : 621.396.822 **982**

The Interstellar Hydrogen Line at 1 420 Mc/s and an Estimate of Galactic Rotation.—C. A. Muller & J. H. Oort: J. L. Pawsey. (*Nature, Lond.*, 1st Sept. 1951, Vol. 168, No. 4270, pp. 357–358.) Using equipment which measured the difference between the radiation received over two frequency bands 25 kc/s wide and 110 kc/s apart, curves showing the intensity of galactic radiation as a function of right-ascension have been obtained. From the observed characteristics of the 1 420 Mc/s line, galactic rotational velocities are deduced which agree with values calculated from a schematic model of the galactic system.

551.510.535 **983**

The Negative-Ion Concentration in the Lower Ionosphere.—D. R. Bates & H. S. W. Massey. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 1–13.) The processes leading to the formation and destruction of negative ions in the upper atmosphere are considered in detail. The day-time equilibrium concentration of negative ions in the D and E regions does not seem to be sufficiently high to be readily reconcilable with the dynamo theory of the L (lunar) and S (solar) magnetic variations. Thus the calculated transverse conductivity of the E region is found to be such that the magnitude of the L current can apparently only be explained if the local tidal motion is 6 000 times greater than at ground level, but this is not supported by radio observations. The transverse conductivity of the D region also appears to be less than that demanded by present theory. The fade-out enhancement of the electron concentration in the lower ionosphere is briefly discussed; the possibility that it arises from photo-detachment by Lyman (α) radiation from solar flares cannot be excluded.

551.510.535 **984**

Comparison of the Results of Measurements of Ionospheric Absorption made at Two European Stations.—K. Rawer. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 38–50. In French.) Absorption measurements at Slough show that (*a*) the seasonal variation of the noon absorption follows a cos $\chi^{\frac{3}{4}}$ law (χ is the sun's zenith angle), (*b*) the influence of the solar cycle is important, (*c*) day-to-day fluctuations are considerable. Daily values measured at Slough and at Freiburg (distance 400 km) mostly show a positive correlation, with coefficient about 0·4. A new method of correlation is proposed.

551.510.535 **985**

The Contributions of the D and E Regions in Measurements of Ionospheric Absorption.—K. Bibl & K. Rawer. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 51–65. In French.) Observations of ionospheric absorption made on different frequencies reveal an important contribution of selective absorption occurring in the E layer, which can be calculated by assuming a parabolic variation of the electron density of the E layer and an exponential variation of the collision number. Using the results of this calculation, a new evaluation of the measurements at Slough and Freiburg has been made for a period of 24 months. A graphical method is described which gives the separate contributions of the E and D layers. The values of D-layer absorption thus obtained are lower than those given by the usual method, where the influence of the E layer is neglected. In spite of a considerable dispersion, this evaluation leads to the adoption of a fairly constant value for the influence of the E layer. It is proposed to use this value in future for reduction of absorption measurements; the corresponding E-layer collision number is certainly $< 10^4$/sec, a value in good agreement with recent temperature and pressure data at 125-km height. See also 984 above (Rawer).

551.510.535 **986**

Absorption of Solar Energy by Oxygen Molecules in the E Layer.—R. Penndorf. (*J. Met.*, June 1950, Vol. 7, No. 3, pp. 243–244.) Results of calculations of energy absorption for heights ranging from 90 to 125 km are plotted; the curve has a maximum at about 102 km and resembles closely that previously given for the number of quanta absorbed (2224 of 1949).

551.510.535 **987**

Determination of the Number of Collisions in the Ionosphere E and F Regions.—K. Rawer, K. Bibl & É. Argence. (*C. R. Acad. Sci., Paris*, 17th Sept. 1951, Vol. 233, No. 12, pp. 667–669.) The collision frequencies were deduced by the classical method from the observed amplitudes of multiple reflections. For the F layer measurements were made at four frequencies, during the night, to eliminate the influence of the D layer, from November 1949 to April 1950 and during February 1951. For the E layer measurements were made at five frequencies, at midday, over a period of two years. Values of 2×10^2 per sec and $6 \cdot 7 \times 10^3$ per sec respectively were found for the collision frequencies at the middle layers of the F and E regions. The double focusing effect due to the variation of curvature of the F layer is taken into account.

551.510.535 **988**

The Temperature of the Upper Atmosphere.—D. R. Bates. (*Proc. phys. Soc.*, 1st Sept. 1951, Vol. 64, No. 381B, pp. 805–821.) Thermal equilibrium in the F layers is studied. The most important energy-loss processes are conduction in the gas and emission by the magnetic dipole connecting the two levels of the ground term of atomic oxygen. Total heat loss is greater than is consistent with the energy gain attributed to ionizing photons. It is possible that the estimate of this energy gain, based on radio measurements, may be much too small. Such measurements give only a lower limit to the electron production rate. The suggestion is tentatively made that heat is supplied to the upper atmosphere mainly by non-observed ionization.

551.510.535 **989**

Ionospheric Behaviour in the F_2 Region at Singapore.—B. W. Osborne. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 66–78.) Observations made at Singapore from November 1948 have shown the interdependence of F_2-layer maximum ionization density and height with the seasonal occurrence of thick-layer effects. Anomalous behaviour was frequent during the morning hours of the December solstice. Layer height changes after sunset and the associated occurrence of the equatorial scatter phenomenon are discussed. Consideration is given to the interpretation of vertical-incidence virtual-height measurements under Singapore conditions of a thick F region.

551.510.535 : 551.594.6 **990**

Calculation of the Height of the Lower Layers of the Ionosphere from the Waveform of Atmospherics.—F. Schindelhauer, A. Schrader & C. Horinger. (*Z. Met.*, Sept. 1951, Vol. 5, No. 9, pp. 277–284.) Ray-geometry methods of calculation introduced by Laby et al. (2184 of 1940) and developed by Schonland et al. (35 of 1941) are outlined. Oscillographic records of atmospherics obtained between September 1948 and September 1950 at Potsdam indicate reflection at heights of 35, 65 and 80–100 km, corresponding respectively to the C, D and lower E layers. Graphs are presented showing the variation of these heights with time and with solar activity. Sources of disturbance other than lightning discharges are considered responsible for some of the atmospherics. The C layer may correspond to the upper boundary of the ozone layer.

551.594.1 **991**

Diurnal Variations of Atmospheric Electricity and Air-Mass Exchange in the High Alps. The Atmospheric-Electrical Conditions on the Jungfraujoch (3472 m).—H. Israël, H. W. Kasemir & K. Wienert. (*Arch. Met. Geoph. Bioklimatol. A*, 10th Aug. 1951, Vol. 3, No. 5, pp. 357–381.) Observations made during periods of several weeks at different seasons are reported and discussed. In summer, the conditions are of continental type, with potential gradient varying in the opposite sense to conductivity and vertical current. In autumn the conditions tend to oceanic type, with conductivity varying in the opposite sense to potential gradient and vertical current. These results are explained on the basis of the movements of air masses between upper and lower levels.

551.594.22 **992**

The Work of the Bernard Price Institute of Geophysical Research, 1938–1951.—B. F. J. Schonland. (*Trans. S.*

Afr. Inst. elect. Engrs, Aug. 1951, Vol. 42, Part 8, pp. 241–254.) Some account of the Institute's research on lightning, and of war-time radar developments, is included.

LOCATION AND AIDS TO NAVIGATION

621.39.001.11 : 621.396.9 **993**

Information Theory and the Design of Radar Receivers.—P. M. Woodward. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1521–1524.) "Deals with the problem, frequently encountered in radar, of extracting simple numerical information from a noisy waveform. It is suggested that the only ideal way of doing this is to use the principle of inverse probability and convert the waveform into a probability distribution for the quantity sought. The method is applied to the problem of determining the time delay of a periodically modulated r.f. waveform in the presence of white Gaussian noise when the undelayed waveform without noise is exactly known. As a result, the matched predetection filter of Van Vleck & Middleton (1919 of 1947) is automatically specified, and the theory of ideal detection is briefly indicated."

621.396.9 **994**

Terrestrial Radar.—A. Flambard. (*Onde élect.*, June & July 1951, Vol. 31, Nos. 291 & 292, pp. 261–270 & 320–328.) Review of development, with descriptions of European and American equipment.

621.396.9 : 523.5 **995**

Variation of Meteor-Echo Rates with Radar-System Parameters.—D. W. R. McKinley. (*Canad. J. Phys.*, Sept. 1951, Vol. 29, No. 5, pp. 403–426.) Observations made with crossed-polarization radar systems do not support the suggestion that an ionized meteor trail may act as a strong filter-polarizer of the incident radio wave. Experiments to determine the variation of normal meteor echo rates with transmitter power, aerial gain, and radio wavelength, all confirm Lovell's scattering formula, provided account is taken of the effective broadening of the scattering pattern of the meteor trail with increased wavelength.

621.396.9 : 526.9 **996**

Recent Lorac Developments.—J. E. Hawkins. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 218–226.) Description of a new method of radio surveying which has been tested experimentally in the Gulf of Mexico area, using frequencies near 1 772 and 1 798 kc/s. Intersecting hyperbolic interference patterns are produced by the radiations from the two pairs of transmitters; 'fixes' are obtained directly from the readings of two phase meters in the mobile receiver.

621.396.93.089.6 **997**

The Calibration of Aircraft Direction-Finders with Particular Reference to Site Selection.—J. H. Moon. In 143 of January, for 'Vol. 15' please read 'Vol. 14'.

621.396.932/.933].2 + 621.317.789 **998**

Direction Finder and Flow Meter for Centimeter Waves.—K. Morita. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1529–1534.) The direction of arrival of the wave is indicated by the minimum response of a dipole located at the focus, and along the axis, of a small paraboloidal reflector. The performance of an experimental model for 10-cm waves is described.

The flow meter indicates directly the active power flow of the wave in magnitude and direction, even when there are standing waves. An energy flow of 1 μW/cm² at a wavelength of 10 cm has been measured.

621.396.932/.933].2 **999**

Selecting Critical Components for Matched-Channel Radio Receiving Systems.—H. D. Webb. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 206–217.) In some d.f. systems receivers are necessary with two or more channels matched in phase and gain characteristics. Matching may be made less critical by using wide-band i.f. amplifiers having selective filters, or wide-band r.f. amplifiers with suitably selected bandwidths, or a combination of the two. Selective grading and matching is proposed as an economical means of providing components with the close tolerances requisite for such receivers.

621.396.933 **1000**

The Civil Aeronautics Administration V.H.F. Omnirange.—H. C. Hurley, S. R. Anderson & H. F. Keary. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1506–1520.) Detailed description of a system operating in the frequency band 112–118 Mc/s and producing two signals, one providing a reference phase, while the phase of the other varies directly with the magnetic bearing of an aircraft from the ground station. A phase comparator in the aircraft enables the pilot to determine his magnetic bearing with respect to the station, and to select and fly on a course on any desired bearing. The accuracy of the system, which has been in continuous operation for more than three years, is within about 1·5°.

621.396.933 **1001**

Rotating Radio Beacon with Angle-Dependent Frequency for Position Finding.—H. H. Rust. (*Arch. elekt. Übertragung*, Sept. 1951, Vol. 5, No. 9, pp. 421–424.) A position-finding system designed for maximum ease of operation on board ship or aircraft uses a rotating beacon in which either carrier or modulation frequency is a function of angle, so that every direction of radiation is associated with a definite frequency. With two such beacons at a suitable separation, position can be determined without ambiguity.

621.396.933 : 621.317.755 **1002**

A Video Oscillograph for Testing Distance Measuring Equipment.—I. A. Hood. (*A.W.A. tech. Rev.*, Sept. 1951, Vol. 9, No. 2, pp. 73–97.) Account of the design, construction and performance of apparatus developed specially for testing the equipment described by Lindsay, Blom & Gilchrist (1659 of 1951).

621.396.933.23 **1003**

Flight-Path Control.—D. L. Markusen. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 227–237.) Analysis of the dynamic-stability problems concerned in automatically guiding an aircraft along a path in space as defined by such radio beams as ILS (Instrument Landing System) or omni-range. A control system in which the bank angle is proportional to the off-course error is shown to be inherently unstable, but may be stabilized over a limited range of gain by the addition of a simple phase-adjusting network or by use of a heading reference. The design of such a stabilized system is a compromise between gain, damping, and response to noise; the analytical work has been confirmed by flight tests.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 **1004**

Calibration of Ionization Gauges for Various Gases at Low Pressures.—S. Wagener & C. B. Johnson. (*J. sci. Instrum.*, Sept. 1951, Vol. 28, No. 9, p. 278.)

535.215 **1005**

An Outline of Some Photoconductive Processes.—A. Rose. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part I, pp. 362–414.) Analysis of photoconductive processes, with the introduction of the concept of a steady-state Fermi limit, to facilitate comparison with experimental results. Explanations are found for *(a)* current-versus-light curves with exponents between 0·5 and 1·0, *(b)* the lack of reciprocity between sensitivity and speed of response, *(c)* the low values of currents excited in thin evaporated insulating films by light or electron bombardment. The ratio of life time of a free carrier to the observed time-constant is proposed as a figure of merit for a photoconductor.

535.215 : 546.23 **1006**

Photoconductivity in Amorphous Selenium.—P. K. Weimer & A. D. Cope. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part 1, pp. 314–334.) Sustained photocurrents are obtained from films of amorphous Se which are highly insulating in the dark. Hole conduction is predominant over electron conduction, the range of the holes in the Se exceeding 10^{-3} cm. The spectral response is a maximum at 4 000–4 500 Å and extends into the far ultraviolet. The maximum response does not coincide with the optical absorption edge, as is usually the case: the absorption edge occurs at much longer wavelengths (6 000 Å). The time constant for the rise and decay of photocurrent is $<50\ \mu$s. Space-charge-limited currents are observed with excess light for low fields across the material. Evidence for primary and secondary photocurrents is discussed. The application of the television scanning method to photoconductive measurements is described and its advantages and limitations are discussed.

535.215 : [546.482.21 + 546.482.31 **1007**

Some Aspects of the Photoconductivity of Cadmium Sulfide.—R. W. Smith. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part 1, pp. 350–361.) Measurements of photocurrent, time constant, Hall effect and potential distribution for CdS and CdSe crystals are described, and are discussed with reference to quantum yield.

535.215 : 546.863.221 : 621.397.611.2 **1008**

Properties of some Photoconductors, principally Antimony Trisulfide.—S. V. Forgue, R. R. Goodrich & A. D. Cope. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part 1, pp. 335–349.) A study of those properties of red Sb_2S_3 which make it promising as a target material for television pickup tubes. Plots of the variation of dark current with voltage and of photocurrent with illumination are given, and spectral response, sensitivity and quantum efficiency are discussed. The effects of impurity, heat treatment and method of preparation are considered.

537.311.3 + 538.632] : 546.723-31 **1009**

Electrical Properties of αFe_2O_3 and αFe_2O_3 containing Titanium.—F. J. Morin. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, pp. 1005–1010.) Electrical conductivity, Hall effect and Seebeck effect were measured for polycrystalline samples of αFe_2O_3 and of αFe_2O_3 containing 0·05–1·0% Ti (*n*-type impurity). In one set of samples there was a 0·6% excess of Fe (*n*-type impurity) and a 0·6% deficiency of Fe (*p*-type impurity) in the second set. Carrier-concentration results indicate that each added Ti ion contributes approximately one electron to the conduction process. Mobilities are found to be $<2{\cdot}0$ cm/s per V/cm, suggesting that conduction involves electrons in the *d* level of Fe.

537.311.33 **1010**

Some Results concerning the Partial Differential Equations describing the Flow of Holes and Electrons in Semiconductors.—R. C. Prim, III. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1174–1213.) The subject equations are investigated with the aim of establishing some general properties of the flow fields. The results include a number of geometrical characteristics of the vector fields, a suggested reformulation of the partial differential equations restricting carrier concentration and electrostatic potential, and a family of solutions in closed form for the steady-state no-recombination case.

537.311.33 : 546.289 **1011**

Hot Electrons in Germanium and Ohm's Law.—W. Shockley. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 990–1034.) Ryder's data (1398 of 1951) on the mobility of electrons in electric fields up to 4×10^4 V/cm are analysed. The mobility decreases because of the influence of scattering by optical modes and because of increases in electron energy. Electron temperatures estimated at 4 000°K have been produced in specimens having atomic-vibration temperatures of 300°K. The critical drift velocity above which there are deviations from Ohm's law is about $2{\cdot}6 \times 10^6$ cm/s.

538.652 : 669.157.82 **1012**

The Magnetostriction of Single Crystals of Iron-Silicon Alloys.—W. J. Carr & R. Smoluchowski. (*Phys. Rev.*, 15th Sept. 1951, Vol. 83, No. 6, pp. 1236–1243.)

546.23 : [535.323 + 535.343 **1013**

Optical Properties of Selenium.—J. J. Dowd. (*Proc. phys. Soc.*, 1st Sept. 1951, Vol. 64, No. 381 B, pp. 783–789.) The refractive index of amorphous Se and the absorption coefficient of amorphous and single-crystal Se were measured over a wide frequency range. The edge of the absorption band for amorphous Se is at about 0·60 μ, corresponding to a band spacing of 2·05 eV. A discrepancy between the value obtained for the infrared refractive index and that deduced from the dielectric constant of amorphous Se indicates a further absorption band in the infrared.

546.28 : 548.55 **1014**

Measurement of the Elastic Constants of Silicon Single Crystals and their Thermal Coefficients.—H. J. McSkimin, W. L. Bond, E. Buehler & G. K. Teal. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, p. 1080.) Measurements of velocities of propagation for both shear and longitudinal waves were made at frequencies in the range 8–12 Mc/s, and the three independent elastic constants and their temperature coefficients were evaluated. The results are tabulated.

546.74 : 539.32] : 534.321.9.012 **1015**

Frequency Dependence of Elastic Constants and Losses in Nickel.—Bozorth, Mason & McSkimin. (See 887.)

621.315.61 **1016**

Electro-Ceramics, with Special Reference to Pyrophyllite.—N. E. Hyde. (*Electronic Engng*, Sept. 1951, Vol. 23, No. 283, pp. 336–340.) A brief review of the origin and history of ceramics, and of the processes involved in their manufacture, together with an account of the more important characteristics of pyrophyllite ceramics recently developed. Mechanical impact strength is higher than that of any other electrical porcelain, volume resistivity is high at elevated temperatures, shrinkage after firing is very small, and the ease of machining the unfired material makes it particularly suitable for the rapid construction of a wide range of shapes.

621.315.61 : 539.23 **1017**

The Electric Tunnel Effect across Thin Insulator Films in Contacts.—R. Holm. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, p. 1217.) Correction to paper abstracted in 2455 of 1951.

621.315.612.8 : 546.817.831.4 **1018**

Antiferroelectric Structure of Lead Zirconate.—E. Sawaguchi, H. Maniwa & S. Hoshino. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, p. 1078.) Oscillation and powder photographs, and polarization-microscope examination of a small untwinned crystal, indicate an orthorhombic structure and that the material is antiferroelectric.

621.315.612.8 : 546.817.831.4 : 537.228.1 **1019**

Piezoelectric Effect in Lead Zirconate.—S. Roberts. (*Phys. Rev.*, 1st Sept. 1951, Vol. 83, No. 5, p. 1078.) Piezoelectric effects in $PbZrO_3$ ceramic disks were found to be only just detectable. This is explained by the structure of the material discussed by Sawaguchi, Maniwa and Hoshino (1018 above).

621.315.614.617 : 53.093/.096 **1020**

Deterioration of Organic Polymers.—B. S. Biggs. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1078–1102.) A general review of deterioration processes in polymers. Changes with time of the properties of such materials as rubbers, plastics, textiles, and varnishes are usually the result of chemical reaction with components of the atmosphere. The mechanisms of these reactions and some methods of preventing or retarding them are discussed.

669.3 : 621.314.632.1 **1021**

Carbon, Oxygen, and Sulfur Content of Chilean Coppers as Related to Cuprous-Oxide Rectifiers.—C. C. Hein & W. M. Hickam. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1192–1195.)

MATHEMATICS

681.142 **1022**

A Quarter-Square Multiplier using a Segmented Parabolic Characteristic.—B. Chance, F. C. Williams, Chia-Chih Yang, J. Busser & J. Higgins. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 683–688.) Description of equipment with a delay time <40 μs and accuracy within $\pm 1\%$.

681.142 **1023**

An Analog Computer for Indeterminate Mechanical Structures.—J. P. Corbett & J. F. Calvert. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 315–332.)

681.142 **1024**

A Versatile Small-Scale Analog Computer.—J. T. Carleton. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 308–314.) Description of a small computer which can be used either by itself or in conjunction with the larger Anacom equipment (2547 of 1949).

681.142 : 512.3 **1025**

A Circuit for Generating Polynomials and Finding their Zeros.—F. W. Bubb, Jr. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1556–1561.)

517.564.3(083.5) **1026**

Tables of the Bessel Functions of the First Kind of Orders Seventy-Nine through One Hundred and Thirty-Five. [Book Notice]—Staff of Computation Laboratory of Harvard University. Publishers: Harvard University Press, 1951, $8.00. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, p. 1579.)

MEASUREMENTS AND TEST GEAR

621.3.018.4(083.74) : 621.396 **1027**

[British] **Standard Frequencies.**—(*Wireless World*, Dec. 1951, Vol. 57, No. 12, p. 501.) A revised schedule of transmissions from Rugby is announced, as follows: 60 kc/s, 1029–1130 and 1429–1530 GMT; 5 Mc/s, 0544–0615 GMT; 10 Mc/s, 0629–0700 GMT.

621.3.087.45 : [551.594.11 + 551.594.13 **1028**

An Apparatus for Simultaneous Registration of Potential Gradient and Air-Earth Current (Description and First Results).—H. W. Kasemir. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 32–37.) Description of equipment in which the capacitors and resistors are chosen so that minor variations of gradient and current are suppressed, in order to give a clear display of the diurnal variations.

621.317 : 621.396.82 **1029**

Technique for the Measurement of Interference Voltages.—J. Pfister. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, 1st Sept. 1951, Vol. 29, No. 9, pp. 321–328. In German.) Reference is made to standards formulated by the C.I.S.P.R. for the measurement of interference voltages. An evaluation is made of the output from the selective i.f. circuit of the measuring apparatus, comprising two cascaded critically coupled band-pass filters, in response to an input of given transient character, viz. a.c. pulse, a.c. step, d.c. pulse or d.c. step. The voltage obtained in the following detector stage is then determined. The results are compared with those obtained by the same calculation procedure for a filter with an ideal narrow pass band, and are discussed in relation to the method of calibrating the apparatus.

621.317 : 621.396.822.029.3 **1030**

The Generation and Measurement of Low-Frequency Random Noise.—R. R. Bennett & A. S. Fulton. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1187–1191.) Methods are discussed for determining the important characteristics of low-frequency noise, such as the mean value, spectral density, amplitude distribution and autocorrelation function. Particular attention is devoted to the length of time necessary to establish satisfactory estimates of the properties of low-frequency noise. A brief description is given of a noise generator with a uniform power spectrum from zero to 25 c/s.

621.317.335.2.029.5† **1031**

Measurement of Capacitances at High Frequency.—H. H. Emschermann & O. Zinke. (*Arch. tech. Messen*, Sept. 1951, No. 188, pp. T100–T101.) Principles of measurement, range and accuracy of various circuits are noted. In the simple current/voltage measurement, error can be kept below 1%. Arrangements of the double voltage divider for large and small capacitances, and bridge circuits for measurements down to 1 pF at 20 Mc/s and to 10^{-4} pF at 1 Mc/s, are shown. Application of the double-T bridge is described for impedances in which the resistive component is large.

621.317.335.3† : 621.365.55† **1032**

Measuring Dielectric Properties during H.F. Heating.—E. Mittelman. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, p. 79.) Summary only. Description of method and apparatus, with experimental results.

621.317.335.3.029.64† **1033**

Measuring the Dielectric Constant and the Loss Angle of Solids at 3000 Mc/s.—M. Gevers. (*Philips tech. Rev.*, Sept. 1951, Vol. 13, No. 3, pp. 61–70.) Dielectric constant and loss angle are determined from the detuning and reduction of quality factor of a cavity resonator (E_{010} mode) when the dielectric is introduced. Quality factors as high as 17 190 (98% of the theoretical value) have been obtained by careful surface finishing of the resonators. A reflex klystron is used as oscillator and a silicon crystal as detector. Measured values for various materials are tabulated and formulae and graphs are given which simplify the determination of the required quantities.

621.317.361 **1034**

Frequency Checking of Mobile Equipment.—M. H. Diehl & C. J. Statt. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 138 . . 184.) Outline of a method of making fine frequency adjustments. A standard input frequency of 10 kc/s triggers a multivibrator circuit producing 0·2-μs pulses with a recurrence frequency of 3333⅓/sec which are fed to a mixer together with the frequency under test. Channel spacing is such that for correct adjustment of the frequency under test, one mixer output frequency is 1⅔ kc/s. This is used in conjunction with a frequency of 1⅔ kc/s derived from the 10-kc/s standard input to produce a stationary ellipse pattern on a c.r.o. when the transmitter frequency is exactly adjusted.

621.317.382 **1035**

New Method for Measurement of Active Power of H.F. Generators.—F. Alf. (*Elektrotech. Z.*, 15th Sept. 1951, Vol. 72, No. 18, pp. 541–543.) The grid rectification inherent in the operation of the oscillator valve is used to obtain a d.c. proportional to the power output. Ordinary commercial wattmeters are used, and measurements are accurate to within 5–7%. Results are independent of frequency up to frequencies at which transit-time effects occur. Application is to generators for industrial heating.

621.317.441 + 538.311 **1036**

Axially Symmetric Systems for Generating and Measuring Magnetic Fields: Part 1.—Garrett. (See 964.)

621.317.7 **1037**

Self-Balancing Instruments and their Application to Nucleonic Measurement.—R. S. Medlock & W. A. Kealy. (*J. Brit. Instn Radio Engrs*, Sept. 1951, Vol. 11, No. 9, pp. 393–405.) 1951 Radio Convention (London) paper. General principles of operation of mechanical and electronic self-balancing recording instruments, and standard applications of these instruments to measurements associated with atomic-energy projects, are described. Circuits are illustrated which permit simple addition, subtraction, multiplication, division, derivation of roots and powers, differentiation and integration.

621.317.7 : 621.396.932/.933 **1038**

Remote Monitoring of Naval and Air Service Transmissions.—J. Marique. (*Onde élect.*, Aug./Sept. 1951, Vol. 31, Nos. 293/294, pp. 331–341.) Report of monitoring measurements carried out in Brussels by the C.C.R.M. (Centre de Contrôle des Radiocommunications des Services mobiles), including frequency checking, field-strength measurement, panoramic recording of frequency bands occupied, and spectrum analysis. Equipment is described and results are shown diagrammatically of a statistical analysis of frequencies in use, deviations from allocated frequency, and beacon field strengths.

621.317.71 **1039**

A Combined Current Indicator and Integrator.—W. A. Higinbotham & S. Rankowitz. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 688–690.) An instrument suitable for measuring the beam current in a particle accelerator and covering the range 0·01 to 500 μA full scale.

621.317.71.029.6 **1040**

Measurement of Current at High Frequency.—O. Zinke. (*Radio franç.*, Sept. 1951, No. 9, pp. 6–18.) Description of photoelectric, rectifier and thermocouple instruments. Design and application of the latter are particularly discussed; rating and characteristics of different elements are tabulated (*a*) for vacuum-type thermo-elements rated up to 200 mA, (*b*) for Miniwatt-S types rated up to 300 mA.

621.317.725 **1041**

Valve Voltmeter without Calibration Drift.—M. G. Scroggie. (*Wireless World*, Jan. 1952, Vol. 58, No. 1, pp. 14–18.) Description of adaptor, of infinite input resistance and zero output resistance, for use with any d.c. voltmeter.

621.317.733 **1042**

A Self-Tracking Bridge Detector for Audiofrequency Bridges.—J. L. Upham, Jr. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 659–664.) A tuned bridge detector with constant bandwidth of 5 c/s is described. It is automatically tuned to the output frequency of the associated signal generator. Its sensitivity and inherent noise are such as to permit detection of a 0·5-μV signal across a 1-MΩ input over the frequency range 100 c/s–20 kc/s. The detector may be used for maintaining bridge balance during slow variations of the measured impedance, the variations being recorded in terms of the movements of the balancing controls.

621.317.74 : 621.397.2 **1043**

Television Streaking Test Set.—R. K. Seigle. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 96–99.) Description of equipment for point-to-point testing of television transmission systems such as coaxial cables, r.f. links, etc. The apparatus requires blanking and synchronization inputs, and generates stepped waveforms at video frequency which, when fed into properly adjusted television receiving equipment, produce a sharp rectangular picture. The height, width and location of the rectangle can be varied. Maladjustment of the receiving equipment is manifested as a smearing of the rectangular picture.

621.317.755 **1044**

A Scanning Method for Simultaneous Display of Several Phenomena with a Single-Beam Oscillograph.—R. Classen, F. W. Gundlach & F. Lentze. (*Arch. tech. Messen*, Sept. 1951, No. 188, pp. T106.) Description of the equipment, with details of the pulse-generator circuits.

621.317.755 **1045**

A Cathode-Ray Oscillograph for Impulse Testing.—W. G. Fockler. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 391–399.)

621.317.755 : 535.88 **1046**

A Portable Projection Oscilloscope.—V. Wouk. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 380–390.) Description of equipment producing an oscillogram 14 in. × 11 in. which can be viewed in rooms with normal lighting.

621.317.755 : 621.317.6.029.3 **1047**

The Production Model of the Automatic A.F. Response Curve Tracer.—G. L. Hamburger. (*J. Brit. Instn Radio Engrs*, May 1951, Vol. 11, No. 5, pp. 165–201.) Description of the production instrument developed from the experimental model previously described (148 of 1949). Layout and circuit details are given and operation of circuits is analysed. The curves are displayed against a logarithmic-coordinate framework and can be either viewed for about one minute or photographed.

621.317.755.087.6 **1048**

A Six-Channel Cathode-Ray Recording Oscillograph.—W. D. Tilton, Jr. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 373–379.)

621.317.755.088 **1049**

Astigmatism Correction for Oscilloscopes.—H. O. Hoadley. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 706–708.) Astigmatism in oscilloscopes produces different degrees of sharpness for horizontal and vertical lines. It may be corrected by altering the average potential of one pair of deflector plates relative to the other. Circuits for balanced and unbalanced signal inputs are given. Some commercial oscilloscopes provide for astigmatism correction by varying the potential of the second anode, and this method is recommended for oscilloscopes using d.c. amplifiers.

621.317.757 **1050**

A High-Resolution Spectrum Analyzer.—T. Miller & D. Sims. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 513–517.) Description, with circuit details, of an analyser for use at frequencies of the order of 50 kc/s. A continuous automatic recording system gives records with a frequency range of 3 kc/s and resolution to 20 c/s. Typical records of sideband structure are shown for carriers with modulation frequencies of 250 c/s and 20 c/s respectively.

621.317.761 : 538.569.4.029.64/.65 **1051**

New Techniques in Microwave Spectroscopy.—W. E. Good. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 29–37.) A description is given of a sensitive microwave spectroscope, with particular reference to the low-noise input circuit and the characteristics of the Si-crystal detector. The absorption lines to be measured are displayed in frequency by application of an 85-kc/s alternating voltage to the absorbing gas. The resultant modulation is detected by the crystal, amplified by a high-gain 85-kc/s amplifier, and displayed on a c.r.o. Lines of frequency up to 40 kMc/s can be measured with an accuracy to within one part in 10^6. The frequency-measurement equipment, for which a detailed circuit diagram is given, uses a Si crystal for generation of a series of harmonic frequency markers spaced at intervals of 500 Mc/s, interpolation being effected by means of a calibrated receiver.

621.317.78 : 535.214 **1052**

The Measurement of Microwave Power by Radiation Pressure.—A. L. Cullen. (*Engineering, Lond.*, 21st Sept. 1951, Vol. 172, No. 4469, pp. 377–378.) The method used is analogous to that for the classical measurements of radiation pressure of light. Microwave radiation is directed on to a reflector surface within the vertical leg of a waveguide T system, this arrangement ensuring that reflected power is not returned to the source. It is emphasized that the method is absolute. See also 2771 of 1951.

621.317.789 + 621.396.932/.933].2 **1053**

Direction Finder and Flow Meter for Centimeter Waves.—Morita. (See 998.)

621.317.799 : 621.396.61 : 621.396.931 **1054**

Mobile-Transmitter Testing Set.—G. J. Kent. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 106–109.) Outline description of test equipment for p.m. or f.m. transmitters operating in the frequency bands 30–44 Mc/s and 152–175 Mc/s. With minor modifications, a.m. transmitters operating at any frequency from 540 kc/s to 110 Mc/s can also be tested. Measurements of r.f. power output, a.f. sensitivity, signal/noise ratio, and speech intelligibility, can be made in a few minutes.

621.396.615.14 : 621.396.621.001.4 **1055**

Microwave Generator with Crystal Control.—W. F. Marshall. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 92–95.) Based on 1950 National Electronics Conference paper (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 504–512). A portable 3 100-Mc/s signal generator for field work. The initial oscillation is crystal-controlled near 50 Mc/s within a band of width $1\frac{3}{8}$ Mc/s. After multiplication to 300 Mc/s the signal is passed to a Si crystal, where S-band harmonics are produced. Preselection is required to utilize a specific output harmonic. Substituting control crystals in two channels gives a frequency range of up to 600 Mc/s without changing other circuit components.

621.396.615.14.029.62/.63 **1056**

A 20 to 1 000 Mc/s Sweep Oscillator.—J. E. Ebert & H. A. Finke. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 499–503.) The full range is covered in a single continuous tuning adjustment. The tank circuit is a conventional $\lambda/4$ coaxial-line resonator at the higher frequencies; as its plunger moves back for the lower frequencies, the exposed section of the inner conductor gradually changes from a solid rod to a helix of increasing pitch. The frequency sweep is effected by rotation of a specially shaped capacitor plate close to the high-impedance end of the tank cavity.

621.397.61.001.41 **1057**

Measuring Television Transmitter Amplitude Characteristics.—Ruston. (See 1133.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

548.0 : 537.228.1].001.8 **1058**

Piezoelectric Crystals as Sensing Elements of Pressure, Temperature, and Humidity.—E. A. Roberts & P. Goldsmith. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, pp. 776–780.) Summary of 1951 A.I.E.E. Great Lakes District Meeting paper. Pressure may be measured either by its effect on the Q-value of a crystal or by the change in crystal frequency due to differential air loading. Temperature is measured by its effect on frequency, and humidity by the change of frequency due to the deposit of moisture on the crystal surface. Types and cuts of crystal for optimum results are discussed.

621-52 : 681.142 **1059**

Simulation—Its Place in System Design.—H. H. Goode. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1501–1506.) Discussion with particular reference to the application of analogue and digital computers.

621.3.012.8 : 629.11.012.8 **1060**

Construction of an Electrical Analogue of a Motor-Car Suspension System.—R. Lansard. (*Onde élect.*, July 1951, Vol. 31, No. 292, pp. 307–312.)

621.316.7 **1061**

Automatic Control.—A. Tustin. (*Nature, Lond.*, 8th Sept. 1951, Vol. 168, No. 4271, pp. 404–406.) Report of conference organized by the Department of Scientific and Industrial Research, July 1951. Research aspects of the subject were discussed, and the relation between work in different fields was emphasized.

621.316.722.1 **1062**

Automatic Control of Inaccessible Terminal Voltages.—R. L. Cosgriff & E. H. Gamble. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 434–442.)

621.317.083.7 + 621.395.44] : 621.316.1 **1063**

Telemetering Systems and Channels for a Large Interconnected Power System.—G. K. Duff. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, pp. 796–801.) Essential text of 1951 A.I.E.E. Summer General Meeting paper.

621.365.54† **1064**

Induction Heater Control System.—R. W. Ketchledge. (*Bell Lab. Rec.*, Sept. 1951, Vol. 29, No. 9, pp. 405–409.) A photometric comparison method, based on the radiation from the heated work piece, is used for controlling the operation of the power source in brazing and soldering work. Powers up to 40 kW at 10 kc/s are thus controlled.

621.38.001.8 **1065**

Industrial Metal-Detector Design.—C. R. Schafer. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 86–91.) Discussion of the principles of design for specific purposes, with description and circuit diagrams of two R.C.A. metal detectors.

621.383 : 621.384.6 **1066**

Photoelectric Control Circuits for the Ion Source of a Pressure-Type Electrostatic Generator.—L. O. Herwig. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 668–672.)

621.384.3 : 621.383.27† **1067**

An Image Converter to Extend the Useful Range of Photo-multipliers to Longer Wavelengths.—F. R. Holiday & W. Wild. (*J. sci. Instrum.*, Sept. 1951, Vol. 28, No. 9, pp. 282–283.) A spectrographic application is discussed in which a photomultiplier whose upper wavelength limit is 0·63 μ is to be used for measurements up to 1·2 μ. The adaptation is effected by interposing an infrared-sensitive image converter with Ni-treated ZnS screen between source and multiplier.

621.384.6† **1068**

Linear Electron Accelerator to One Million Volts.—A. T. Starr, G. King & L. Lewin. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 186–194.) Problems arising in the design of various types of linear accelerator are discussed. A description of one using an E_0 guide with irises is given and its performance is outlined: a velocity corresponding to 1–2 MeV is achieved in a metre-length of guide. A possible line of development for a 10-MeV accelerator is briefly considered.

621.384.6† : 621.319.339 **1069**

The Electrostatic Accelerator as a Source of Ionizing Energy.—J. G. Trump. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, pp. 781–787.) Essential text of 1951 A.I.E.E. Summer General Meeting paper. Discussion of application of e.s. accelerators in nuclear research, for production of very-high-voltage X rays, and for sterilization by means of high-energy electrons.

621.384.612.2† : 621.396.6 **1070**

Radio Frequency for a Synchrocyclotron.—A. J. Poté. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 100–105.) Description of the construction of the oscillator and modulator units, with the connections to the dee, of a f.m. cyclotron. Vacuum-type capacitors were found effective for insulation of the dee, on which an average voltage of 8·5 kV was obtained with an oscillator anode power input of 4·5 kW.

621.385.833 **1071**

Inorganic Replication in Electron Microscopy.—C. J. Calbick. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 798–824.)

621.385.833 **1072**

A Theory of the Nearly Symmetrical Independent Electrostatic [electron] **Lens.**—É. Regenstreif. (*Ann. Radioélect.*, July 1951, Vol. 6, No. 25, pp. 244–267.) Explicit formulae are established for the fundamental optical properties of the elliptical lens in terms of its geometrical and electrical structure. See also 2793 of 1951.

621.385.833 **1073**

An Experimental Study of the Illuminating System of the Electron Microscope.—M. L. De. (*Indian J. Phys.*, July 1950, Vol. 24, No. 7, pp. 303–308.) Observations were made of the variation of electron-crossover size with grid-filament spacing for the electron microscope at Calcutta [1747 of 1949 (Dasgupta et al.)]. Results are discussed in relation to the quality of the final image.

621.385.833 **1074**

The Aberrations of Magnetic Electron Lenses due to Asymmetries.—P. A. Sturrock. (*Philos. Trans. A*, 6th July 1951, Vol. 243, No. 868, pp. 387–429.) The relation between machining defects of the objective lens in an electron microscope and the resulting aberrations is investigated mathematically. A computational procedure is developed for fixing tolerances for a proposed lens design.

621.385.833 **1075**

On a New Test Method for Spherical Aberration of Electron Lenses.—O. Klemperer. (*Proc. phys. Soc.*, 1st Sept. 1951, Vol. 64, No. 381B, pp. 790–794.) "The focus of rays from a lens with spherical aberration appears as a spot surrounded by a discrete halo ring if a diaphragm with fine circular aperture is placed across the beam in front of the focus. The diameter of the halo allows an estimate of the magnitude of the aberration involved. The geometry of rays forming the halo is explained here by schematic drawings. The practical application of the halo test is illustrated by some examples. In particular, the negative spherical aberration produced by an electronic space charge in a saddle field lens is demonstrated."

621.385.833 **1076**

The Electron-Optical Characteristics of Electrostatic Systems with Apertures Deviating from the Idealized Form.—B. M. Rabin, A. M. Strashkevich & L. S. Khin. (*Zh. tekh. Fiz.*, April 1951, Vol. 21, No. 4, pp. 438–447.) The fields of rectangular and elliptic apertures were investigated experimentally using an electrolyte tank. Formulae are derived for determining the potential distribution with great accuracy.

621.385.833 : 541.183.5 **1077**

The Use of the Field Emission Electron Microscope in Adsorption Studies of W on W and Ba on W.—J. A. Becker. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 907–932.)

621.387.424† **1078**

Construction of the External-Cathode Geiger Counter.—M. L. MacKnight & R. L. Chasson. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, pp. 700–701.) Description of the method adopted for quantity production.

621.387.424† **1079**

A Circular Geiger-Mueller Counter.—G. Ensell & S. D. Chatterjee. (*Rev. sci. Instrum.*, Sept. 1951, Vol. 22, No. 9, p. 700.) Description of the method of construction.

621.38.001.8 **1080**

Electronics. [Book Review]—J. Millman & S. Seely. Publishers: McGraw-Hill, New York, 2nd edn 1951, 559 pp., $7.25. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, p. 1578.) ". . . primarily intended as a textbook to set the ground work for specialized courses in communications, electron-tubes, industrial electronics, etc., it would be a welcome addition to the reference shelf of most practicing engineers."

PROPAGATION OF WAVES

538.566 : 551.510.535 **1081**

The Production of Harmonics in the Ionosphere.—

K. Försterling & H. O. Wüster. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 22–31. In German.) Fuller discussion than that previously given (721 of 1951).

621.396.11 **1082**

The Source of Long-Distance Backscatter.—W. G. Abel & L. C. Edwards. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1538–1541.) By comparing the delay time of the leading edge of the back scatter with that of the response from a beacon transponder, the source of back scatter was shown to be the ground at and beyond the skip distance. The scatter forming the leading edge of the pattern sometimes arrived from directions off the principal axis of the aerial, and preceded the amplitude peak due to scatter along the axis.

621.396.11 : 551.510.535 **1083**

The Paths of an Electromagnetic Signal in the Ionosphere.—É. Argence. (*C. R. Acad. Sci., Paris*, 10th Sept. 1951, Vol. 233, No. 11, pp. 607–608.) The approximate formulae derived previously for the refractive index of an ionized medium (3094 of 1951) are used as a basis for further investigations of wave paths in the plane of the magnetic meridian. Analysis is given for the ordinary and the extraordinary ray for vertical incidence, and for the ordinary ray for oblique incidence. Three types of path are possible in each case.

621.396.11.029.62 **1084**

Propagation of V.H.F. via Sporadic E.—T. W. Bennington. (*Wireless World*, Jan. 1952, Vol. 58, No. 1, pp. 5–9.) Clouds of abnormally dense ionization occurring within the normal E region can enable waves of frequency up to 100 Mc/s to be propagated over oblique paths. Temperate-zone-type records of sporadic-E obtained by vertical-incidence measurements at Slough, Fraserburgh, De Bilt, Lindau, Freiburg, Domont and Poitiers are analysed. It appears possible to trace the growth and movement of these clouds from observations at a number of stations.

621.396.11.029.62 **1085**

U.S.W. Propagation in the 30–100-Mc/s Range.—J. Grosskopf. (*Fernmeldetech. Z.*, Sept. & Oct. 1951, Vol. 4, Nos. 9 & 10, pp. 411–414 & 441–451.) Field-strength measurements made by the German Post Office over distances within the range of vision are analysed. Factors investigated include the influence of ground reflections near transmitter or receiver on the shape of the field, the influence of undulations or hilliness on attenuation, and the diffracting effect of obstacles. Irregularities in the field-strength/distance curves are traced to relatively simple causes such as double reflections. Diffraction measurements indicate that in general the classical diffraction formulae are inapplicable, and that the analysis of the propagation process must take account of the nature of the terrain.

621.396.8.029.55 **1086**

Reception of Transatlantic Signals of Frequencies near 30 Mc/s.—J. Maire. (*Ann. Radioélect.*, July 1951, Vol. 6, No. 25, pp. 197–204.) Results of systematic reception tests made near Paris from 1937 to 1940 and from 1948 onwards are shown diagrammatically and discussed. The transmissions recorded were from the New York region, including harmonics of commercial transmissions, from WWV and amateur stations, and also from Buenos Aires. Regularity of reception throughout the above two periods is related to the sunspot cycle, the time of sunset at the mid-point of the path, and the predicted m.u.f. Reception of the Buenos Aires transmissions on about 27·5 Mc/s was, with minor exceptions, consistently good during much of the daytime from 1946 to 1950, and was almost completely free from echo disturbances.

621.396.812.3 **1087**

Further Statistics of Fade-Outs.—D. Stranz. (*J. atmos. terr. Phys.*, 1951, Vol. 2, No. 1, pp. 79–82.) Investigation, for the year 1949, of the weak absorption effect of the corpuscular cone emitted by the sun, previously detected by the statistical method of sample days, gave a result very similar to that for 1948 (3146 of 1950).

621.396.812.3 : 551.510.535 **1088**

Some Random Fading Records with Short-Wave Signals.—P. M. Das & S. R. Khastgir. (*Indian J. Phys.*, July 1950, Vol. 24, No. 7, pp. 277–280.) Rapid variations of the intensity at Dacca of 4·84-Mc/s signals from Calcutta did not agree with Rayleigh's random-scattering formula. Possible explanations of the discrepancy based on diverse reflections from the ionosphere are discussed. See also 981 of 1951 (Khastgir & Das).

RECEPTION

621.396.62 + 534.874.1] : 519.272.11 **1089**

The Correlation Function in the Analysis of Directive Wave Propagation.—Nodtvedt. (See 882.)

621.396[.397].621 **1090**

The Design of a Combined Television and Radio Receiver.—A. B. Bamford. (*J. Televis. Soc.*, July/Sept. 1951, Vol. 6, No. 7, pp. 253–263.) Details are given of a circuit designed to facilitate the production of a range of receivers providing complete coverage of the television channels and the medium- and long-wave broadcasting bands, using a common chassis and as many common sub-assemblies as possible.

621.396.621 **1091**

Radio Feeder Unit.—J. F. O. Vaughan. (*Wireless World*, Dec. 1951, Vol. 57, No. 12, pp. 480–484.) Detailed description of a pretuned receiver providing switched selection of four stations, three medium-wave and one long-wave. The circuit comprises two r.f. stages, diode detector and a.f. amplifier to compensate for losses due to tone-control circuits, with separate a.f. valve for pickup input. The output is suitable for feeding to a high-quality amplifier.

621.396.621 **1092**

Sensitive T.R.F. Receiver.—S. W. Amos & G. G. Johnstone. (*Wireless World*, Nov. 1951, Vol. 57, No. 11, pp. 452–456.) Details are given of the circuit and adjustment of a 3-valve receiver with 'amplified a.g.c.', for medium and long waves.

621.396.621 : 621.396.619.13 **1093**

Design for an F.M. Receiver.—J. G. Spencer. (*Wireless World*, Nov. & Dec. 1951, Vol. 57, Nos. 11 & 12, pp. 440–444 & 487–490.) Detailed description of a simple 7-valve receiver for the 90-Mc/s band, designed to be comparable in cost with a medium-price a.m. receiver while realizing the improvement in background noise associated with f.m. A triple-diode-triode valve serves for both discriminator and first audio amplifier. The i.f.-alignment method is described and performance figures are given.

621.396.621.54 **1094**

The Ionodyne, invented for the Ionophone: will it replace the Superheterodyne?—M. Bonhomme. (*Toute la Radio*, Nov. 1951, No. 160, pp. 291–295.) Discussion of a proposed receiver circuit for use with the ionophone (896 above). Two oscillators are used, one corresponding to the local oscillator of the normal superheterodyne, while the second is tuned to the frequency of the received carrier wave, by which it is accurately synchronized. This enables the depth of modulation of the input signal to be varied, a variable proportion of the input signal and of an unmodulated voltage from the synchronized oscillator being applied to the mixer together with the local-oscillator voltage. The arrangement thus resembles Tucker's synchrodyne circuit (525 and 526 of 1948), but in the ionodyne the synchronization is effected in phase, whereas it is in antiphase in the synchrodyne.

621.396.822 **1095**

The Power Spectrum of a Narrow-Band Noise Passed through a Nonlinear Impedance Element.—J. L. McLucas & R. C. Raymond. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1211–1213.) Noise centred at 1·6 Mc/s, with the spectral shape of an error function with standard deviation of 3·8 kc/s, was passed through a Type-1N34 crystal diode. The observed spectral distributions of output power up to the sixth harmonic are in good qualitative agreement with Middleton's theory (3238 of 1948).

621.396.822 **1096**

On the Theory of Random Noise. Phenomenological Models: Parts 1 & 2.—D. Middleton. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1143–1163. Correction, *ibid.*, Nov. 1951, Vol. 22, No. 11, p. 1326.) Various models of electron noise are considered: (*a*) non-overlapping periodic noise waves, (*b*) non-overlapping non-periodic disturbances, (*c*) Poisson noise. An attempt is made to determine explicitly the second-order probability density in the important stationary cases. For the first two models this is impracticable except in the simplest cases, while for Poisson noise an explicit treatment is possible for impulsive random noise, nearly normal random noise and normal random noise. In part 1 the general probability density is formally obtained; in part 2 the distribution density for nearly random noise is derived and the first and second (second-order) moments and energy spectral distributions are determined.

621.396.62 + 621.396.645 **1097**

Verstärker und Empfänger (Amplifiers and Receivers). [Book Review]—Strutt. (See 955.)

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **1098**

Entropy of a Transmission System.—J. A. Ville. (*Câbles & Transmission, Paris*, July 1951, Vol. 5, No. 3, pp. 189–198.) Entropy is interpreted in terms of the number of binary code signals required for the transmission of an intelligible message. A single information unit is expressed as the sum of three terms representing information, redundancy and error. Limiting values of entropy for known probability conditions are evaluated.

621.39.001.11 **1099**

Information Theory and Most Efficient Codings for Communication or Memory Devices.—L. Brillouin. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1108–1111.) Shannon's theorem about the capacity of a channel is discussed; the most efficient coding is the one yielding the most probable distribution of the code symbols. A specific rule is obtained for this most probable distribution. Most efficient coding is essential for communication channels or memory devices in large-scale computers.

621.394.441 **1100**

Two New Voice-Frequency Telegraph Systems.—H. Gardère. (*Câbles & Transmission, Paris*, July 1951, Vol. 5, No. 3, pp. 199–225.) Technical and economic aspects of multiplex telegraphy are reviewed. Critical discussion of various proposed systems indicates that those using ph.m. or f.m. are preferable. A description is given of a ph.m system.

621.395.44 + 621.317.083.7] : 621.316.1 **1101**

Telemetering Systems and Channels for a Large Interconnected Power System.—G. K. Duff. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, pp. 796–801.) Essential text of 1951 A.I.E.E. Summer General Meeting paper.

621.395.44 : 621.316.1 **1102**

Co-ordination of a Power-Line Carrier Network.—G. E. Burridge & A. S. Jong. (*Elect. Engng, N.Y.*, Sept. 1951, Vol. 70, No. 9, p. 803.) Summary of 1951 A.I.E.E. Summer General Meeting paper.

621.396.619.11/.13 : 621.396.8 **1103**

Comparative Tests of Communication by Amplitude Modulation and Frequency Modulation carried out by the Société française Radioélectrique.—(*Ann. Radioélect.*, July 1951, Vol. 6, No. 25, pp. 287–288.) In districts where traffic and industrial electrical interference are very considerable, f.m. was found definitely superior to a.m. but where traffic and interference were of small amount the two systems were found nearly equal as regards range and quality of modulation.

621.396.619.16 **1104**

Cross-Talk Considerations in Time-Division Multiplex Systems.—S. Moskowitz, L. Diven & L. Feit. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 209–216.) Reprint. See 740 of 1951.

621.396.65 **1105**

Long-Distance Telephone Links and the Microwave Copenhagen-Skamlebaek Installation.—G. Pedersen. (*Teleteknik, Copenhagen*, July 1951, Vol. 2, No. 2, pp. 153–155.) General considerations with reference to the Danish Post Office 24-channel link, operating on 20-cm wavelength.

621.396.65 **1106**

Radio-Telephone Communications in Jutland.—H. Schouboe-Madsen. (*Teleteknik, Copenhagen*, July 1951, Vol. 2, No. 2, pp. 155–160.) An account of the 24-channel Aarhus-Hammel link, which uses pulse-phase modulation and operates on frequencies between 1 400 and 1 500 Mc/s, and a description of emergency R/T equipment to provide a service for islands, when cable breakage interrupts communications.

621.396.65 **1107**

24-Channel Pulse-Modulated Microwave Telephony Equipment of Danish Manufacture.—L. Christensen. (*Teleteknik, Copenhagen*, July 1951, Vol. 2, No. 2, pp. 161–167.) Description of equipment for the Aarhus-Hammel and Copenhagen-Skamlebaek links.

621.396.65 **1108**

Notes on an Automatic Radio-Frequency Repeater System.—J. A. Craig. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1524–1529.) Discussion of the basic principles of r.f. relay systems and description of the system and equipment for the broadcasting network covering the whole of Cuba.

621.396.65.029.64 **1109**

The TD-2 Microwave Radio Relay System.—A. A. Roetken, K. D. Smith & R. W. Friis. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1041–1077.) The relay system is designed to supplement the coaxial-cable telephone system for long-distance communication and to provide facilities for wide-band signals. The system uses f.m. and provides twelve wide-band channels, six in each direction, spaced 40 Mc/s apart in the 3 700–4 200-Mc/s band. Each channel may be used to provide a large number of message circuits 4 kc/s wide or a single video circuit 4 Mc/s wide. The repeater stations are located 25–30 miles apart in line-of-sight steps and only the main stations, which are situated every few hundred miles, are attended. Details of the equipment are given. See also 450 of 1951 (Clutts).

621.396.65.029.64 **1110**
An Unattended Broad-band Microwave Repeater for the TD-2 Radio Relay System.—R. W. Friis & K. D. Smith. (*Elect. Engng, N.Y.*, Nov. 1951, Vol. 70, No. 11, pp. 976–981.) Essentials of 1951 A.I.E.E. Summer General Meeting paper. Description of the equipment for converting the incoming 4-kMc/s signals to an i.f. band centred at 70 Mc/s (where 75% of the required gain is provided) and for reconversion to a microwave band offset 40 Mc/s from the carrier frequency of the incoming signals. See also 1109 above (Roetken, Smith & Friis).

621.396.66 : 621.396.97 **1111**
Precision A.M. Frequency Monitor.—R. S. McKinney. (*Broadcast News*, May/June 1951, No. 64, pp. 48–50.) Description and circuit of R.C.A. Type-BW-11A broadcasting monitor. The transmitter signal, after passage through an untuned wideband amplifier, is heterodyned with an accurate reference signal of frequency 1 kc/s below the assigned frequency.

621.396.665.1 **1112**
Instantaneous Compandors on Narrow-Band Speech Channels.—J. C. Lozier. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1214–1220.) The conditions are examined for the distortionless transmission of compressed speech over a system with a pass band no wider than that occupied by the uncompressed speech. The analysis indicates that more severe requirements must be imposed on the attenuation and phase characteristics of the system when this reduced-bandwidth mode of operation is used.

621.396.712 **1113**
The Operation of Broadcasting Studios, and New Equipment of the S.F.R. (Société française Radio-électrique).—Cordonnier & Bernard. (*Ann. Radioélect.*, July 1951, Vol. 6, No. 25, pp. 268–285.) Economic and technical aspects of studio and control-room arrangements are discussed, particularly personnel requirements, size of apparatus, ease of operation and running costs. A comprehensive description is given of studio arrangements and types of equipment proposed by the S.F.R. to satisfy the essential conditions of both quality and efficiency.

621.396.931 **1114**
Radio for Taxis.—(*Wireless World*, Dec. 1951, Vol. 57, No. 12, pp. 491–493.) Description and discussion of two-way a.m. radiotelephone system for taxicabs in London. Operation is in the frequency band 100–184 Mc/s. Only one fixed station, located on very high ground, is used.

621.396.932/.933 : 621.317.7 **1115**
Remote Monitoring of Naval and Air Service Transmissions.—Marique. (See 1038.)

621.396.932 : 623.98 (44) **1116**
Radiocommunication in the [French] Navy.—P. David. (*Onde élect.*, July 1951, Vol. 31, No. 292, pp. 297–306.) Survey of the problems of equipping a naval unit for long- and short-range communication, navigation, telecontrol and enemy interception. Selection of frequencies, disposition of apparatus in the vessel, arrangement of aerials, and recent improvements in d.f. technique are discussed, reference being made to relevant published papers.

SUBSIDIARY APPARATUS

621-526 **1117**
Nonlinear Techniques for Improving Servo Performance.—D. McDonald. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 400–421.)

621-526 **1118**
Servo Mechanisms.—A. L. Whiteley. (*Proc. Instn elect. Engrs*, Part 1, Sept. 1951, Vol. 98, No. 113, pp. 289–297.) A review of progress. 46 references.

621.314.5 **1119**
Comparative Representation of Various D.C./A.C. Converters.—H. Tigler. (*Arch. elekt. Übertragung*, Sept. 1951, Vol. 5, No. 9, p. 439.) Correction to paper noted in 504 of February.

621.314.6.012.8 **1120**
The 'Transrector', the Quadripole Equivalent of a Rectifier.—H. Marko. (*Frequenz*, July 1951, Vol. 5, No. 7, pp. 196–203.) A method of calculating the approximate impedance characteristics of a rectifier circuit is developed. The rectifier is represented as an ideal transformer connecting the a.c. and d.c. parts of the circuit; the resistances of these parts determine the turns ratio to be assumed. Simple transformer theory is then applied. Applications to a mains rectifier, a modulator and a meter rectifier illustrate the method.

621.314.632.1 **1121**
The Copper Oxide Rectifier.—W. H. Brattain. (*Rev. mod. Phys.*, July 1951, Vol. 23, No. 3, pp. 203–212.) The conductivity of the oxide layer in Cu_2O rectifiers can be explained on the basis of the usual energy-band representation of semiconductors only by assuming the presence of some donor-type impurities in addition to the usual acceptor type. Applying the Schottky theory of the space-charge exhaustion layer, the dependence of the capacitance of the rectifier on bias voltage shows that the density of ion charge in the rectifying layer is of the same order of magnitude as the difference between the donors and acceptors found from the conductivity, thus checking the theory. Analysis of the d.c. characteristic and its dependence on temperature indicates that the Cu-Cu_2O interface is not uniform, but acts like a patchy surface over which the potential maximum varies in magnitude.

621.316.543.2.029.64 : 621.392.26† **1122**
A High-Speed K-Band Switch.—M. W. Long. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1566–1567.) Description and performance characteristics of a 3-way rotary switch for 1·25-cm waves, with application to rapid scanning.

621.316.722.1 : 621.314.3 **1123**
Magnetic-Amplifier Voltage Regulator.—J. L. Wolff. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 45–51.) Description of equipment which uses a magnetic amplifier as the control element. Regulation is effected at 160–400 V to within 0·5% for load currents from 0 to 500 mA (line voltage constant), or for line-voltage changes of ± 10% and load currents from 0 to 300 mA.

TELEVISION AND PHOTOTELEGRAPHY

621.396/.397].621 **1124**
The Design of a Combined Television and Radio Receiver.—Bamford. (See 1090.)

621.397.335 : 621.317.35 **1125**
Producing and Interpreting the Pulse Cross.—D. M. Launer. (*TV Engng, N.Y.*, Sept. & Nov. 1951, Vol. 2,

Nos. 9 & 11, pp. 12–15, 29 & 21, 29.) An oscillographic technique developed by Loughren & Bailey (1937 of 1941) is adapted for examining the phasing and duration of synchronizing signals transmitted with the television picture signal. The 'pulse-cross' pattern is obtained by modulating the intensity of the scanning beam and adjusting the phase of the synchronizing pulses so that they appear near the centre of the raster. Photographs of patterns thus obtained are shown and interpreted.

621.397.5 **1126**

The Evaluation of Picture Quality with Special Reference to Television Systems: Part 1.—L. C. Jesty & N. R. Phelp. In 275 of January, for 'Vol. 15' please read 'Vol. 14'.

621.397.5 **1127**

The Evaluation of Picture Quality with Special Reference to Television Systems: Part 2.—L. C. Jesty & N. R. Phelp. (*Marconi Rev.*, 4th Quarter 1951, Vol. 14, No. 103, pp. 156–186.) Previous experimental results (275 of January) show useful correlation with system performance and provide a basis for an explanation of the relations between the limits of resolution, brightness and signal/noise ratio (i.e., graininess). Further experimental data are needed at low values of contrast. A fundamental performance figure is proposed, measured in terms of the number of quanta per picture element per picture required for a signal/noise ratio of unity. For comparing systems, both the viewing distance and the ratio viewing-distance/picture-height must be specified. The quality of 35-mm motion pictures can be attained in a 600-line television system with spot wobbling. Application of the results to the determination of television standards is discussed.

621.397.5 **1128**

Flicker in Television Pictures.—J. Haantjes & F. W. de Vrijer. (*Philips tech. Rev.*, Sept. 1951, Vol. 13, No. 3, pp. 55–60.) Flicker in television pictures is compared with that in motion-picture projection. The origin of flicker and the related properties of the eye are discussed. Experiments show that with a suitable phosphor for the c.r. screen and a frame frequency of 50 c/s, a high-light luminance of 200 cd/m² is permissible without causing troublesome flicker.

621.397.5 : 535.623 **1129**

Oscillating Color Sequence in Color TV.—R. G. Peters. (*TV Engng, N.Y.*, Sept. 1951, Vol. 2, No. 9, pp. 18–19.) An account of the essential features of the technique described by Loughlin (826 of March).

621.397.5 : 535.767 : 621.398 **1130**

Stereo-Television in Remote Control.—H. R. Johnston, C. A. Hermanson & H. L. Hull. (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 170–177.) Description of an experimental system developed as a viewing system for remote manipulation in atomic-energy research.

621.397.5 : 535.88 **1131**

Large Screen Television in the Festival of Britain Telekinema.—T. M. C. Lance. (*J. Televis. Soc.*, July/Sept. 1951, Vol. 6, No. 7, pp. 266–271.) The general problems connected with the production of bright television pictures of diagonal up to 26 ft and with methods of distributing programmes to a large number of cinemas are outlined. A description is given of instantaneous projection equipment for both 405- and 625-line pictures and of its associated power supply and remote-control units. A cinema specially designed to house this equipment, as well as film projectors, is described. The projection box is built under the front of the balcony to keep the television projector on the centre line of the metallized-fabric directional-viewing screen. The latter has an illuminated surround of fixed low intensity to enhance the contrast of the television picture. The foyer, which can be used as a studio for closed-circuit work, has one wall of glass allowing the public to see into the projection room.

621.397.6 **1132**

Considerations on Television Pickup Technique: Part 2.—G. Goebel. (*Fernmeldetech. Z.*, Sept. 1951, Vol. 4, No. 9, pp. 403–406.) The discussion of studio problems presented in part 1 (3483 of 1940) is continued, with particular reference to the illumination required when using an iconoscope or supericonoscope camera tube.

621.397.61.001.41 **1133**

Measuring Television Transmitter Amplitude Characteristics.—J. Ruston. (*Tele-Tech*, Sept. 1951, Vol. 10, No. 9, pp. 30–31 . . 72.) Description of a method of measurement in which the frequency of a test voltage applied to the video input terminals of the transmitter is swept over a 5-Mc/s range. A c.r.o. is used to display the desired overall transmitter response. An inductive probe inserted in the output transmission line of the transmitter picks up a signal which is mixed with the 'swept' signal by means of a crystal diode.

621.397.611.2 **1134**

The Vidicon — Photoconductive Camera Tube.—P. K. Weimer, S. V. Forgue & R. R. Goodrich. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part 1, pp. 306–313.) Reprint. See 2040 of 1950.

621.397.611.2 : 535.215 : 546.863.221 **1135**

Properties of some Photoconductors, principally Antimony Trisulfide.—Forgue, Goodrich & Cope. (See 1008.)

621.397.62 **1136**

A Time-Selection Circuit for Frame Sync-Separation.—W. R. Luckett. (*Electronic Engng*, Sept. 1951, Vol. 23, No. 283, p. 343.) Description, with diagram, of a circuit that is stable in operation and does not require critical adjustment of component values. The basic principle is the selection of pulses that occur within a fixed period after the occurrence of a reference pulse.

621.397.62 **1137**

A Single-Valve Line-Scan and E.H.T. Generator.—C. H. Banthorpe. (*Electronic Engng*, Sept. 1951, Vol. 23, No. 283, pp. 349–352.) A brief review of single-valve circuits is given. Their advantages of compactness, cheapness and stability are offset by poor linearity and interdependence of controls. By suitable design and the use of modern components these disadvantages can be minimized. The design of a suitable circuit is described in some detail.

621.397.62 **1138**

Paris Television Show.—A. V. J. Martin. (*Wireless World*, Nov. 1951, Vol. 57, No. 11, pp. 459–460.) Of the 87 different receivers shown, the majority were for the 819-line standard, and six were for both the 819-line and the 441-line standard. See also *Radio franç.*, Oct. 1951, No. 10, pp. 18–22.

621.397.62 : 621.396.662 **1139**

Utilization of Printed Components in a Television Tuner.—D. Mackey & E. Sass. (*RCA Rev.*, Sept. 1951, Vol. 12, No. 3, Part 1, pp. 293–302.) "Printed coils are found to be practical in a turret-type television tuner which has individual coil strips. The coils on these strips are produced by means of a relatively unconventional photoetching process. Possible economic and electrical advantages of using the printed coils are indicated, and a circuit employing the coils is described."

621.397.62 : 621.396.68 **1140**

Ringing-Choke E.H.T. Systems.—W. T. Cocking. (*Wireless World*, Nov. & Dec. 1951, Vol. 57, Nos. 11 & 12, pp. 444–447 & 513–516.) Systems are described in which line-scan sawtooth voltage is applied to the grid of a pentode with a RLC circuit connected to its anode; voltage oscillations produced when the anode current cuts off are passed to a rectifier of either half-wave or voltage-doubler type. Because the high-voltage and scanning circuits are separate, optimum design and voltage regulation are facilitated.

621.397.621 : 621.314.2 **1141**

Study of a Line [-scan] **Transformer considered as a Pulse Transformer.**—H. Gilloux. (*Radio franç.*, Sept. 1951, No. 9, pp. 1–5.) Design calculations are made for a transformer for the timebase circuit of a 441-line or 819-line scanning system. The critical value of leakage inductance is determined by considering a pulse transformer operating with pulse duration twice the flyback time. The transformer windings comprise five series-connected primary sections and four interleaved parallel-connected secondary sections. Additional windings may be incorporated to supply a voltage-doubler circuit for the h.v. required for the c.r. tube.

621.397.621.2 **1142**

The Davisson Cathode Ray Television Tube using Deflection Modulation.—A. G. Jensen. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 855–866.) Description of an experimental c.r. tube, nearly 5 ft long, designed by C. J. Davisson and used during 1937 in demonstrations of television transmission from New York to Philadelphia by coaxial cable. See also 4462 of 1938 (Strieby).

621.397.82 **1143**

Internal Television-Receiver Interference.—B. Amos & W. Heiser. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 122–125.) Minimum interference from harmonics of the sound and video channel intermediate frequencies occurs when 21·75 Mc/s is used for the sound i.f. The possible harmonics due to the video detector are analysed and optimum frequencies are given for intercarrier-sound and 41-Mc/s operation.

621.397.82 : 621.365.55† **1144**

Curing Industrial TVI.—P. S. Rand, A. J. Riley & J. J. Lamb. (*QST*, Sept. 1951, Vol. 35, No. 9, pp. 29–33.) Account of methods for preventing radiation from r.f. heating installations. Essential requirements are complete shielding and r.f. filtering for all leads.

621.397.5 **1145**

Elements of Television Systems. [Book Review]—G. E. Anner. Publishers: Prentice-Hall, New York, 1951, 771 pp., \$10.35. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, p. 1579.) " . . . written primarily for physicists and engineers rather than for those who are interested in the very elementary aspects of the subject."

621.397.621 **1146**

Theory and Design of Television Receivers. [Book Review]—S. Deutsch. Publishers: McGraw-Hill, New York, 1951, 521 pp., \$6.50. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1577–1578.) "The author has attempted to cover a large field and, on the whole, has done so in an excellent fashion."

VALVES AND THERMIONICS

621.383 **1147**

Relations for Barrier-Layer Photocells and Theory derived therefrom of the Barrier-Layer Photo- and Rectifier-Effect.—P. E. Weber. (*Optik*, July 1951, Vol. 8, No. 7, pp. 302–310.) From the variation of the no-load terminal voltage and of the short-circuit current of barrier-layer photocells with the intensity of illumination an equivalent circuit is derived and its resistance values are determined, assuming a constant e.m.f. This circuit consists of a voltage source with internal resistance inversely proportional to the radiation intensity, and a constant resistance across the terminals. These resistances are estimated. From this equivalent circuit, theory is developed according to which the barrier layer consists of oxygen atoms bound to copper atoms and, in order that it may be conductive, one electron of the N-shell must be taken from an unbound copper atom of the Cu_2O layer by photoionization or polarization.

621.383.4 : 546.482.31 **1148**

Photoconductive Cells of Cadmium Selenide.—E. Schwarz. (*Proc. phys. Soc.*, 1st Aug. 1950, Vol. 63, No. 368B, pp. 624–625.) Polycrystalline layers of CdSe have been produced by methods previously described (1102 of 1949). Preliminary measurements on cells with such layers indicate a wide range of sensitivity. The properties of the cells are governed by the amount and form of oxygen present in the CdSe layer. Experiments support the view that adsorption of oxygen on the grain boundaries is an essential condition for the production of photocells with a high quantum yield. See also 3578 of 1949.

621.383.4 : 546.482.31 **1149**

Photoconductive Cells of Cadmium Selenide.—E. Schwarz. (*Proc. phys. Soc.*, 1st Sept. 1951, Vol. 64, No. 381B, pp. 821–822.) Data are tabulated for an improved cell (1148 above) of high sensitivity for low applied voltage (6V). Minimum detectable energy for a bandwidth of 1 c/s is of the order of 2×10^{-12} W for a wavelength of $0{\cdot}7\mu$. The sensitivity range extends from the X-ray region to about $1{\cdot}4\mu$ in the infrared, and the permissible current through the cell is of the order of 20–30 mA.

621.385 **1150**

Rare Metals in Electron Tubes.—D. A. Wright. (*J. Brit. Instn Radio Engrs*, Sept. 1951, Vol. 11, No. 9, pp. 381–392.) 1951 Radio Convention (London) paper. The part played by rare metals in overcoming difficulties encountered in the manufacture of different types of electron tubes is discussed. Typical applications mentioned are: electrodes for high-temperature operation; getters; high-emission electrodes; grids with reduced emission; materials for brazing and soldering.

621.385.029.63/.64 **1151**

Effect of Hydrostatic Pressure in an Electron Beam on the Operation of Traveling-Wave Devices.—P. Parzen & L. Goldstein. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 228–232.) Reprint. See 2580 of 1951.

621.385.029.63/.64 : 621.396.822 **1152**

Noise Measurements on a Travelling-Wave Tube.—B. N. Agdur & C. G. L. Åsdal. (*Acta polyt., Stockholm*, 1951, No. 86, 9 pp.) Noise measurements were carried out on a valve operating at a frequency of about 10 kMc/s. Measured values of gain as a function of beam current (1–5 mA) are in fairly good agreement with Rydbeck's theory (2962 of 1948), according to which the gain should vary as $I_b^{\frac{1}{3}}$, where I_b is the beam current. The observed gain was about 12 db for a beam current of 3 mA. Noise factor increased from 16 db at $I_b = 1$ mA to about 19 db at $I_b = 5$ mA. This change is of the same order as that predicted by present theories, which yield noise-factor values apparently much too high.

621.385.029.64/.65 **1153**

Periodic-Waveguide Traveling-Wave Amplifier for Medium Powers.—G. C. Dewey, P. Parzen & T. J. Marchese. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 220–227.) Reprint. See 2047 of 1951.

621.385.029.64 **1154**

Experimental Observation of Double-Stream Amplification.—B. N. Agdur. (*Acta polyt., Stockholm*, 1951, No. 86, 13 pp.) Description of the construction and performance of a double-stream microwave valve giving a gain of 30 db at 3 kMc/s. The two helices, wound with 30 turns of 0·4-mm Mo wire, are 25 mm long with a diameter of 8 mm, the intervening interaction space being of length 20 cm. Accelerating voltages of 250–290 V were found suitable. Beam current was 30 mA. The change of gain characteristics with relative modulation of the beams is in fair agreement with the theory of Rydbeck & Forsgren (2866 of 1951). The gain/cathode-voltage-difference curve obtained is generally similar to that given by Haeff (1825 of 1949).

621.385.032.213.2 **1155**

Thermionic Data for some Capillary Metal Cathodes.—H. Katz & K. L. Rau. (*Frequenz*, July 1951, Vol. 5, No. 7, pp. 192–196.) The variations of the emission from a thoria cathode coating on a tungsten core with step variations of the heater current are shown graphically and explained by means of Richardson's law. With a metallic thorium emitting surface on a porous tungsten base the operating temperature can be considerably lower than for the thoria cathode; at 1 370°C a current density of 1 A/cm² was obtained. Emission characteristics for a metallic barium layer are shown, the cathode construction being modified to provide an auxiliary heating coil for the stock material.

621.385.032.216 **1156**

The Work Function for Oxide Cathodes.—G. Jähnig. (*Funk u. Ton*, Feb. 1951, Vol. 5, No. 2, pp. 95–100.) Treatment of the thermodynamics of electron emission by Fermi statistics. The influence of cathode activation on work function is discussed.

621.385.032.216 **1157**

The Effect of Ion Bombardment on the Emission from Oxide-Coated Cathodes.—P. A. Redhead. (*Canad. J. Phys.*, Sept. 1951, Vol. 29, No. 5, pp. 362–369.) The decay of emission was measured under different sets of conditions. It is caused by a reduction of the number of impurity centres in the cathode coating and by sputtering of the cathode surface by heavy ions. The life of an oxide cathode is not reduced even when the current density is increased by 20 times, provided there is no positive-ion bombardment.

621.385.032.216 : 621.386 **1158**

A Study of Oxide-Coated Cathode by X-Ray Diffraction Method.—E. Yamaka. (*J. appl. Phys.*, Aug. 1951, Vol. 22, No. 8, pp. 1087–1088.) Correlated measurements of the crystal size and thermionic emission of BaO, (BaSr)O and (BaSrCa)O as affected by previous heat treatment.

621.385.1 : 537.525.5 **1159**

Studies of Externally Heated Hot-Cathode Arcs: Part 1—Modes of the Discharge.—Malter, Johnson & Webster. (See 959.)

621.385.15 : 537.533.8 **1160**

Development of a Secondary-Electron Multiplier with Aluminium-Oxide-Caesium Emitters.—D. M. Khorosh. (*Zh. tekh. Fiz.*, April 1951, Vol. 21, No. 4, pp. 397–404.) Experiments were conducted with oxidized Al foil treated with Cs vapour. The coefficient of secondary emission of the surfaces is 6–7·5. The characteristics are free from any anomalies due to primary electron emission. The construction and production of five-stage electron multipliers using these surfaces is described. The amplification obtained is 900–1 600 for 300 V per stage and 2 000–2 500 for 500 V per stage.

621.385.2/.3 : 546.817.231 **1161**

Crystal Diode and Triode Action in Lead Selenide.—C. A. Hogarth. (*Proc. phys. Soc.*, 1st Sept. 1951, Vol. 64, No. 381B, pp. 822–823.) Single crystals of both *p*- and *n*-type PbSe were examined; whisker contacts of tungsten and phosphor-bronze were used. With rectifier connection, peak inverse voltages were usually between 4 and 7 V and rectification ratios up to 300 : 1 were recorded for small signals, values of 50 : 1 being obtained without difficulty. With transistor connection, power gains of 1·5 to 2 could be obtained, but current gain greater than unity was not observed, the usual value being about 0·3.

621.385.2 : 537.315.6 **1162**

A Method of Calculating the Space Distribution of Potential in a Diode.—H. Bonifas. (*Onde élect.*, Aug./Sept. 1951, Vol. 31, Nos. 293/294, pp. 363–369.) Two expressions are derived which involve the ratio u/α, where u is the initial electron emission velocity normal to the cathode and α the velocity corresponding to cathode temperature. Numerical integration of these expressions determines the curve representing the distribution of potential between the anode and cathode. Calculation for a particular case gives results in good agreement with experimental values.

621.385.2 : 537.525.92 **1163**

The Space-Charge Smoothing Factor.—C. S. Bull. (*Proc. Instn elect. Engrs*, Part III, Nov. 1951, Vol. 98, No. 56, pp. 470–472.) Discussion on 2058 of 1951. See also 2871 of 1951.

621.385.2 : 537.533 **1164**

Electron Streams in a Diode.—F. Gray. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 830–854.) "A general solution of the electron-stream equations is developed for a parallel plane diode, under the assumption that the electron velocity is single valued. This solution contains all particular solutions. It serves to unify the wave theory and the particle theory of electron flow, and it is an approximation for multi-velocity streams over a wide range of conditions."

621.385.3/.4 **1165**

The Development of Electron Tubes for a New Coaxial Transmission System.—G. T. Ford & E. J. Walsh. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 2, pp. 1103–1128.) The fundamental problem in the development of valves for wide-band systems was to devise means of obtaining closer grid-cathode spacings without sacrificing life performance. The closer spacings have been made possible by the use of rigid control-grid supports which can be wound with wire of very small diameter under tension. A flat winding is produced which can be mounted very close to a flat cathode. Construction and operation details are given of three new valves developed for the L3 coaxial-cable system: Type-435A and Type-436A tetrodes and Type-437A triode.

621.385.3/.5].032.24 : 621.317.311 **1166**

Method for the Determination of Extremely Small Grid Currents in Valves.—H. Köppen. (*Elektrotechnik, Berlin*, Sept. 1951, Vol. 5, No. 9, pp. 431–433.) Causes and effects of grid current, especially in output valves, are discussed. Grid current due to thermionic emission caused by heat from the cathode may be 10^{-10} to 10^{-7} A. A method of production testing for this is suggested, using a galvanometer in the balanced anode circuit of a screened-filament electrometer valve, the grid of which is connected through a switch to the grid of the valve on test. Tests on five EF14 valves are reported.

621.385.3 : 546.289 **1167**

Crystal Triodes.—T. R. Scott. (*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 195–208.) Reprint. See 2592 of 1951.

621.385.3 : 546.289 **1168**

The Junction Transistor.—D. G. F. & R. K. J. (*Electronics*, Nov. 1951, Vol. 24, No. 11, pp. 82–85.) A review of recently published material. The junction transistor consists of a single crystal of Ge having a *p*-type section enclosed between *n*-type end pieces, to which the emitter and collector connections are made, the base connection being made to the *p*-type section. Advantages over the point-contact transistor include (*a*) improvement in noise figure of 20–30 db, the equivalent of an absolute noise figure of between 20 and 10 db at 1 kc/s, (*b*) greater gain (40 to 50 db per stage), (*c*) better electrical and mechanical stability. This type of transistor has a larger barrier capacitance than the present point-contact type, resulting in reduced gain at h.f. By means of impedance mismatching the frequency response can be made uniform to at least 1 Mc/s, with some sacrifice of gain.

21.385.3.029.55 **1169**

Manufacture of a High-Frequency Transmitting Tube.—(*Elect. Commun.*, Sept. 1951, Vol. 28, No. 3, pp. 171–185.) Fully illustrated description of the various processes in the manufacture of Type-F-5918 high-power water-cooled h.f. triodes. These valves have an anode dissipation of 70 kW and operate at frequencies up to 22 Mc/s. In push-pull class-C telegraph service, two valves will give an output of 400 kW. In a class-C anode-modulation amplifier the output from two valves is $>$ 200 kW.

621.385.4 : 537.525.92 **1170**

Space-Charge and Ion-Trapping Effects in Tetrodes.—K. G. Hernqvist. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1541–1547.) In the space between the screen grid and the anode in an evacuated tetrode, a potential minimum is formed by the space charge. At low current levels this effect is small and most secondary electrons from the anode pass to the grid. For higher currents the secondary electron flow is reduced, while at still higher currents a virtual cathode is formed. If residual gas is present, the space-charge potential minimum traps positive ions until an equilibrium state is reached with formation of a plasma region near the anode. By using a square cut-off pulse, it has been shown experimentally that ion trapping occurs, is built up in a finite time and alters the division of current between screen grid and anode in a predicted manner.

621.385.832 **1171**

A Gun for Starting Electrons Straight in a Magnetic Field.—J. R. Pierce. (*Bell Syst. tech. J.*, Oct. 1951, Vol. 30, No. 4, Part 1, pp. 825–829.) "In a simple electron gun consisting of a cathode and two apertured planes held at different potentials, the apertures act as electron lenses. When the gun is immersed in a uniform axial magnetic field the aperture spacings and potentials can be chosen so that the emerging electrons have no radial velocities."

621.385.832 : 537.534 **1172**

Hollow Cathode for Positive-Ion Studies in Cathode-Ray Tubes.—C. H. Bachman, H. Eubank & G. Hall. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1208–1210.)

621.385.832 : 621.318.572 **1173**

New Electronic Tubes Employed as Switches in Communication Engineering: Part 1 — Contact Tubes.—J. L. H. Jonker & Z. van Gelder. (*Philips tech. Rev.*, Sept. 1951, Vol. 13, No. 3, pp. 49–56.) Discussion of the principles and description of the construction of experimental 'contact' tubes which have much greater switching speed than e.m. relays and in which no cleaning of contacts is required. A secondary-emission output electrode is used, so that the input and output signals are in phase. See also 1801 and 3213 of 1950 (Jonker).

621.385.832 : 681.142 **1174**

Improvements in Cathode-Ray Tube Storage: Application to a Parallel Type of Digital Computer.—G. H. Perry. (*Nature, Lond.*, 1st Sept. 1951, Vol. 168, No. 4270, pp. 372–373.) The dot-dash system of charge distribution of c.r. tube screen, when applied to the parallel computer, was found to have several undesirable features which are shown to be absent in the defocus-focus system, which is also suitable for serial storage.

621.386.1 **1175**

Special X-Ray Tubes.—B. Combée & P. J. M. Botden. (*Philips tech. Rev.*, Sept. 1951, Vol. 13, No. 3, pp. 71–80.) Details of three tubes for diagnosis, contact therapy and endotherapy respectively.

621.396.615.141.2 **1176**

Enhanced Emission from Magnetron Cathodes.—R. L. Jepsen & M. W. Muller. (*J. appl. Phys.*, Sept. 1951, Vol. 22, No. 9, pp. 1196–1207.) Violations of the Hull cut-off condition considerably more pronounced than those previously reported were found at anode voltages and magnetic fields much larger than those used by earlier investigators. When enhanced emission occurred in magnetrons with pure-metal cathodes, it was found that maximum current limits existed for each value of the magnetic field. The experimental observations are combined into a self-consistent pattern for static magnetrons. It is suggested that the process causing cathode bombardment and enhanced emission is electronic interaction in the region of relatively dense space charge surrounding the cathode.

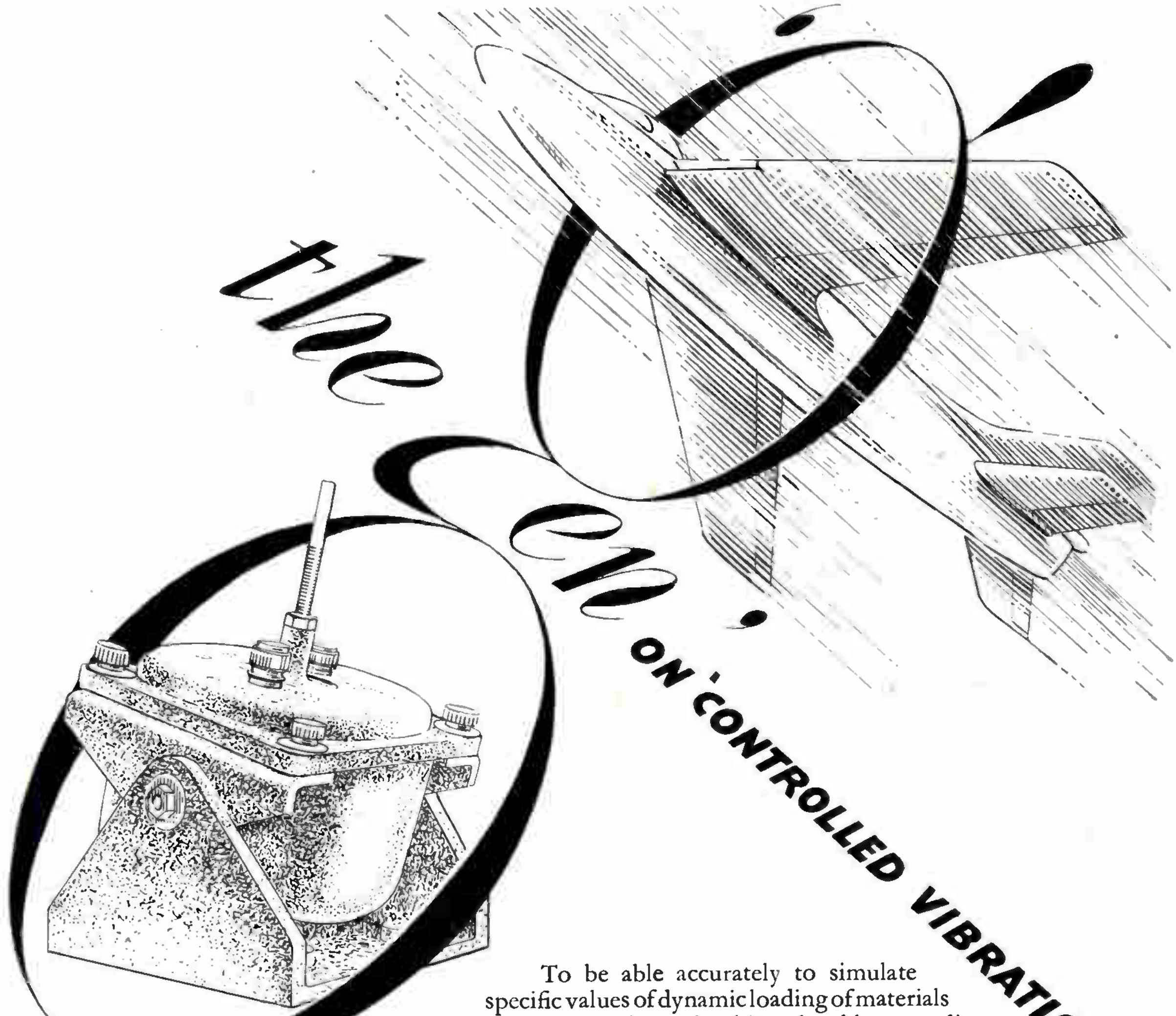
621.396.615.142.2 **1177**

Space-Charge Effects in Reflex Klystrons.—M. Chodorow & V. B. Westburg. (*Proc. Inst. Radio Engrs*, Dec. 1951, Vol. 39, No. 12, pp. 1548–1555.) "Space charge in reflex tubes has an effect which causes considerable departure from the existing reflex theory. To a first order, it modifies the electronic admittance by a bunching effectiveness parameter, designated F, which, in general, takes some value between one and three. Several approximate methods have been used to estimate this F factor and the results are presented graphically for comparison with experimental measurements."

MISCELLANEOUS

061.4 : 621.396.933 **1178**

Aircraft Radio on Show.—(*Wireless World*, Nov. 1951, Vol. 57, No. 11, pp. 448–450.) Features and equipment shown at the exhibition of the Society of British Aircraft Constructors included the design of aircraft structures to accommodate or to serve as aerials; radio altimeters; v.h.f. communication equipment; wide-band aerial amplifiers for reducing multiplicity of aerials at ground stations; distance-measuring equipment operating at 1 kMc/s.



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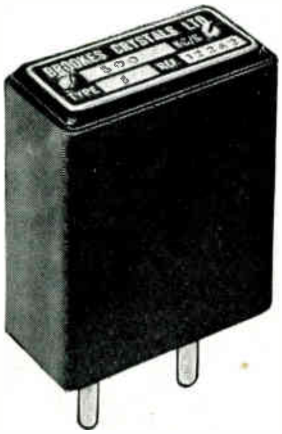

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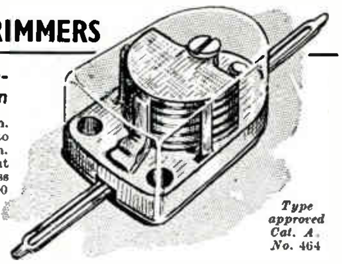

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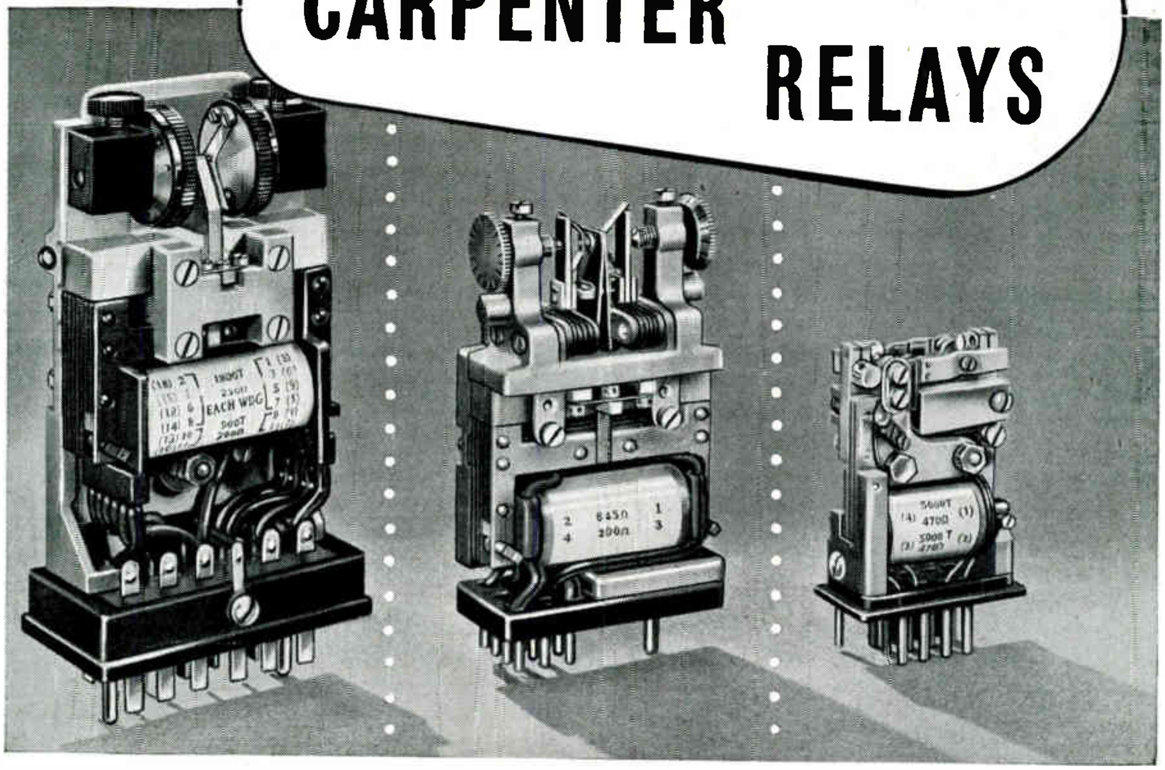

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Apply at once by letter, stating age, full names in block letters, and full particulars of qualifications and experience, and mentioning this paper to the Crown Agents for the Colonies, 4 Millbank, London, S.W.1, quoting on letter M.29100.B. The Crown Agents cannot undertake to acknowledge all applications and will communicate only with applicants selected for further consideration.

Electronic Engineer, with good theoretical background to Higher National Certificate standard and with practical experience in use of valves in audio and radio frequency circuits, is required in the Valve Life Testing Laboratory at the G.E.C. Research Laboratories, Wembley, Middlesex. Apply to the Staff Manager (Ref. GBLC/0/566) stating age and record.

Engineers and Physicists with experience in the following fields are required by the G.E.C. at the Stanmore Laboratories. (*a*) servo-mechanisms; (*b*) microwave aerials, transmitters or receivers; (*c*) small mechanisms; (*d*) electronic circuitry; (*e*) D.C. amplifiers; (*f*) test gear for field trials. Preference will be given to men with good academic qualifications. Applications should be sent to the Staff Manager (Ref. GBLC/551), G.E.C. Research Laboratories, Wembley Middlesex.

Designers and Draughtsmen with experience in electronic or light mechanical engineering are required by the G.E.C. Research Laboratories, Wembley, Middlesex, for work at Stanmore. Vacancies exist for seniors with good academic qualifications and also for juniors. Excellent and interesting openings with good prospects. Apply to the Staff Manager (Ref. GBLC/552), stating age and record.

Development Engineers required for Development Laboratory of light engineering factory in Stockport, experienced in circuit design for Radio, Test-gear, and Electronic Instruments. HNC, City and Guilds (telecommunications) Degree, essential. Write stating age and full details of experience and qualifications.—Box 7630, c/o *Wireless Engineer*.

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Electronic Engineer required for development of specialized electronic machinery. Good academic qualifications, preferably B.Sc., experience in development work and knowledge of practical design are desirable. The post offers good scope for a man to develop new techniques.

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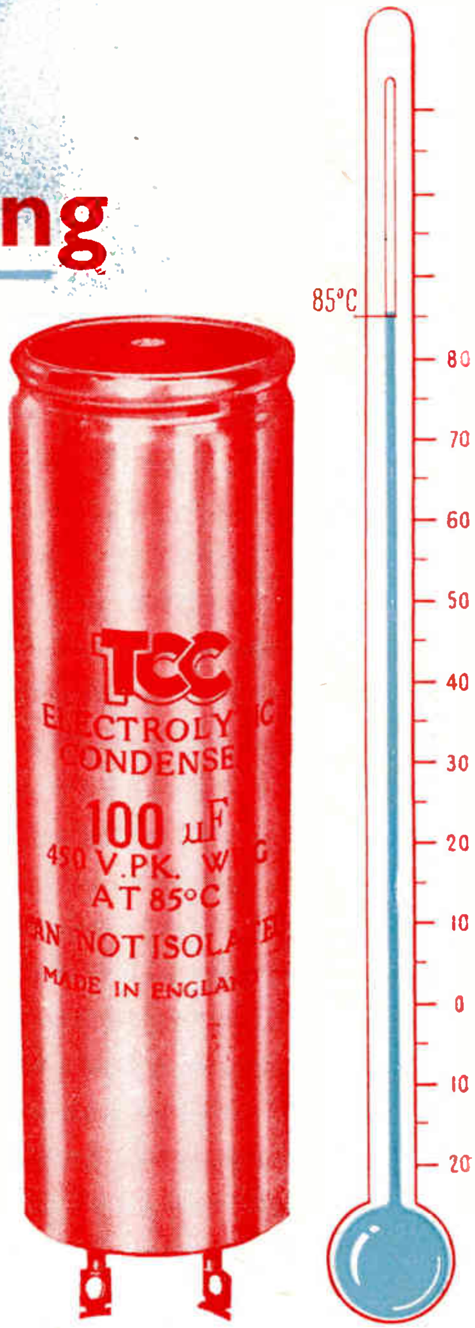

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