ELECTRONIC ENGINEERING

VOL. 29 No. 349

MARCH 1957

Commentary

IN 1945 a committee consisting of members appointed by the Council of the Institution of Electrical Engineers, in consultation with the British Electrical and Allied Manufacturers' Association, the Radio Industry Council and other interested bodies, undertook to look into the question of the practical training of professional electrical engineers.

At that time the re-organization of industry as a whole from a war time to a peace time footing was beginning to take shape and one of the immediate problems demanding attention was an overhaul of the pre-war system of recruitment and subsequent training of technical staff.

The Committee examined the training of both graduate and student apprentices and its report, published two years later, largely set the pattern by which the electrical industry recruited its technical staff.

But in the post war years the electrical industry in this country has doubled in size and, according to statements recently announced, is now Britain's largest exporting industry. This happy state of affairs has, however, led to an increasing demand for professionally qualified electrical and electronic engineers to undertake responsible duties in connexion with research into and the development, installation and operation of a very wide range of equipment.

Not surprisingly, the original committee have found it desirable to revise some of the recommendations made in their report of 1947. They reviewed their earlier work and have now produced another report* dealing with the training of graduates who after a full time education in engineering or science at a University, College of Advanced Technology or Regional College aim to complete their preparation for positions of responsibility as professional electrical and electronic engineers leading to corporate membership of the Institution of Electrical Engineers.

This report, which does not concern itself with graduates in science or mathematics who join the research laboratories of the electrical industry with the intention of devoting themselves to scientific or mathematical research, will be followed later by a further report on the training of student apprentices who seek to reach professional level through the Higher National Certificate or the new Diploma of Technology. It is intended to be no more than a guide to fundamental requirements and recommends that graduates should undergo a two-year course consisting of basic workshop training, general mechanical or electrical training and ending with a final period of directed objective training.

The report underlines what is common knowledge, namely that still more engineers than heretofore will be required if the electrical industry is to play its full part in the country's economy, and on the assumption that the increased numbers will be forthcoming suggests the methods by which they can be most suitably trained in the minimum of time.

The training in the past has for the most part fallen on the larger electrical firms and here the report breaks new ground in stating that the time has now come when this burden can no longer be carried in this way and that smaller firms should join in either by participating in group schemes or by co-operating with the larger firms.

Whether such a suggestion is practical remains to be seen, but it is beyond doubt that the larger firms cannot cope with increasing numbers—they have already been accepting and training between two and three times as many as can be absorbed at the end of the training period. This has not necessarily been a complete loss for it has at least enabled the firm to be more selective when the training is complete and it must be remembered that many trainees are from overseas who eventually return home with a sound knowledge of British electrical techniques and methods—in itself a valuable prestige export.

There nevertheless remains a large number who join the smaller firms in this country having received their training at no expense to their new employers and it is a matter of regret that so many of the smaller firms should have remained content with this state of affairs.

Not all the smaller firms are blameworthy in this respect, for many have for years past run well conducted training schemes, and overcrowding with the larger firms may well be attributable to the desire of the entrants—graduate and apprentice alike—to enter the electrical industry by the larger door. And this desire can be readily understood when one considers the advantages to be gained in so doing.

The fact that an entrant may be no more than a very small cog in a very large machine is no great cause for alarm, nor is it likely that the entrant will worry unduly about his prospects of subsequent employment at the end of his training period. Instead he sees that his training, both practical and theoretical, is well organized and that every encouragement will be given for further study. He will have the companionship of people of his own age and outlook, his leisure activities will be well catered for and in many cases if he is away from home he will be living in a well run hostel.

It is doubtful, therefore, whether the smaller firms can operate as single units in this respect, except at a prohibitive cost to themselves, nor are groups of smaller firms likely to be successful unless they are relatively close to each other.

Perhaps the greatest hope lies in the association of the larger firms concentrated in the industrial areas of the Midlands and the North and the smaller firms who surround them.

Such co-operative training schemes will not be easy to work out but there is no doubt that the electrical industry stands to benefit enormously.

* *The Training of Graduates — A Report of the Joint Committee on Practical Training in the Electrical Engineering Industry.* (*Institution of Electrical Engineers, 1957*).

Out-of-Channel Radiation from Mobile F.M. V.H.F. Transmitters

By A. L. Rowles*, B.Sc.

The peak-clipper method of restricting frequency deviation in mobile f.m. transmitters produces harmonic distortion of the modulating waveform; this can result in adjacent channel interference. Some theoretical consideration is given to the problem of eliminating the effects of clipped-waveform sidebands by the use of a post-limiter filter in the transmitter.

IN allocating frequencies to v.h.f.-equipped services it is clearly essential to lay down limits of bandwidth beyond which radiation must not be of a magnitude likely to interfere with users of adjacent channels.

To achieve this requirement in the case of frequency modulation, it is necessary to prevent the transmitter frequency deviation from rising to values corresponding to transient peaks occurring in a modulating speech waveform adjusted to have an adequate mean level. A common method is to introduce some form of peak clipper in the audio chain, acting as a fixed 'gate' removing any portions of the waveform exceeding a given amplitude.

Thus the modulating signal never exceeds a certain value, and the maximum deviation is fixed.

The maximum permissible deviation is a function of the channel separation imposed, among other factors, and is normally specified by the licensing authority.

However, Fourier analysis shows that squaring of the wave peaks implies the creation of many harmonics of the fundamental (input) frequency. The amplitudes of these components are small for all but the first three or four present, the rate of diminution following a geometric progression.

Now these harmonic components may be considered as individual modulating frequencies producing at the transmitter aerial certain modulation indices. These indices being very small indeed for the frequencies of interest, only the first order sidebands are generally of importance. However, compared with the higher order sidebands of fundamental modulating tones, it is possible for the first order sidebands due to harmonic components in clipped waveforms still to be of appreciable magnitude at frequencies removed from the carrier at which the former reach obscurity.

The sidebands due to distortion products can well be the major factor determining minimum channel separation.

Sideband Power Spectrum

In the case of frequency modulation by a pure tone of frequency f, the modulated wave can be represented by:

$$\begin{aligned} I &= I_0 \sin(2\pi Ft + M \sin ft) \\ &= I_0 [J_0(M) \sin 2\pi Ft \\ &\quad + J_1(M)\left\{\sin 2\pi(F+f)t - \sin 2\pi(F-f)t\right\} \\ &\quad + J_2(M)\left\{\sin 2\pi(F+2f)t + \sin 2\pi(F-2f)t\right\} \\ &\quad + \ldots\ldots\ldots\ldots\ldots\ldots \\ &\quad + J_p(M)\left\{\sin 2\pi(F+pf)t + (-1)^p \sin 2\pi(F-pf)t\right\} \\ &\quad + \ldots\ldots\ldots\ldots\ldots\ldots] \end{aligned}$$

The amplitude of each of the p^{th} side-current pairs (spaced every $\pm pf$ from the carrier frequency F) is given, in terms of the unmodulated carrier amplitude I_0, by the coefficient

$$I_p = I_0 J_p(M),$$

where $p = 0, 1, 2$, etc.

- $p = 0$, referring to signal at unmodulated carrier frequency.
- $J_p(M)$ = Bessel coefficient of first kind of order p and argument M.
- M = modulation index = $(\Delta F/f)$.
- ΔF = maximum deviation of radio frequency from unmodulated carrier frequency.

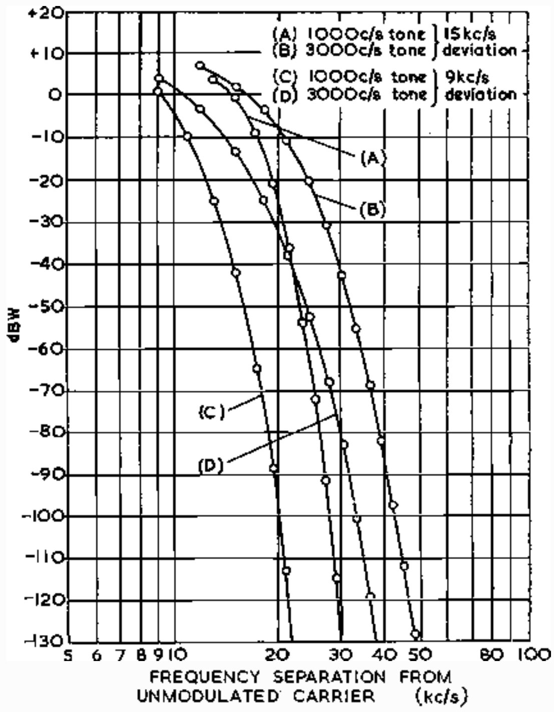


Fig. 1. Sideband power envelopes of 25W transmitter, frequency-modulated by a pure tone

The power radiated at each side-frequency may thus be expressed in terms of the power of the unmodulated carrier as

$$-20 \log (I_0/I_p) = -20 \log \frac{1}{J_p(\Delta F/f)} \text{ dB relative to carrier.}$$

It will be convenient to consider the absolute power levels of sidebands, and, as an e.r.p. of 25W (=14dBW) is the maximum permitted in the United Kingdom for the type of service under consideration, figures will be based upon this carrier power.

* *Marconi's Wireless Telegraph Company Ltd.*

Then:

$$W_p = \left(14 - 20\log\frac{1}{J_p(\Delta F/f)}\right)\text{dBW} \quad \ldots\ldots\ldots (1)$$

An identical result applies to corrected phase modulation, that is, to the modulation resulting from a phase modulator preceded by an audio network having a frequency response proportional to $1/f$. Using equation (1) and a table of Bessel coefficients, sideband powers of a transmitter modulated by a pure 1kc/s or 3kc/s tone to a deviation of 15kc/s are given in Table 1. Since we are interested only in energies at extreme separation from the carrier, the envelope of side-frequency powers drawn for each tone represents the maximum possible energy content for any modulating frequency up to that of the tone. These envelopes are curves (A) and (B) of Fig. 1.

Curves (C) and (D) are derived similarly, but correspond to a deviation of 9kc/s.

To calculate the powers radiated in the sidebands due to distortion products, it is first necessary to find an expression for the amplitudes of the harmonic components. Some basis must be established, and as example is considered a tone input which suffers a degree of clipping of 10dB, this being a degree readily attained in practice.

The product of a sinusoidal tone which has undergone a severe degree of clipping is closely represented by a symmetrical trapezoid wave, as illustrated in Fig. 2.

The amplitude of the n^{th} harmonic component of such a wave is given by:

$$C_n = \frac{2A(d+t)}{T}\left[\frac{\sin(n\pi t/T)}{n\pi t/T}\right]\left[\frac{\sin[n\pi(d+t)/T]}{n\pi(d+t)/T}\right]$$

$n = 1, 2, 3$, etc.

If the degree of clipping of positive and negative half-cycles is equal, then $d + t = T/2$ and hence

$$C_n = A\left[\frac{\sin(n\pi t/T)}{n\pi t/T}\right]\left[\frac{\sin(n\pi/2)}{n\pi/2}\right] \quad \ldots\ldots\ldots\ldots (2)$$

A value for t may be deduced from the 10dB degree of clipping assumed. For this, the amplitude of the clipper output is 0·316 of that of the input sine wave, as illustrated in Fig. 3.

TABLE 1

Power in the Sidebands of a 25W Transmitter Frequency Modulated by a Pure Tone

f=1kc/s		15kc/s deviation		9kc/s deviation	
		$\frac{\Delta F}{f}=15$		$\frac{\Delta F}{f}=9$	
Order p	Frequency separation of sideband from carrier (kc/s)	Coefficient $J_p(15)$	$14\text{-}20\log\frac{1}{J_p(15)}$ (dBW)	Coefficient $J_p(9)$	$14\text{-}20\log\frac{1}{J_p(9)}$ (dBW)
			(A)		(C)
9	9	—	—	0·215	+ 0·7
11	11	—	—	$0{\cdot}622\times10^{-1}$	− 10·1
13	13	0·279	+ 2·9	$0{\cdot}108\times10^{-1}$	25·3
15	15	0·181	− 0·8	$0{\cdot}129\times10^{-2}$	− 43·8
17	17	$0{\cdot}665\times10^{-1}$	− 9·5	$0{\cdot}112\times10^{-3}$	− 65·0
19	19	$0{\cdot}166\times10^{-1}$	− 21·6	$0{\cdot}750\times10^{-5}$	− 88·5
21	21	$0{\cdot}305\times10^{-2}$	− 36·3	$0{\cdot}399\times10^{-6}$	− 114·0
23	23	$0{\cdot}438\times10^{-3}$	− 53·2	$0{\cdot}173\times10^{-7}$	−141·2
25	25	$0{\cdot}506\times10^{-4}$	− 71·9	—	—
27	27	$0{\cdot}483\times10^{-5}$	− 92·3	—	—
29	29	$0{\cdot}388\times10^{-6}$	−114·2	—	—
31	31	$0{\cdot}267\times10^{-7}$	−137·5	—	—

f=3kc/s		$\frac{\Delta F}{f}=5$		$\frac{\Delta F}{f}=3$	
Order p	Frequency separation of sideband from carrier (kc/s)	Coefficient $J_p(5)$	$14\text{–}20\log\frac{1}{J_p(5)}$ (dBW)	Coefficient $J_p(3)$	$14\text{-}20\log\frac{1}{J_p(3)}$ (dBW)
			(B)		(D)
3	9	—	—	0·309	+ 3·8
4	12	0·391	+ 5·8	0·132	− 3·6
5	15	0·261	+ 2·3	$0{\cdot}430\times10^{-1}$	− 13·3
6	18	0·131	− 3·6	$0{\cdot}114\times10^{-1}$	− 24·9
7	21	$0{\cdot}534\times10^{-1}$	− 11·4	$0{\cdot}255\times10^{-2}$	− 37·9
8	24	$0{\cdot}184\times10^{-1}$	− 20·7	$0{\cdot}493\times10^{-3}$	− 52·1
9	27	$0{\cdot}552\times10^{-2}$	− 31·2	$0{\cdot}844\times10^{-4}$	− 67·5
10	30	$0{\cdot}147\times10^{-2}$	− 42·6	$0{\cdot}129\times10^{-4}$	− 83·8
11	33	$0{\cdot}351\times10^{-3}$	− 55·1	$0{\cdot}179\times10^{-5}$	− 100·9
12	36	$0{\cdot}763\times10^{-4}$	− 68·4	$0{\cdot}228\times10^{-6}$	− 118·8
13	39	$0{\cdot}152\times10^{-4}$	− 82·4	$0{\cdot}266\times10^{-7}$	− 137·5
14	42	$0{\cdot}280\times10^{-5}$	− 97·1	—	—
15	45	$0{\cdot}480\times10^{-6}$	−112·4	—	—
16	48	$0{\cdot}768\times10^{-7}$	−128·3	—	—

For $\theta = \theta'$, $y = 0{\cdot}316$.

Now θ' corresponds to interval $t/2$

$\therefore 0{\cdot}316 = \sin\theta' = \sin[(2\pi/T) \,.\, t/2]$

$\therefore t = (T/\pi)\sin^{-1} 0{\cdot}316$ (in radians)

$= T/180 \times 18{\cdot}4$ (in degrees, from tables).

$\therefore t = 0{\cdot}102T$.

Substituting this value of t in equation (2) and making other reductions,

$$C_n = A \,.\, \frac{\sin(0{\cdot}102n\pi)}{0{\cdot}102n\pi} \,.\, \frac{\sin(n\pi/2)}{n\pi/2}$$

The last factor indicates even order harmonics to be

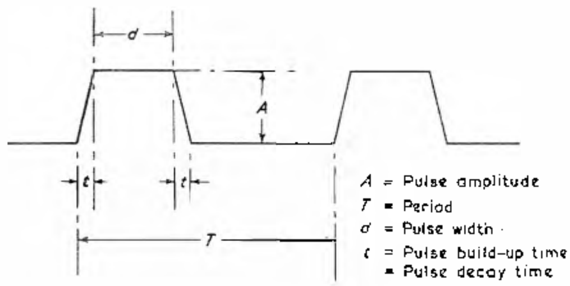


Fig. 2. Symmetrical trapezoid wave

absent and assists in determining the sign, with which we are not concerned. Again, since for the purposes of the present consideration only the 5th and higher harmonics are of interest, it is not necessary to consider the second factor as a function of the type $(\sin x/x)$. Putting $\sin(0{\cdot}102\,n\pi) = 1$ and $\sin(n\pi/2) = 1$ one is merely considering the positive envelope of possible values, and disregarding a 'phase' factor. Any error introduced is 'safe', in that it shows, at worst, the sideband to be greater than is actually the case.

The expression for the amplitude of the n^{th} harmonic is taken as

$$C_n = \frac{A}{0{\cdot}102n\pi \,.\, (n\pi/2)}$$

or

$$C_n = (1{\cdot}99/n^2)\ A \qquad (3)$$

Values of this coefficient for the harmonics it is required to consider are given in Table 2. It is seen that they range in order from 10^{-2} to 10^{-4}.

The general expression for a Bessel coefficient of the first kind is

$$J_p(M) = \sum_{k=0}^{\infty} \frac{(-1)^k}{k!(p+k)!} (M/2)^{p+2k}$$

For small arguments M, of the order 10^{-1} or smaller, quite negligible error is introduced by taking only the first term of this summation; (the error is in the pessimistic or 'safe' direction).

Putting $k = 0$:

$$J_p(M) = (1/p!)\ (M/2)^p$$

Neglection of the sidebands corresponding to the second and higher order coefficients is justified by the rate of

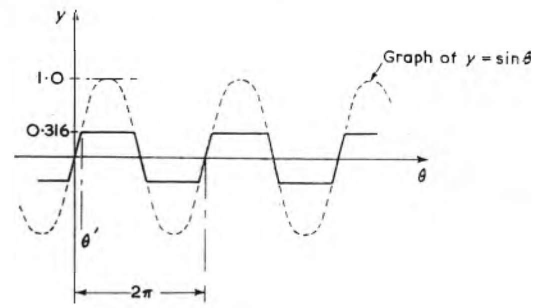


Fig. 3. 10dB clipping of a sinusoidal tone

diminution of the coefficient with increasing order being much greater than the effect of diminution of M with increasing frequency of harmonic, from equation (3). It is sufficient, then, to consider only the case $p = 1$, for which, conditionally,

$$J_1(M) = (M/2) \qquad (4)$$

Using equations (1), (3) and (4), Table 2 gives the powers radiated in each first-order sideband due to frequency modulation of a 25W transmitter by the harmonic content of a 10dB clipped tone. A peak deviation of 15kc/s is considered, and the input tone is 1kc/s.

Since $J_1(M)$ in equation (4) is a linear function of $M[\propto (1/f)]$ and C_n in equation (3) is proportional to $(1/n^2) \propto (1/f^2)$, the graph of sideband powers, when plotted on a decibel ordinate and logarithmic frequency abscissa scale, is a straight line having slope

$$m = 20\log[f^3/(2f^3)] = -60\log 2 = -18\text{dB/octave}$$

The envelope of sideband powers for a 1kc/s clipped tone is curve (a) of Fig. 4.

For an input tone of another frequency

$$C_n = (\text{const}/n^2)$$

TABLE 2

Power in the Sidebands of a 25W Transmitter Frequency-Modulated by a Tone which has Suffered 10dB of Clipping

	15 kc/s deviation by 1kc/s tone which has suffered 10dB clipping							9kc/s deviation	
Sideband frequency (separation from carrier) f (kc/s)	Order of harmonic n	Relative amplitude $\frac{C_n}{A}$	Deviation $\Delta F = 15\frac{C_n}{A}$	Relative 1st order sideband amplitude $\frac{1}{2}\frac{\Delta F}{f}$	Relative 1st order sideband power $20\log\left(\frac{1}{2}\frac{\Delta F}{f}\right)$ (dB)	1st order sideband power due to 25W transmitter (+14dBW) (dBW)	For sideband f due to 3kc/s clipped tone $(a)+19{\cdot}1$dB (dBW)	1st order sideband power due to 1 kc/s clipped tone $(a)-4{\cdot}4$dB (dBW)	1st order sideband power due to 3 kc/s clipped tone $(b)-4{\cdot}4$dB (dBW)
						(a)	(b)	(c)	(d)
9·0	9	$2{\cdot}46\times10^{-2}$	$3{\cdot}69\times10^{-1}$	$0{\cdot}205\times10^{-1}$	33·8	−19·8	—	−24·2	—
15·0	15	$8{\cdot}85\times10^{-3}$	$1{\cdot}33\times10^{-1}$	$0{\cdot}443\times10^{-2}$	−47·1	−33·1	−14·0	−37·5	−18·4
45·0	45	$9{\cdot}80\times10^{-4}$	$1{\cdot}47\times10^{-2}$	$0{\cdot}163\times10^{-3}$	−75·8	−61·8	−42·7	66·2	−47·1
135·0	135	$1{\cdot}09\times10^{-4}$	$1{\cdot}64\times10^{-3}$	$0{\cdot}608\times10^{-5}$	−104·3	90·3	−71·2	−94·7	−75·6

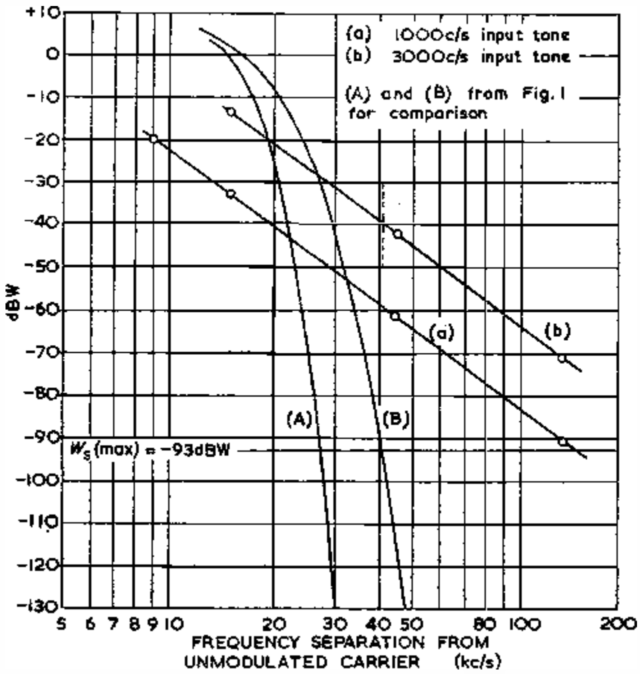


Fig. 4. Sideband power envelopes of 25W transmitter, frequency-modulated to ±15kc/s deviation by 10dB clipped tone

Hence $C_n'/C_n = (n/n')^2$

where C_n' and n' are the amplitude and order appropriate to a harmonic at the same frequency as C_n, but due to an input tone of different frequency.

Then, for the particular case of a 3kc/s input tone,

$$n' = (1/3)n$$

$$\therefore (C_n'/C_n) = 9.$$

Thus, at a given frequency in the sideband spectrum chosen for convenience to be a multiple of 1kc/s and 3kc/s, the power due to a 3kc/s clipped input tone is

$$20 \log (C_n'/C_n) = +19{\cdot}1\text{dB}$$

referred to the power due to a 1kc/s clipped tone.

The envelope in the case of a 3kc/s tone is therefore also a straight line, parallel to curve (a) and displaced upwards by 19·1dB. This is curve (b) of Fig. 4. Curves (A) and (B) of Fig. 1 are superimposed upon this figure for reference.

The 9kc/s deviation case is readily obtained by adding

$$20 \log (9/15) = -4{\cdot}4\text{dB}$$

to the ordinates of curves (a) and (b) respectively, to produce curve (c) and (d). These are shown in Fig. 5, together with curves (C) and (D) of Fig. 1 for reference.

A Criterion for Interference Effects

Measurements indicate that an unwanted signal arriving at an f.m. receiver simultaneously with a wanted signal will be ineffective in creating interference if its level at the limiter is 6dB below the latter. The capture effect, by which a receiver favours the larger of two signals, becomes effective at about this difference of levels. For a good v.h.f. equipment, a wanted carrier resulting in a signal of 0·7μV across the 50Ω load presented by the receiver at its aerial terminals, is just satisfactory. This represents an absolute power level of −140dBW, so that the unwanted signal must contribute a power of less than −146dBW within the pass-band if it is not to be classified as 'interfering'.

It is a reasonable supposition that the aerials of equipments on neighbouring channels (one transmitting and the other receiving) will be separated by not less than 500ft. This condition usually holds between fixed stations and is readily justified in the case of fixed/mobile or mobile/mobile combinations, by observing that the time spent in closer proximity than 500ft will be negligibly small. The following formula then gives approximately the degree of protection which may be assumed to exist by virtue of this spatial separation:

$$A = [20 \log (45DF)]\text{dB}$$

where A = attenuation in decibels

D = separation in miles

F = frequency in megacycles per second.

At a representative frequency of 100Mc/s,

$$A = 53\text{dB}.$$

Hence the unwanted transmitter must not radiate a signal having greater power than

$$W_{s(max)} = (-146 + 53)\text{dBW}$$
$$= -93\text{dBW} \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

in a sideband which under normal conditions could fall within the pass-band of a receiver operating on an adjacent channel.

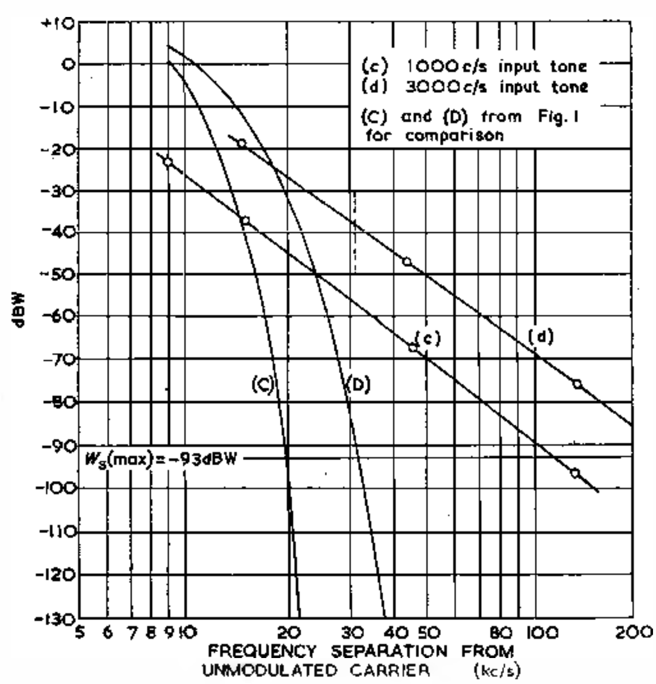
Applying the above power level condition to the case illustrated in Fig. 4, it is seen that pure-tone sidebands of a ±15kc/s deviation transmitter are potentially interfering at frequencies up to ±41kc/s removed from the carrier frequency. This is an inescapable consequence of the process of frequency modulation.

It is also seen, however, that sidebands due to modulation by clipped tones have appreciable power levels at frequency separations from the unmodulated carrier well in excess of ±41kc/s.

Post-Limiter Filtering

Since any avoidable bandwidth occupancy reduces the ultimate information-carrying potentialities of available v.h.f. spectrum, it is desirable to consider the possibility of eliminating clipped-waveform sidebands. To this end, one

Fig. 5. Sideband power envelopes of 25W transmitter, frequency modulated to ±9 kc/s deviation by a 10dB clipped tone



cannot do better than ensure that such sidebands have levels within the pure-tone sideband envelope out to ± 41kc/s, and lower than -93dBW for all frequencies beyond this. For if such conditions obtain, then modulation by clipped waveforms cannot be the cause of interference, and the undesirable sidebands are effectively eliminated.

A method of achieving the foregoing result is to arrange the audio response between limiter and modulator to have a cut beyond the upper limit of the speech band. The extent of this cut must be sufficient to make the clipped-waveform sidebands always smaller than the high order sidebands of a 3kc/s tone at full modulation out to the point at which these latter fall to below -93dBW. Beyond this point, the attenuation introduced must maintain the sideband levels below this interference criterion.

The minimum performance of a filter required to achieve this condition, for the 15kc/s case of Fig. 4, is clearly represented by the difference between curves (B) and (b) for frequencies above the point of intersection; this is given in Table 3 and plotted in Fig. 6(1).

TABLE 3

FREQUENCY (kc/s)	CURVE (B) (dBW)	CURVE (b) (dBW)	DIFFERENCE (B)-(b) (dB)
26·2	−28·6	−28·6	0
28·0	−35·5	−30·4	− 5·1
30·0	−43·0	−32·2	−10·8
32·0	−51·5	−34·0	−17·5
34·0	−60·0	−35·5	−24·5
36·0	−69·5	−37·0	−32·5
38·0	−79·0	−38·3	−40·7
40·0	−88·0	−39·8	−48·2
41·0	−93·0	−40·4	−52·6

Similarly for ± 9kc/s deviation from Fig. 5, the required filter curve is given by the difference between curves (D) and (d), as shown in Table 4 and Fig. 6(2).

TABLE 4

FREQUENCY (kc/s)	CURVE (D) (dBW)	CURVE (d) (dBW)	DIFFERENCE (D)–(d) (dB)
17·3	−22·4	−22·4	0
20·0	−33·2	−26·1	− 7·1
22·0	−42·0	−28·6	−13·4
24·0	−51·5	−31·0	−20·5
26·0	−62·0	−33·0	−29·0
28·0	−72·5	−35·1	−37·4
30·0	−84·0	−36·9	−47·1
31·6	−93·0	−38·2	−54·8

To use filters having such response characteristics would be impractical; moreover, advantage is to be gained by letting the filtering action have its turning point at 3kc/s, from which a fairly gradual slope extends to beyond the lowest required point. Its purpose being to define the upper speech band limit, this filtering action immediately above 3kc/s should take place before the limiter; however, the content of speech above 3kc/s being relatively small, little is lost and considerable economy effected by combining the two filtering actions into a single low-pass filter of moderate performance.

Over the speech band, normally 300c/s to 3kc/s, the relation between limiter output and final frequency deviation should be flat. This applies to corrected phase modulation as well as frequency modulation. A tendency to pre-emphasize results in an effective increase in harmonic content of the modulating signal. This is apparent not only as an increase in the measured distortion but, of greater

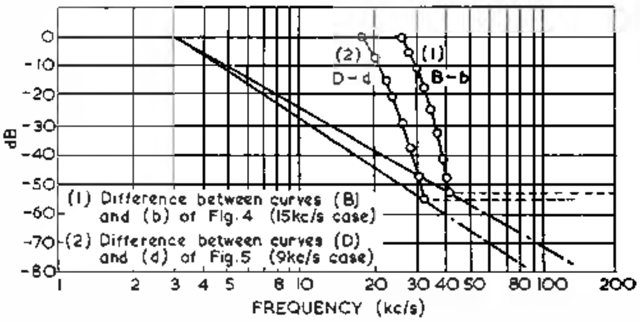


Fig. 6. Minimum filter requirements

consequence, in producing higher energies in the sidebands more remote from the unmodulated carrier. A desired pre-emphasis characteristic may be applied prior to the limiter, but complete correction of the emphasis inherent in a phase modulator should occur between limiter and modulator (immediately preceding the latter, for preference), together with the higher frequency filtering advocated.

It is convenient to have a comparative measure of the higher frequency filtering requirements, which may be obtained by considering the slope (convenient as dB/octave) of an idealized filter curve, the straight line joining the 'knee' point (3kc/s, 0dB) to the lowest required point on Fig. 6. In the 15kc/s case, the lowest point is $-52{\cdot}6$dB and occurs at 41kc/s. The number of octaves is given by:

$$N = 3{\cdot}33 \log (f_2/f_1)$$
$$= 3{\cdot}33 \log (41/3)$$
$$= 3{\cdot}78$$

The minimum required filter slope is thus

$$\alpha_{15} = \frac{-52{\cdot}6}{3{\cdot}78} = -13{\cdot}9\text{dB/octave}$$

For the 9kc/s case, a filter cutting at a constant rate from 3kc/s requires a slope of at least

$$\alpha_9 = \frac{-54{\cdot}8}{3{\cdot}33 \log (31{\cdot}6/3)} = \frac{-54{\cdot}8}{3{\cdot}40} = -16{\cdot}1\text{dB/octave}$$

It is important to observe that these rates of cut must be maintained up to frequencies well in excess of the 'audio' definition, in fact to at least 41kc/s and 31·6kc/s respectively. Beyond this limit, the rate of cut may become zero, or positive to a restricted extent, without invalidating the result.

The degrees of cut indicated may be achieved with relatively simple *LC* filter sections. Simple *RC* filter sections have a turnover which is too slow to enable their response to meet the speech-band requirements while cutting effectively at higher frequencies. An alternative to the passive *LC* filter is, however, the active filter combining the asymptotic -18dB/octave cut of a three-section *RC* filter with a selective feedback amplifier. The amplifier is arranged to peak at a point which makes the turnover of the *RC* sections effectively more rapid, the knee occurring at 3kc/s.

The result of applying the specified $-16{\cdot}1$dB/octave post-limiter response cut in the case of ± 9kc/s deviation is illustrated in Fig. 7, for which curve (d′) is obtained from Table 5.

A further necessary condition for the efficacy of this method of restricting sidebands is that no significant non-linearity occurs subsequent to the low-pass filter. The most serious possible cause of such distortion is in the modulator itself: in particular, reactance valves must be capable of handling maximum signals without grid-limiting. When

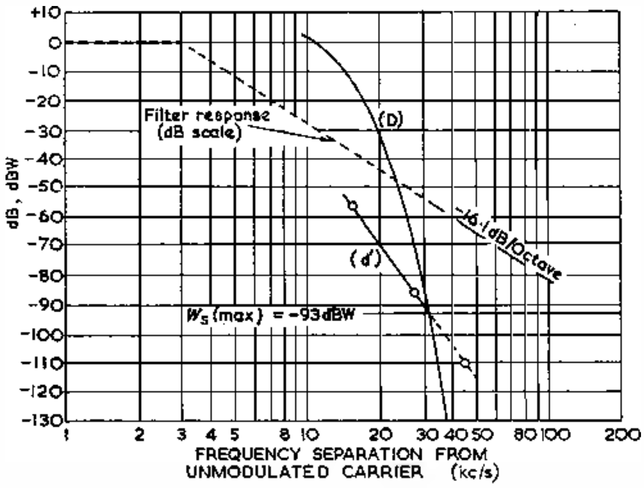


Fig. 7. Suppression of clipped-tone sidebands by post-limiter low-pass filter. (Compare Fig. 5)

TABLE 5

FREQUENCY (kc/s)	SIDEBAND POWER ENVELOPE (from Fig. 5) (dBW)	FILTER RESPONSE (from Fig. 7) (dB)	RESULTING SIDEBAND POWER ENVELOPE (dBW)
	(d)		(d′)
15·0	−18·4	−37·5	−55·9
27·0	−33·0	−52·0	−85·0
45·0	−47·1	−63·0	−110·1

an active filter is employed, the non-linearity condition must be held to apply to the gain stage of this also. Trouble due to this cause is minimized if the gain stage precedes the *RC* sections.

Comments

Mention should be made of several points affecting the results indicated. Fortunately these will be seen to make the results appear pessimistic, so that their stringency, rather than their validity, may be debated.

A true clipped sine-wave, and more particularly the practical result of passing a sine waveform through an actual peak limiter, will have a smaller harmonic content than the trapezoid wave assumed. Also, quite distinct from the mathematical assumptions made to simplify the derivations of the section dealing with the sideband power spectrum, several other factors conspire to make the quantitative results of that section somewhat approximate. For example, due to valve and stray capacitances a section of normal audio circuit is unlikely to pass freely signals having frequencies of the order of 50 to 100kc/s or so, as implied for the results of Figs. 4 and 5. For these reasons the clipped waveform reaching the modulator will appear appreciably 'less square' at the corners, and the reduced magnitudes particularly of the higher harmonics will give rise to a decrease of sideband power with increasing frequency, superimposed upon the indicated slope.

Also, the input signal will in practice be speech waveform, hence the regular and constant nature of the clipped sine wave will be lost. Clipping will occur, however, and 10dB of clipping represents a degree likely to be achieved. Thus the sideband powers calculated will occur only whenever speech sounds give rise to this assumed degree of clipping. This may be interpreted as an easing of the calculated conditions, since the interference caused will be less than continuous. Questions of general 'speech factor', language and speaking-distance from the microphone, among others, make it extremely difficult to arrive at a figure for the easement afforded.

Further, due again to speech factor, the distribution of energy within the 300c/s to 3kc/s speech input band will be such that very much more energy occurs towards the low end of this band than at the high end. In this respect, curves (B) and (D) of Fig. 1 will be less frequently achieved than curves (A) and (C), plotted for the same respective maximum frequency deviations. If full modulation at 3kc/s, necessary to produce (B) and (D), is not frequently achieved, then curves (b) and (d) of Figs. 4 and 5, requiring an input level 10dB higher, rather lose significance. Perhaps these latter are valid for only, say, 1 per cent of the total transmission. Under these circumstances it might be more realistic to accept curves (a) and (c) as defining the outermost envelopes of clipped-waveform sidebands, though to do this involves dependence upon the same indeterminate factors referred to above.

It should be pointed out that since the considerations of this article are directed towards the possibility of a transmitter giving rise to sufficient sideband radiation to cause interference with a receiver on an adjacent channel, such interference is most likely to occur between installations operating on a common-frequency simplex basis. However, frequency allocations are commonly made in 'blocks', all headquarters transmitters occupying one portion of the band, and mobiles radiating in a separate portion, several megacycles distant. The possibilities of interference are thus confined to the case of a mobile receiver being in the immediate vicinity of a headquarters transmitter operating on the adjacent channel, and to a limited extent vice versa.

Conclusions

To meet the requirement of restricting occupied bandwidth to a minimum in the case of a mobile-class f.m. v.h.f. transmitter, a suitable arrangement would appear to be as follows:

(a) Speech signals from the microphone circuit pass, via an amplifier if necessary, through a network giving the required response. For example, this response might take the form of a 6dB/octave pre-emphasis, arranged to be complementary to a desired de-emphasis characteristic of an associated receiver.

(b) The signals are now fed, at fairly low impedance, through a symmetrical clipper-type limiter, preferably employing thermionic diodes and arranged to clip at a level of at least a volt or so to ensure a clean action.

(c) Following the limiter is a low-pass filter, having a cut, beyond the speech band upper limit, as steep as economically practicable, but at least that derived in the section on post-limiter filtering.

(d) The output of the filter passes to a phase modulator (with correcting network) or direct frequency modulator, having a linear amplitude characteristic. For either modulating method, within the speech band the whole post-limiter response should be flat; that is, the frequency deviation produced is proportional only to the amplitude of the signal emerging from the clipper, and independent of the frequency.

Acknowledgment

Acknowledgment is made to the engineer-in-chief of Marconi's Wireless Telegraph Co. Ltd for permission to publish this article, also to Mr. E. F. Cranston of that company for setting the author on this line of investigation and for his constructive criticism during the preparation of the article.

Frequency Transformations and Dissipative Effects in Electric Wave Filters

By D. J. H. Maclean*, B.Sc., M.S., A.M.I.E.E.

The object is to obtain a simple method for estimating the effects of parasitic dissipation on the amplitude response of filters. The method is based on the zero-pole pattern of the lossless network, and is thus general in scope. In particular, filters based on the common frequency transformations, e.g. low-pass to band-pass, are considered. To utilize the method, a new approach is made to these frequency transformations. The root pattern of a filter based on a low-pass equivalent is derived by a straightforward process. The resulting pattern is used in a relatively quick graphical estimation of the effects of dissipation. The effects of varying the fractional bandwidth can also be studied.

THE effects of parasitic dissipation on the performance of networks designed to meet stringent requirements are of practical importance. Darlington[1] and Scowen[2] have considered the design of networks having dissipative inductors and capacitors. Mole[3] has given useful information for calculating the effects in image parameter filters. The present method is based on a knowledge of the root positions and so is particularly suited to simple ladder filters where the roots can be calculated easily. Green[4] gives useful information for such filters, and his terminology has been used here.

The method consists of

(1) Computing the root positions of the low-pass equivalent network.

(2) Applying the relevant frequency transformation to the roots.

(3) Using a scale plot of the resulting pattern to estimate the response at points of interest.

The treatment is in terms of explicitly normalized variables to aid readers unfamiliar with the technique, and is mathematically simple.

The physical meaning of the transformations and the mathematical aspects have long been known. However, the author has not seen a thorough treatment, in terms of zeros and poles, for all the transformations. The present article arose from work on narrow bandwidth band-stop filters and the theory developed was found to give reasonable agreement with the experimental results.

The low-pass to band-pass case is treated in detail, but examples are also given for band-stop filters. Fig. 1 illustrates some of the corresponding responses.

The Low-Pass to Band-Pass Transformation

The transformation can be written

$$(\lambda/\omega_\beta) = (1/w)\,[(p/\omega_r) + (\omega_r/p)] \quad \ldots\ldots\ldots (1)$$

where λ denotes the low-pass frequency variable, ω_β is the radian bandwidth, w is the fractional bandwidth, p is the band-pass frequency variable and ω_r is the radian centre frequency; all are defined in the list of symbols. λ and p are both equal to $(\sigma + j\omega)$, two symbols being used only to avoid confusion. The properties of this transformation are well known, but the focal point now is that equation (1) represents a mapping of a point in the low-pass, λ-plane into two points in the band-pass, p-plane. The axes in both planes are real and imaginary frequencies.

The solution for p/ω_r can be found by simple algebra and is

$$(p/\omega_r) = (w/2)\,.\,(\lambda/\omega_\beta) \pm j\,[1 - (w/2\,.\,\lambda/\omega_\beta)^2]^{\frac{1}{2}} \quad \ldots (2)$$

This equation allows one to go from a known value of λ/ω_β to the unknown p/ω_r. For practical purposes, equation (2) will be modified. It is convenient firstly to multiply and divide λ/ω_β by j, then divide through the equation by j, i.e. rotate through $-90°$ so that the positive real frequency axis is horizontal and directed to the right. Equation (2) becomes

$$(p/j\omega_\beta) = (w/2)\,(\lambda/j\omega_\beta) \pm [1 + (w/2\,.\,\lambda/j\omega_\beta)^2]^{\frac{1}{2}} \quad \ldots (3)$$

Before simplifying further, consider equation (3). Several points are immediately obvious:

(a) for $(\lambda/j\omega_\beta) = 0$, $(p/j\omega_r) = \pm 1$

The origin maps into the two band centres. Applied in reverse is ' transferring the carrier frequency to zero ', used in studying amplitude modulation.

(b) for $\lambda = \pm j\infty$, $(p/\omega_r) = \pm j\infty$ or 0

(c) for $(\lambda/j\omega_\beta) = +(\omega/\omega_\beta)$, $(p/j\omega_r) = (w/2)\,.\,(\omega/\omega_\beta) \pm$

$$[1 + (w/2\,.\,\omega/\omega_\beta)^2]^{\frac{1}{2}} \quad \ldots\ldots\ldots (4)$$

From this equation, it is obvious that real frequencies transform into real frequencies.

(d) To obtain the positive frequency points corresponding to the low-pass points $\pm\lambda/j\omega_\beta$, the positive square root is taken.

(e) The band edges are obtained by putting $\lambda/j\omega_\beta = \pm 1$

LIST OF SYMBOLS

ω =	angular frequency, radians per second
ω_β =	angular frequency at band edge (see Fig. 1)
ω_1, ω_2 =	angular frequencies at band edges in band-pass and band-stop cases (see Fig. 1)
ω_r =	geometric mid-band, $= (\omega_1\omega_2)^{1/2}$
w =	fractional bandwidth $= (\omega_2 - \omega_1)/\omega_r$, $\omega_2 > \omega_1$
λ =	complex frequency variable in low-pass case
p =	complex frequency variable in other cases
k =	constant equal to product of low-pass and high-pass normalized frequency variables
k_1 =	constant relating band-pass and low-pass bandwidths $= \omega_\beta/(\omega_2 - \omega_1)$
k_2 =	constant relating low-pass to band-pass or band-stop fractional bandwidth $= \omega_\beta/w$
k_3 =	constant equal to radius of circle in Type B response
V =	output voltage
V_p =	maximum output voltage
V/V_p =	response function
θ, ϕ =	angles
n =	number of branches in a network

* Barr & Stroud Ltd., formerly with the G.E.C. Research Laboratories, Wembley, and Stanford University, California

and, for the positive frequencies in the p-plane

$$p/\mathrm{j}\omega_r = \pm(w/2) + [1 + (w/2)^2]^{\frac{1}{2}} \quad \ldots\ldots\ldots \quad (5)$$

A radical is very inconvenient for numerical calculation since λ is in general complex. Equation (3) can be expanded by the Binomial Theorem with the restriction shown, thus:

$$p/\mathrm{j}\omega_r = (w/2\ .\ \lambda/\mathrm{j}\omega_\beta) \pm [1 + 1/2(w/2\ .\ \lambda/\mathrm{j}\omega_\beta)^2 - 1/8(w/2\ .\ \lambda/\mathrm{j}\omega_\beta)^4 \ \ldots],\ (w/2\ .\ \lambda/\mathrm{j}\omega_\beta)^2 \leqslant 1$$

To illustrate corresponding real frequencies imagine a transparent cylinder around which circles are drawn as shown in Fig. 2. The lines are marked off in frequencies, positive and negative. The direction of increasing frequency from minus infinity to plus infinity is shown by the arrows. Corresponding frequencies lie on the same vertical line. One revolution on the low-pass line corresponds to one on the high-pass line and two for the band-pass and band-stop cases.

TABLE 1

RESPONSE	FREQUENCY VARIABLE	RELATIONSHIP BETWEEN VARIABLES
Low-pass	$\lambda/\mathrm{j}\omega_\beta$	—
High-pass	$\lambda'/\mathrm{j}\omega_\beta$	$\lambda'/\mathrm{j}\omega_\beta = -k/\lambda/\mathrm{j}\omega_\beta$
Band-pass	$p/\mathrm{j}\omega_r$	$p/\mathrm{j}\omega_r = \left(\frac{w}{2}\cdot\frac{\lambda}{\mathrm{j}\omega_\beta}\right) \pm \sqrt{\left[1 + \left(\frac{w}{2}\cdot\frac{\lambda}{\mathrm{j}\omega_\beta}\right)^2\right]}$
Band-stop	$p'/\mathrm{j}\omega_r$	$p'/\mathrm{j}\omega_r = \left(\frac{w}{2}\cdot\frac{\lambda'}{\mathrm{j}\omega_\beta}\right) \pm \sqrt{\left[1 + \left(\frac{w}{2}\cdot\frac{\lambda}{\mathrm{j}\omega_\beta}\right)^2\right]}$

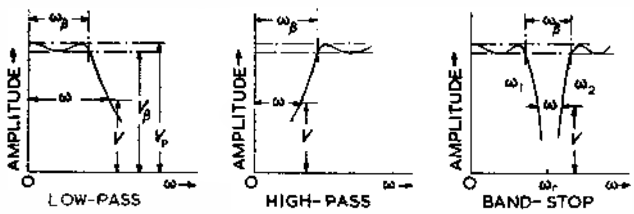


Fig. 1. Typical amplitude responses (after Green)

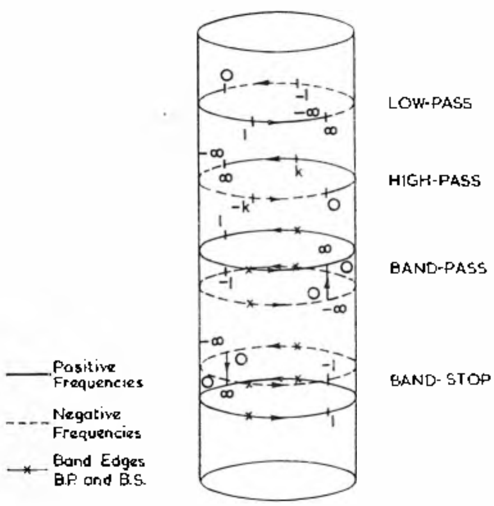


Fig. 2. Illustration of the relationship between real frequencies

or

$$p/\mathrm{j}\omega_r = \pm 1 + (w/2\ .\ \lambda/\mathrm{j}\omega_\beta) \pm [1/2(w/2\ .\ \lambda/\mathrm{j}\omega_\beta)^2 - 1/8\,(w/2\ .\ \lambda/\mathrm{j}\omega_\beta)^4 - \ldots] \quad \ldots\ldots\ldots\ldots \quad (6)$$

It should be noted that the original relation, equation (1) is valid for any finite w, the filter need not be narrow-band. For computation, however, series (6) should converge rapidly; this is usually so in practical filters, where w is often less than 30 per cent.

The relationship in the case of the high-pass and band-stop transformations can be obtained similarly and are given in Table 1. Fig. 1 illustrates typical amplitude responses.

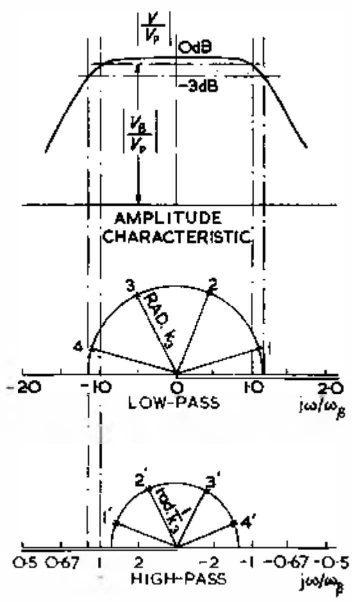


Fig. 3. Correspondence between low-pass and high-pass pole positions. Type B response, $n = 4$. *Underlined portions of the frequency axis positive frequencies*

To go from low-pass to high-pass, it is helpful to regard the variables as vectors of product $-k$ (see Table 1). The meaning of the minus sign is easily seen from Fig. 3 to be an interchange of the positive and negative halves of the complete frequency axes. The high-pass points are then found by inverting the low-pass points with respect to k (see Fig. 3). This is an example of the useful device known as electrical inversion.

The Effects of Dissipation

Providing the dissipation is small, it can be well represented in practice by the mean decrement

$$\Delta = 1/2\ (R/L + G/C) \quad \ldots\ldots\ldots\ldots \quad (7)$$

where the ratios are the average ratios for all reactances of the appropriate kind. In normalized terms,

$$(\Delta/\omega_\beta) = 1/2\ (R/\omega_\beta L + G/\omega_\beta C) = 1/2\ (1/Q_L + 1/Q_C) \quad \ldots\ldots\ldots \quad (8)$$

the Q's are seen to be those measured at the band edge.

Equation (6) can be modified to include the effects of parasitic dissipation by a suitable change of the low-pass frequency variable. This change simply replaces $\lambda/j\omega_\beta$ by $(\lambda + \Delta)/j\omega_\beta$, hence

$$p_d/j\omega_m = \pm 1 + (w/2)\left(\frac{\lambda + \Delta}{j\omega_\beta}\right) \pm \left[1/2\left\{w/2\frac{\lambda + \Delta}{j\omega_\beta}\right\}^2 - \ldots\right] \quad \ldots\ldots\ldots\ldots (9)$$

$$= [\pm 1 + (w/2 \,.\, \lambda/j\omega_\beta) \pm 1/2(w/2 \,.\, \lambda/j\omega_\beta)^2 \pm 1/8 \, (w/2 \,.\, \lambda j\omega_\beta)^4 \ldots] + [w/2 \,.\, \Delta/j\omega_\beta) \pm (w/2 \,.\, \lambda/j\omega_\beta) \,.\, (w/2 \,.\, \Delta/j\omega_\beta) \ldots] \quad \ldots\ldots\ldots\ldots (10)$$

$$= (p/j\omega_r) + (d/j\omega_r) \quad \ldots\ldots\ldots\ldots (11)$$

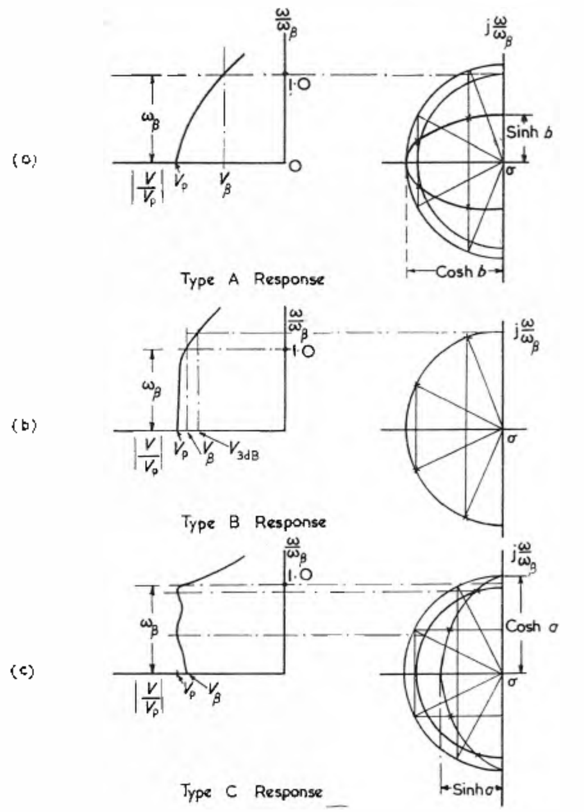


Fig. 4. Amplitude characteristics and pole patterns of low-pass ladder configurations (after Green)

One sees that just as in the low-pass case, the addition of dissipation amounts to a change of the complex frequency variable.

Examining the dissipation term more closely, one notes that:

(a) The first approximation for d/ω_r is $w/2 \,.\, \Delta/j\omega_\beta$. This can conveniently be expressed in terms of the mid-band Q's as follows:

$$w/2 \,.\, (\Delta/j\omega_\beta) = \frac{\omega_2 - \omega_1}{j\omega_r} \,.\, \tfrac{1}{4}(R/\omega_\beta L + G/\omega_\beta C)$$

$(\Delta/j\omega_r) = \tfrac{1}{2}(1/Q_L + 1/Q_C)$, provided that $(\omega_2 - \omega_1) = \omega_\beta$.

The Q's are the mid-band values.

(b) If the first approximation is sufficient, and if Δ can be taken as constant within the pass-band, the effect is a displacement of the roots to the left. This can conveniently be represented by taking an auxiliary $j\omega$-axis to the right of the original axis, and evaluating the response function along this new axis.

(c) The second approximation is to include the product term; it should be noted that this is a vector having the angle of the particular value of $\lambda/j\omega_\beta$ considered.

The above treatment stems from the work of Bode[5] and Boothroyd[6].

Use of the Root Pattern as a Vector Diagram

The aid of a diagram of the poles and zeros of a network function (impedance, admittance, transmission properties, for instance) can be enlightening because the driving point and transfer immittances are defined to within a constant multiplier by their roots in the complex plane[7]. This fact is very important for the present purpose. One may represent the response function, defined as V_p/V (i.e. the insertion voltage ratio) in the form

$$(V_p/V) = K\frac{\pi_r(p - p_r)}{\pi_s(p - p_s)} \quad \ldots\ldots\ldots\ldots (12)$$

For restrictions on K, and on the roots, the reader is referred to works on network theory, e.g. Darlington[1], Bode[3]. This is a quotient of products of factors, and the roots p_r, p_s may be real, imaginary, or complex. If not real, the roots must occur in conjugate pairs. The roots of equation (12) can be plotted to a suitable scale, and each term in equation (12) represents a vector with its tip at some value

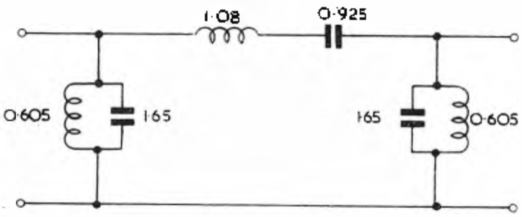


Fig. 5. Simple band-pass filter, normalized element values

of p, and its tail at the root. The magnitude of V_p/V at real frequencies ($p = j\omega$) is of primary concern here, consequently one considers the magnitude of equation (12). For the right-hand side, this is the product of the magnitudes of the individual factors. As $p = j\omega$ is varied along the real frequency axis, the vectors extend or contract so the behaviour of V_p/V can be estimated. Fig. 4 shows the amplitude characteristics and pole patterns of low-pass ladder configurations.

The reason for developing equations (6) or (10) should now be clear, since through them the transformed pattern can be obtained, and the above graphical method used. The following section has some examples to illustrate the technique.

It may be noted that the graphical method is applicable to any vector diagram of the roots of a network function, when suitably interpreted, not just to the response function considered above. Neither is the method confined to filters, but applies to finite, lumped networks in general.

Applications to Filters

The first example is a simple, three-branch band-pass filter (Fig. 5) designed to the specification in Table 2.

The first step is to compute the derived quantities; these are shown in Table 2. The next step is to find the root positions in the low-pass case. These are found by standard formulæ (see, for example, Green[4] pp. 18-20) and appear in Table 3. The roots are then expressed in terms of $\lambda/j\omega_\beta$. Following this, the various transformed root positions are found as shown in Table 3, where it may be noted that polar co-ordinates are used. The third order term of equation (6) is negligible, even for this wide-band case.

The first order band-pass and the second order approximation are shown in Fig. 6. The contour shown is obtained by transforming a sufficient number of points from the low-pass elliptical locus. It can be seen that the contour has become egg-shaped; this is characteristic for type-C response. There is, of course, a mirror image in the real axis centred about $p/j\omega_r = -1$. By the properties of the transformation, there are also third order roots at the origin and at infinity.

The amplitude response for the first order approximation has been computed and is shown in Fig. 7. It is seen to be the same shape as the low-pass response, but centred on $+1$. This response ignores the effect of the roots at the origin and at negative frequencies. The exact amplitude response has also been computed and is shown on the same figure. This response will agree closely with that found by taking account of the second order approximation and using the graphical method of a scale and all the finite plane roots, i.e. six poles in two clusters and a third order zero at the origin.

It can be observed from Fig. 6 that the pole position below band centre is much closer to the $j\omega$-axis; this immediately suggests that the effects of dissipation will be greater there. Quantitatively, one requires the band edge points, which from equation (5) are

(a) upper band edge at $p/j\omega_r = 1{\cdot}414$

(b) lower band edge at $p/j\omega_r = 0{\cdot}707$

and the first order dissipation term in equation (10). This displacement of ·003 units to the left can conveniently be made by taking an auxiliary axis as mentioned previously and leaving the roots undisturbed. With a fine scale, e.g. a millimetre scale, and a suitable scale plot, the ratio of the distances from a frequency on the auxiliary and original axes to a given root is found. This is repeated for each root, and the product of the ratios taken. For this case, to obtain a result to within a few per cent, the effects of the zeros at the origin and the pole cluster about -1 are negligible. As an example of the ratios for the lower band edge, the author obtained

$$\frac{1{\cdot}17}{1{\cdot}10} \times \frac{7{\cdot}25}{7{\cdot}20} \times \frac{1{\cdot}492}{1{\cdot}490} = 1{\cdot}071$$

TABLE 2

SPECIFICATION	DERIVED QUANTITIES
Allowable pass-band ripple = 0·5dB	$\omega_r = (\omega_1\omega_2)^{1/2}$
$\omega_\beta = 2\pi \times 10^6$ rad/sec	$= 2 \times 2\pi \times 10^6$ rad/sec
$\omega_2 = 4\pi \times 10^6$ rad/sec	$w = \dfrac{\omega_2 - \omega_1}{\omega_r}$
$\omega_1 = 2\pi \times 10^6$ rad/sec	$= \dfrac{1}{\sqrt{2}} = 0{\cdot}707$
$Q_L = 200$	$\dfrac{\triangle}{j\omega_r} = \frac{1}{2}\left(\dfrac{1}{200} + \dfrac{1}{1000}\right)$
$Q_c = 1000$	$= 0{\cdot}003$

The results are tabulated in Table 4, together with the computed ratios of the exact response for comparison. The computed response of the high-loss network is shown on Fig. 7; the computations are in the Appendix.

The accuracy required in the above graphical construction and in the computation depends upon the accuracy required in the final result. The examples here were worked out to about 0·3 per cent accuracy using good millimetre

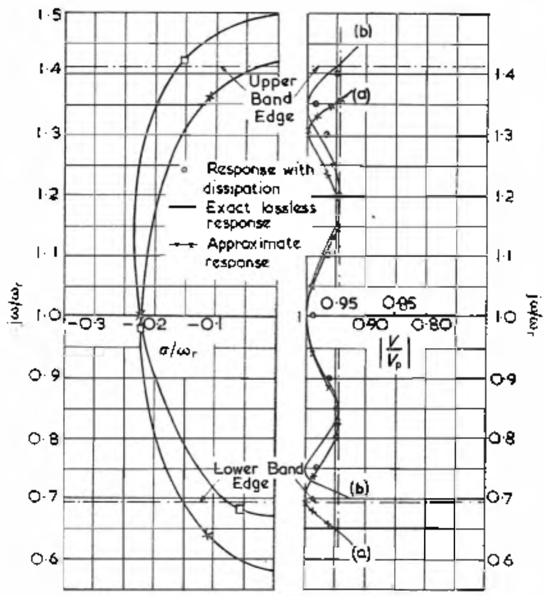


Fig. 6 (left). Pole locations for simple band-pass filter
Fig. 7 (right). Amplitude response for simple band-pass filter

graph paper and a slide-rule. The degree of accuracy could be improved if required by not using graph paper in the plot.

The next example is a band-stop filter. The specification and computation are as follows:

Specification: Type-B response
Band centre, 1·002Mc/s
Fractional bandwidth, $w = 0{\cdot}05$
Number of branches, $n = 5$
Bandwidth defined by 1dB points
Maximum attenuation > 100dB

TABLE 4

FREQUENCY	COMPUTED	GRAPHICAL
Lower band edge	1·073	1·071
Upper band edge	1·017	1·032
Mid-band	1·010	1·023

TABLE 3

ROOTS IN THE LOW-PASS PLANE		COMPONENTS IN THE BAND-PASS PLANE	
$\lambda/j\omega_\beta$	$\lambda/j\omega_\beta$	FIRST ORDER	SECOND ORDER
$\dfrac{\lambda_{1,3}}{\omega_\beta}$	$\dfrac{\lambda_{1,3}}{j\omega_\beta}$	$\dfrac{w}{2} \cdot \dfrac{\lambda_{1,3}}{j\omega_\beta}$	$1/2\left(\dfrac{w}{2} \cdot \dfrac{\lambda_{1,3}}{j\omega_\beta}\right)^2$
$-0{\cdot}3145 \pm j\ 1{\cdot}022$	$\pm 1{\cdot}022 + j0{\cdot}3145$	$0{\cdot}377 \angle 17{\cdot}1°$ or $162{\cdot}9°$	$0{\cdot}072 \angle 34{\cdot}2°$ or $325{\cdot}8°$
$\dfrac{\lambda_2}{\omega_\beta}$	$\dfrac{\lambda_2}{j\omega_\beta}$	$\dfrac{w}{2} \cdot \dfrac{\lambda_2}{j\omega_\beta}$	$1/2\left(\dfrac{w}{2} \cdot \dfrac{\lambda_2}{j\omega_\beta}\right)^2$
$-0{\cdot}629$	$+ j0{\cdot}629$	$0{\cdot}222 \angle 90°$	$0{\cdot}025 \angle 180°$

The computation, using Green's symbols, follows.

The available decrement is $\Delta/\omega_1 = 1{\cdot}39 \times 10^{-3}$.

(a) Calculation of radius of circular locus, k_3;

by definition, $k_3 = \left\{|V_p/V_\beta|^2 - 1\right\}^{-(1/10)} = 1{\cdot}144$ for 1dB points.

(b) Calculation of low-pass pole positions on the locus; from Green, angles relative to real frequency axis are

$$\phi_q = \frac{2q-1}{n} \cdot (\pi/2),\ q = 1, 2, 3, 4, 5$$

a constant, which is given by calculation (c) above. The results are shown in Table 6 and the measured insertion loss characteristic is shown in Fig. 9.

If the mean value of the two band edges is taken, that is, making the edges symmetrical, the difference between this mean value and 2dB is 0·1dB.

An example starting from an elliptical low-pass distribution is the band-stop filter specified in Table 7.

The values for the low-pass positions were obtained from Green[3], the high-pass positions by inversion with respect to unity. To obtain the band-stop positions, the equation

TABLE 5

FREQUENCY PLANE	LOCUS RADIUS	ANGULAR POSITION IN DEGREES				
		ϕ_1	ϕ_2	ϕ_3	ϕ_4	ϕ_5
Low-pass	1·144	18	54	90	126	162
High-pass	0·873	162	126	90	54	18
Band-stop First term	$\frac{w}{2} \cdot \frac{1}{k_3} = 21.83 \times 10^{-2}$	162	126	90	54	18
Second term	$\frac{w^2}{8} \cdot \frac{1}{k^3} = 0{\cdot}283 \times 10^{-3}$	−216	−288	−360	−72	−144

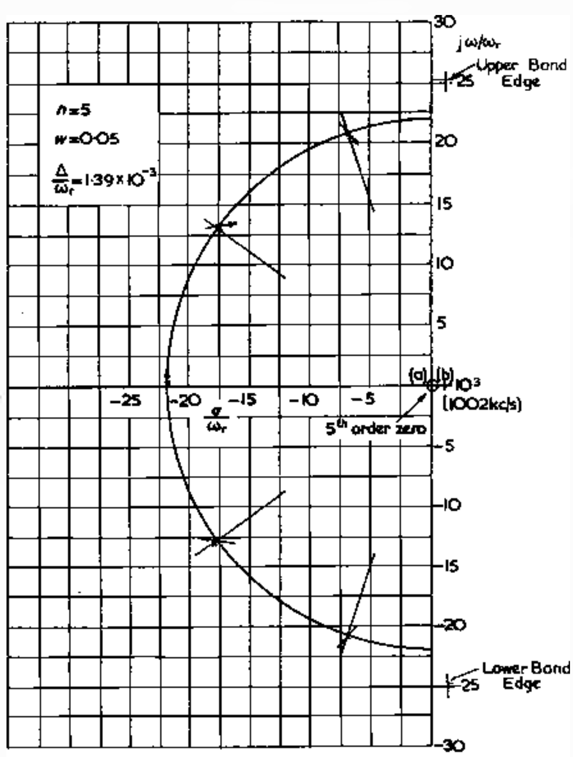


Fig. 8. Root pattern for band-stop filter with Type B response

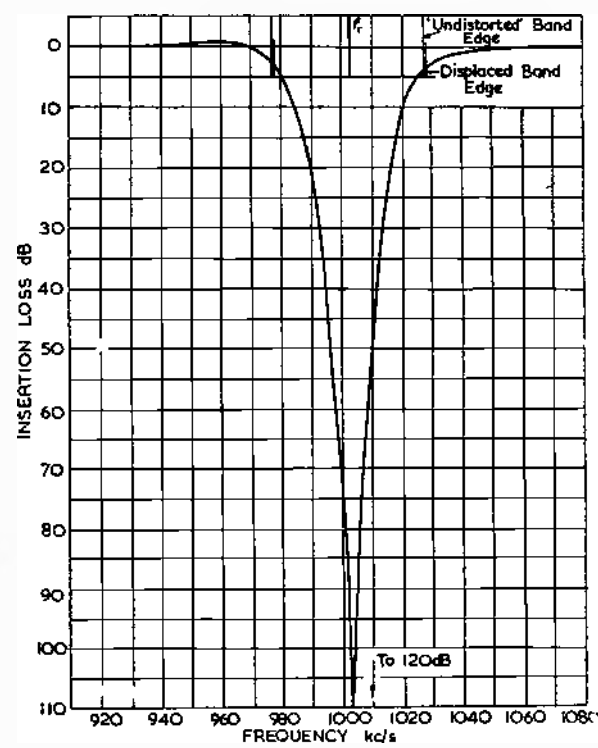


Fig. 9. Measured insertion loss characteristic of band-stop filter. Type B response

(c) Calculation of the constant multiplier of V_p/V, where V_p/V is given by

$$V_p/V = \left\{|V_p/V_\beta|^2 - 1\right\}^{-\frac{1}{2}} \times \text{product of five factors.}$$

The computation for the transformation is given in Table 5.

The resulting root pattern is shown in Fig. 8. By the scale diagram method the effect of finite Q is estimated at the two band edges. Estimating the maximum insertion loss can be done as follows: the numerator of the response is proportional to the distance from (b) to the zeros, counted according to their multiplicity. The denominator is proportional to the distance from (b) to the poles, which are single. The ratio gives the response of the lossy network to within

used was

$$p'/j\omega_1 = \pm 1 + (w/2 \,.\, \lambda'/\omega_\beta) \pm$$
$$[\tfrac{1}{2}(w/2 \,.\, \lambda'/j\omega_\beta)^2 - \tfrac{1}{8}(w/2 \,.\, \lambda'/j\omega_\beta)^4 - \ldots]$$

The results obtained are given in Table 8 and the resulting pattern for the roots is shown in Fig. 10.

The dissipative response is obtained as before. In this

TABLE 6

FREQUENCY	MEASURED	GRAPHICAL
Lower edge	1·7dB	2·1dB
Upper edge	2·5	1·9
Mid-band	>120	0·129

TABLE 7

FILTER SPECIFICATION	POLE POSITIONS					
	LOW-PASS ($\lambda_m/j\omega_\beta$)			HIGH-PASS ($\lambda'_m/j\omega_\beta$)		
Type—C response	m	modulus	argument	m	modulus	argument
$n = 5$ $w = 0{\cdot}05$	1	1·024	6·3°	1	0·9755	173·7°
ω_β defined by 0·5dB points	2	0·6915	28	2	1·448	152
centre frequency 500kc/s	3	0·364	90	3	2·848	90
attenuation>90dB for 550 ± 2kc/s	4	0·6915	152	4	1·448	28
$\Delta/\omega_r = 1{\cdot}14 \times 10^{-3}$	5	1·024	173·7	5	0·9755	6·3

TABLE 8

BAND-STOP ROOT POSITIONS OF V/Vp				
ZEROS	FIRST ORDER POLES		SECOND ORDER POLES	
	MODULUS	ARGUMENT	MODULUS	ARGUMENT
Fifth order zeros at positive and negative band centres	$2{\cdot}44 \times 10^{-2}$	173·7°	$2{\cdot}98 \times 10^{-4}$	347·4°
	3·62 ,,	152	6·54 ,,	304
	7·12 ,,	90	25·4 ,,	180
	3·62 ,,	28	6·54 ,,	56
	2·44 ,,	6·3	2·98 ,,	12·6

TABLE 9

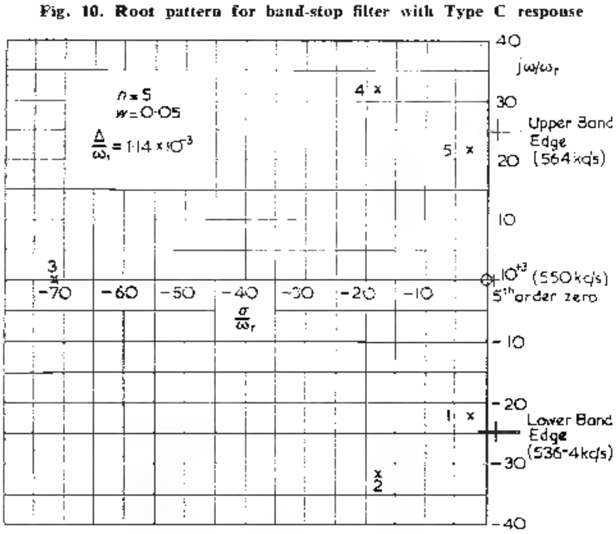
FREQUENCY	MEASURED	GRAPHICAL
Lower edge	2·6dB	2·5dB
Upper edge	4·0	2·5
Mid-band	>120	129

case, the product term is negligible so only the first approximation is required. The results are given in Table 9 and the measured insertion loss characteristic is shown in Fig. 11.

The agreement at the upper band edge is poor; however, no attempt was made to make the filter response symmetrical; if this had been tried the agreement would probably have been improved.

As a final example, consider a low-pass filter which was designed by the Darlington insertion loss method to the specification below. As the actual design of the filter is

Fig. 10. Root pattern for band-stop filter with Type C response



not strictly relevant, the details have been omitted, but the resulting normalized filter is shown in Fig. 12. The filter response for a Q of 100 is given in Table 10.

Specification: Pass-band $0 < \omega < 1$

pass-band ripple 0·4dB

frequencies of infinite loss $\omega = 1{\cdot}25, \infty$

The graphical response was computed exactly as in the

Fig. 11. Measured insertion loss characteristics of band-stop filter, Type C response

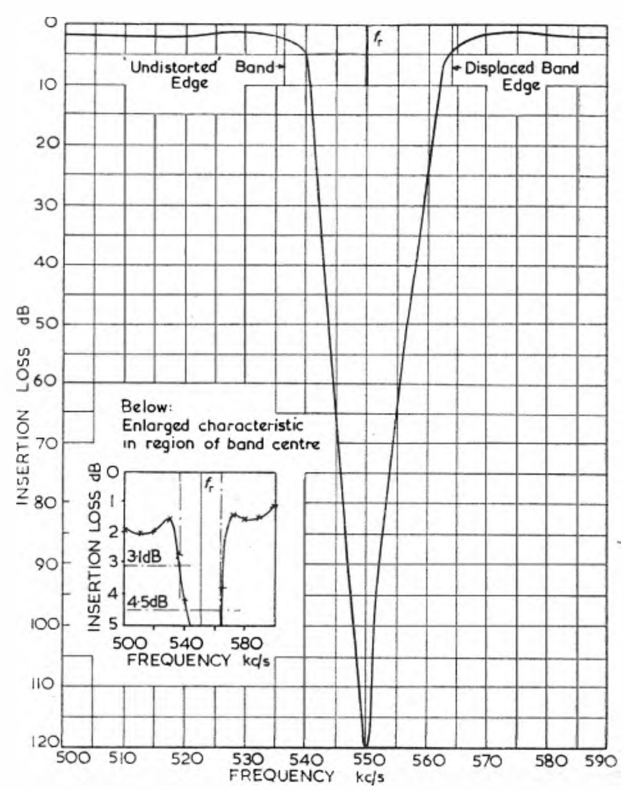


TABLE 10

ω	INSERTION LOSS (dB)									
	0	0·2	0·4	0·6	0·8	1·0	1·05	1·2	1·25	1·3
Ideal	0	0·09	0·16	0·41	0·09	0·41	1·45	17·1	∞	20.5
Computed $Q = 100$ . .	0·06	0·16	0·35	0·50	0·35	0·50	1·52	16·8	32·6	20·1
Graphical $Q = 100$ . .	0·10	0·19	0·26	0·53	0·24	0·51	1·95	17·6	34·6	—

previous examples except for the frequency of zero transmission. The response there was established as follows: the ratio V_p/V was scaled from the diagram for several values near $\omega = 1{\cdot}25$. This ratio was compared with the computed ratio to give a scale factor (it is not constant) and the estimated value at $\omega = 1{\cdot}25$ used to find the insertion loss with $Q = 100$. As will be seen from Table 10, the agreement is good.

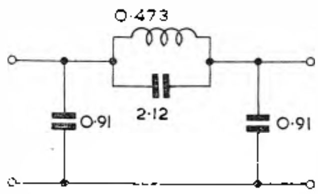


Fig. 12. Low-pass filter, Darlington insertion loss design

Conclusions

The agreement between computed and graphical results is satisfactory. In the case of the measured filter responses, the graphical results do not agree so well; this is to be expected, since component tolerances and alignment procedures contribute to a response which differs from the ideal. However, the agreement is reasonable.

The method of series expansion for the band-pass and band-stop cases is believed to be new and is of considerable utility. Using the transformed root positions, the graphical estimation can be done simply and rapidly.

It would appear that the application of the method in the preliminary design stage of a network, using the poles and zeros of the network function, might be of value as an indication of the effects to be expected from parasitic dissipation.

Acknowledgments

The author wishes to acknowledge the assistance given by colleagues at the General Electric Company's Research Laboratories, Wembley, and at Stanford University, California, in helpful discussions and criticisms.

Appendix

Formulæ required for the example of the three-branch band-pass filter specified in Table 2.

From Green[4] page 20. equation (39), one can write

$$V_p/V = 2^{3-1}\left\{|V_p/V_\beta|^2 - 1\right\}^{1/2}[(\lambda/\omega_\beta) + 0{\cdot}629]\,[(\lambda/\omega_\beta)^2 + 0{\cdot}629\,(\lambda/\omega_\beta) + 1{\cdot}1429] \quad \ldots\ldots (13)$$

and for 0·5dB ripple

$$\left\{|V_p/V_\beta|^2 - 1\right\}^{\frac{1}{2}} = [(1{\cdot}059)^2 - 1]^{\frac{1}{2}} = 0{\cdot}3464$$

The multiplier is thus $4 \times 0{\cdot}3464 = 1{\cdot}3856$.

The square of the magnitude of V_p/V can be written

$$|V_p/V|^2 = \tfrac{1}{2}[|V_p/V_\beta|^2 + 1] + \tfrac{1}{2}[|V_p/V_\beta|^2 - 1]\,T_{2n}(\lambda/\omega_\beta) \quad \ldots(14)$$

where T_{2n} is a Chebyshev polynomial of order $2n$. For $|\lambda/j\omega_\beta| < 1$,

$$T_{2n}(\lambda/j\omega_\beta) = \cos[2n\cos^{-1}(\lambda/j\omega_\beta)],$$

thus for the case in hand

$$(V_p/V)^2 = 1{\cdot}06 + 0{\cdot}6\cos[6\cos^{-1}(\lambda/j\omega_\beta)] \quad \ldots\ldots (15)$$

This equation was used for computation. For the band-pass first approximation, the values of equation (15) were assigned to band-pass frequencies related to the low-pass frequencies by

$$(\omega/j\omega_r) = 1 \pm (w/2)\,.\,(\omega/j\omega_\beta)$$

To obtain the exact ideal band-pass response, one uses the fact that the transformation amounts to a distortion of the frequency scale (see Fig. 2). The responses at correctly related frequencies are the same.

For the calculation of the exact response with the first approximation to the dissipation, one can use the band-pass variable below, and relate the frequencies as before;

$$1/w\left(\frac{p+d}{\omega_r} + \frac{\omega_r}{p-d}\right) \quad \ldots\ldots\ldots\ldots (16)$$

For real frequencies, equation (16) becomes

$$1/w\left(\frac{d+j\omega}{\omega_r} + \frac{\omega_r}{d+j\omega_r}\right)$$

$$= 1/w\left[\frac{d+j\omega}{\omega_r} + \frac{\omega_r(d-j\omega)}{d^2+\omega^2}\right]$$

$$= 1/w\left[\frac{d+j\omega}{\omega_r} + \frac{\omega_r(d-j\omega)}{\omega^2}\right], \text{ assuming } d^2 \ll \omega^2$$

$$= 1/w\ [d/\omega_r(1+(\omega_r^2/\omega^2)) + j(\omega/\omega_r - \omega_r/\omega)]$$

$$= 1/w\ [d/\omega_r\,.\,\omega_r/\omega(\omega/\omega_r - \omega_r/\omega) + j(\omega/\omega_r - \omega_r/\omega)] \quad \ldots\ldots\ldots\ldots (17)$$

For $d = 0$, equation (17) reduces to the known form. Denoting by Ω the frequency variable of equation (17), the band-pass case can be written down from equation (13) as

$$|V_p/V| = 1{\cdot}3856\,|\Omega + 0{\cdot}629|\,.\,|\Omega^2 + 0{\cdot}629\Omega + 1{\cdot}1429| \quad \ldots\ldots\ldots\ldots (18)$$

The response indicated by the circles in Fig. 7 was obtained by calculation from equation (18) with the values

$$(1/w) = 1{\cdot}414,\ (d/\omega_r) = 3 \times 10^{-3}$$

REFERENCES

1. Darlington, S. Synthesis of Reactance 4-Poles. *J. Math. Phys. 18*, 257 (1939).
2. Scowen, F. An Introduction to the Theory and Design of Electric Wave Filters. (Chapman & Hall, 1950.)
3. Mole, J. H. Filter Design Data for Communication Engineers. (E. & F. N. Spon Ltd., 1952.)
4. Green, E. Amplitude-Frequency Characteristics of Ladder Networks. (Marconi's Wireless Telegraph Co., 1954.)
5. Bode, H. W. Network Analysis and Feedback Amplifier Design. Ch. 10, Section 2 (D. Van Nostrand, Inc.)
6. Boothroyd, A. R. Design of Electric Wave Filters with the Aid of the Electrolytic Tank. *Proc. Instn. Elect. Engrs. 98*, Pt. 4, 65 (1951).
7. Bode, H. W. *loc. cit.* 25.
8. Bode, H. W. *loc. cit.* 209.

Heat Control in Electronic Equipment

(Part 3)

By E. N. Shaw*, A.M.I.E.E.

Control of Temperature Differentials

Ventilation having produced the initial precipitous temperature reductions shown in the histogram, Fig. 9, the final series of experiments, recorded on Plates 7 and 8, aimed at consolidating these 'brute force' gains by a closer approach to the concept of a thermally neutral dust cover. This required a much more subtle application of the means of heat transfer by radiation and conduction, with the triple purpose of lowering the vertical temperature differential, raising horizontal differential and attaining further temperature reductions by more effective control of heat flow into, and through, the front panel. The successful transposition of the differentials is recorded at the top of the histogram, accompanied by temperature reductions indicating progressive improvements in total dissipation.

Borrowing first from the interim solution, an inclined plate extending from the aperture top to the deck was attached to the front panel (Exp. 40) to deflect heat-laden air before it reached the top of the dust cover which, for the first time, became cooler than the adjacent front panel. Having a clean, low-absorptivity, surface, the deflector also acted as a reflector of radiated heat, as shown by the outward displacement of the 10° isotherm in the lower part of the external pattern. A third function was added by attaching the deflector-reflector plate to the deck (Exp. 41) to form a conducting path between the heat source and the front panel.

Continuing this process below the heat source, an inclined plate was attached to the base of the aperture (Exp. 42), its action as a reflector of radiated heat being shown by the outward displacement of the 10° isotherm in the upper part of the external pattern, with the louvres permitting the cold air from the inlet to be drawn up past the valves. Attaching the plate to the deck (Exp. 43) provided a conducting path to the front panel unimpaired by the louvres, which were placed along the line of flow instead of transverse to it, as in the instrument case.

Having eliminated 'aimless' conduction by a double path to the front panel, the flow along both paths was encouraged by giving the panel a high-emissivity surface (Exp. 44), with a similar treatment of the hot, but very restricted, surface of the deck (Exp. 45) to enable it to radiate directly into free space. No sensational temperature reductions were expected because the treated areas were a small portion of the whole. Nevertheless, the reductions obtained were not accompanied by a single rise, showing that the heat was, in fact, dissipated without transfer to other structural members.

Pausing for a moment at this stage of development, it should be pointed out that this form of construction could have been arrived at with equal logic by a different approach. As advocated earlier, the heat source could have been mounted directly on the front panel with resistors inside the container and valves outside. To avoid the possibility of physical damage to the valves and to withdraw them from the convection stream of a unit below, the panel could have been recessed, with the sloping sides of the recess kept clean to reflect radiation from the valves, which could have been cooled by an individual supply of cold air from the rear of the container passing through louvres in the lower slope of the recess. While results would have been identical, the chosen method of exposition was considered superior because it retained the heat source at the centre of the container, where its potentialities as a thermal menace had been demonstrated in earlier experiments, with the stage-by-stage simulation of the recessed panel countering this menace by a process of progressive segregation. With convection deflected, radiation reflected and conducted heat dissipated, segregation accomplished the virtual withdrawal of the valves from the centre of the container.

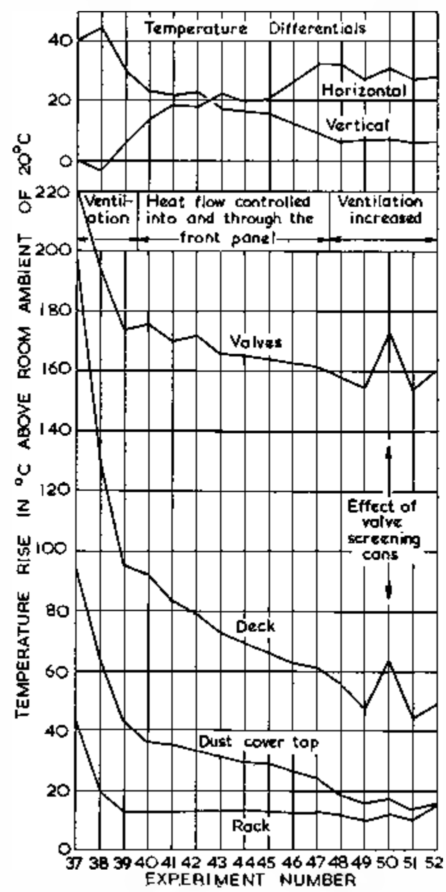


Fig. 9. Temperature reductions and transposition of temperature differentials during evolution of the final unit

With the valves effectively segregated, attention was focused on the 30 watts dissipated inside the container by the load resistors. Although concern was no longer felt for their operating conditions, the top of the dust cover still frustrated their convection stream, some of which escaped through the upper vent, while the rest turned anti-clockwise in the familiar internal convection pattern to resist the upward flow of cooling air, the majority of which moved horizontally between the inlet vent and the louvres to cool the valves. The existence of this main ventilating flow suggested a different method of dealing with the internal

* *Radar Research Establishment, Malvern.*

convection stream responsible for the 16° to 18° differential in Exp. 45. Instead of attempting to deflect it, the resistors were transferred—at first tentatively (Exp. 46) and then completely (Exp. 47)—to new positions below the louvres where their heat loss by convection was added to the main flow, as shown by the outward displacement of the 20° isotherm in the lower portion of the recess. This move not only reduced the vertical differential by a alf to 8° to 9° in Exp. 47, but also increased the horizontal differential by greater heat transfer to the lower part of the front panel.

Final Cooling by Ventilation

Having attained the desired transposition of differentials with progressive improvements in total dissipation, the final move was designed to obtain temperature reductions throughout the container by more vigorous cooling of the heat source itself. With valves and resistors ventilated by the same stream, admitting a greater quantity of cold air through an enlarged inlet vent in the rear of the dust cover (Exp. 48) had the expected effect, all temperatures being reduced and the differential dropping to a mere 4° to 6°. The one fault with this arrangement was that some air detached itself from the carefully cultivated horizontal flow to travel up and out via the exhaust vent. This was countered by increasing the size of the inlet vent until the rear of the dust cover had been cut away to within 2in of the top (Plate 8, Exp. 49) enabling a large mass of cold air to be drawn into the interior with little departure from the horizontal until it had done the useful work indicated by further temperature reductions.

With so much reliance placed on ventilation it becomes important to avoid impairing convection loss by illogical actions, such as placing screening cans on the valves. Their effect is seen in the regressive Exp. 50, with all temperatures increased because cooling by convection had been reduced, more heat from the valves conducted to the deck and less dissipated by reflected radiation, as shown by the virtual collapse of the 'radiation bulge' forward of the panel aperture.

Removal of the screening cans drew attention to the valveholder skirts as possible barriers to convection cooling of the first $\frac{3}{8}$in of valve length. Cutting away the skirts (Exp. 51) improved valve cooling and reduced container temperatures by quite appreciable amounts. It was noted that the initial target had been reached, both valves and deck being cooler than in free air.

A halt was called at this stage because it was felt that further minor improvements were a matter of detail and that the important thing was to put the system itself to the acid test of sandwiching the unit of Exp. 51 between others in a rack assembly to see how it improved upon the interim solution.

Behaviour of the Final Unit in a Rack Assembly

Two units identical with that in Exp. 51 were sandwiched between a 'cold' unit above and a special unit (with deflector plate as in the interim solution) below, the assembly being shown in Exp. 52. Lacking any absolute standards, the success of the final design was judged by comparison with the interim solution of Exp. 36.

(1) The 'cold' unit of Exp. 36 grew warmer by 7° due to a total of 356W below it in two units needing deflector plates in inter-unit spaces, while the smaller rise of 5° was attained in Exp. 52 without increased spacing or deflectors, and with a further unit added for good measure to make a total of over $\frac{1}{2}$kW below the 'cold' unit.

(2) Rack temperatures in Exp. 52 were lower than those in Exp. 36 despite 50 per cent greater loading, indicating more effective heat dissipation by each unit.

(3) Valve and deck temperatures in the singly-mounted unit of Exp. 34 rose by 14° and 15° respectively when the unit was sandwiched in Exp. 36. Sandwiching the singly-mounted unit of Exp. 51 increased valve and deck temperatures by only 8° and 4° respectively in the upper unit of Exp. 52, with the possibility of even smaller rises if the lower of the sandwiched units had not become hotter than the upper one due to 'heat exchanger' transfer from the special unit below.

(4) While valve temperatures were not as low as in the interim solution, their segregation in the final unit enormously increased the area within the dust cover with a temperature rise below 20°.

These comparisons show that building deflectors into the final units avoided 'chimney effect' at least as successfully as in the interim solution, and that segregation of the heat source enabled the concept of a thermally neutral dust cover to be approached sufficiently closely to permit the direct sandwiching of properly designed units without fear of appreciable 'heat exchanger' action.

This does not imply that Exp. 51 represents the ultimate in heat dissipation. It might be worth while, for instance, to increase conduction to the front panel by using thicker material for both deflector and reflector plates; to polish their outer surfaces to improve reflectivity and their inner surfaces to reduce emissivity within the container; to have louvre apertures wider than the $\frac{1}{8}$in used in the experiments, and to shorten the conducting path to the front panel and make the deflector angle less acute by using louvres $2\frac{1}{2}$in long instead of 3in (the shortest available). It is probable that wider louvres would be of the greatest benefit in lowering the impedance to the flow of ventilating air.

An interesting point arising out of the concept of a thermally neutral dust cover is that the closer the approach to neutrality, the less need there is for an enclosure of any kind. For example, it is very hard to justify the presence of two horizontal surfaces between the upper and lower sandwiched units in Exp. 52. Removing them would merely allow air from the lower unit to escape through the louvres in the upper unit instead of emerging from the exhaust vent, which would become unnecessary. This might have the appearance of encouraging 'chimney effect' but, with the main convection flow deflected into an external 'chimney', the residue is harmless and some benefit might be derived from removing the last vestige of frustration due to horizontal dust cover surfaces.

Final units may therefore be used in cubicle assemblies without dust covers or exhaust vents if provision is made for removing residual air from the topmost unit through vents in the cubicle top. No other precautions are necessary if the units are mounted on the cubicle front, but a completely enclosed cubicle must provide an external 'chimney' by mounting the units at least 5 to 6in behind the cubicle doors, which should be painted matt black to absorb radiated energy without reflecting it back to the units. Inlet vents in either case should be provided from top to bottom in the rear of the cubicle to make the entire assembly into a large-scale version of a single unit.

Providing an adequate supply of air is available in the room containing the equipment, each unit or assembly of units will 'inhale' horizontally as much cool air as it requires for natural convection cooling and 'exhale' heated air in a vertical stream easily deflected to an exhaust vent in the shape of a window, the analogy being completed by an open door acting as an inlet vent.

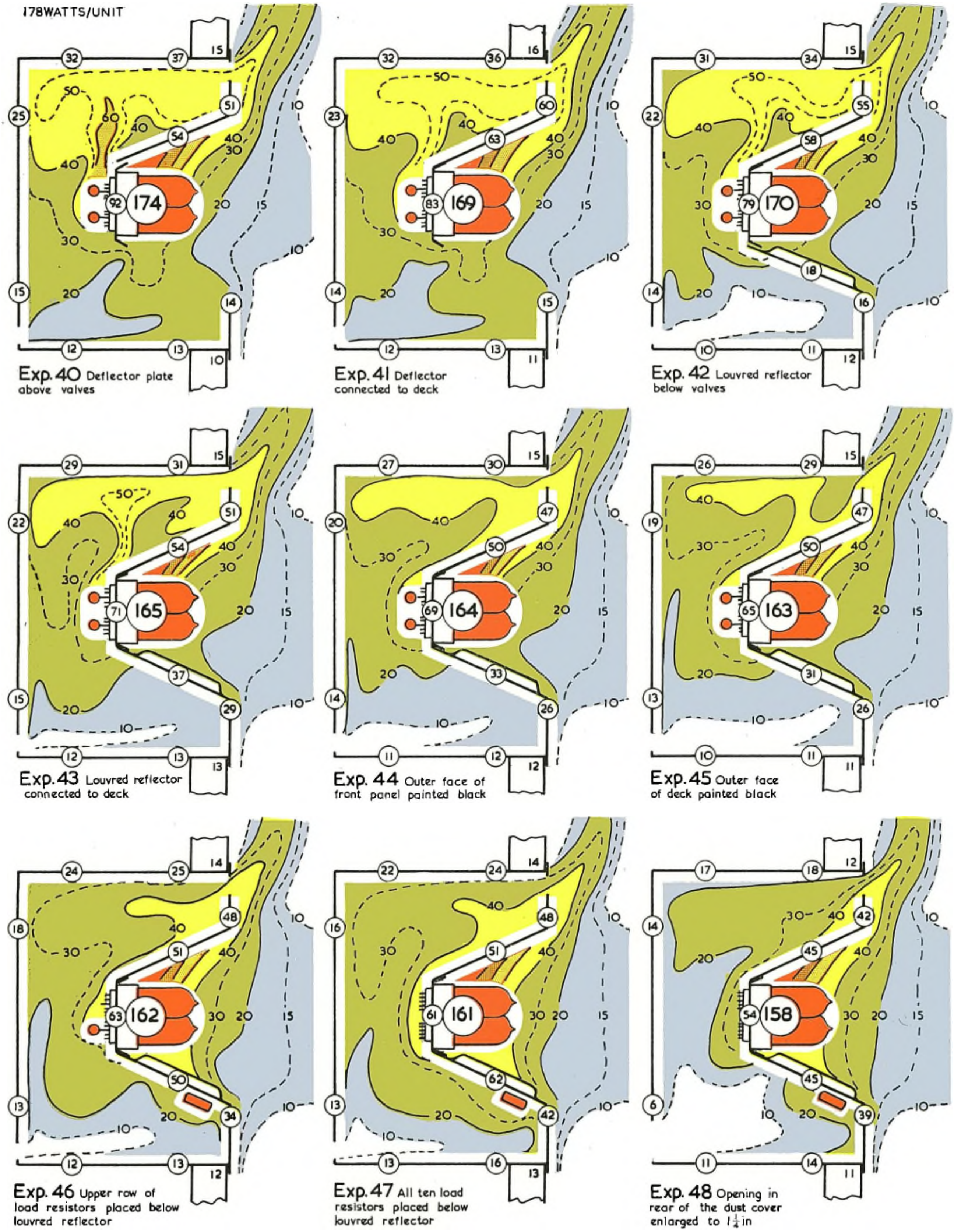


Plate 7. Control of heat flow by various means

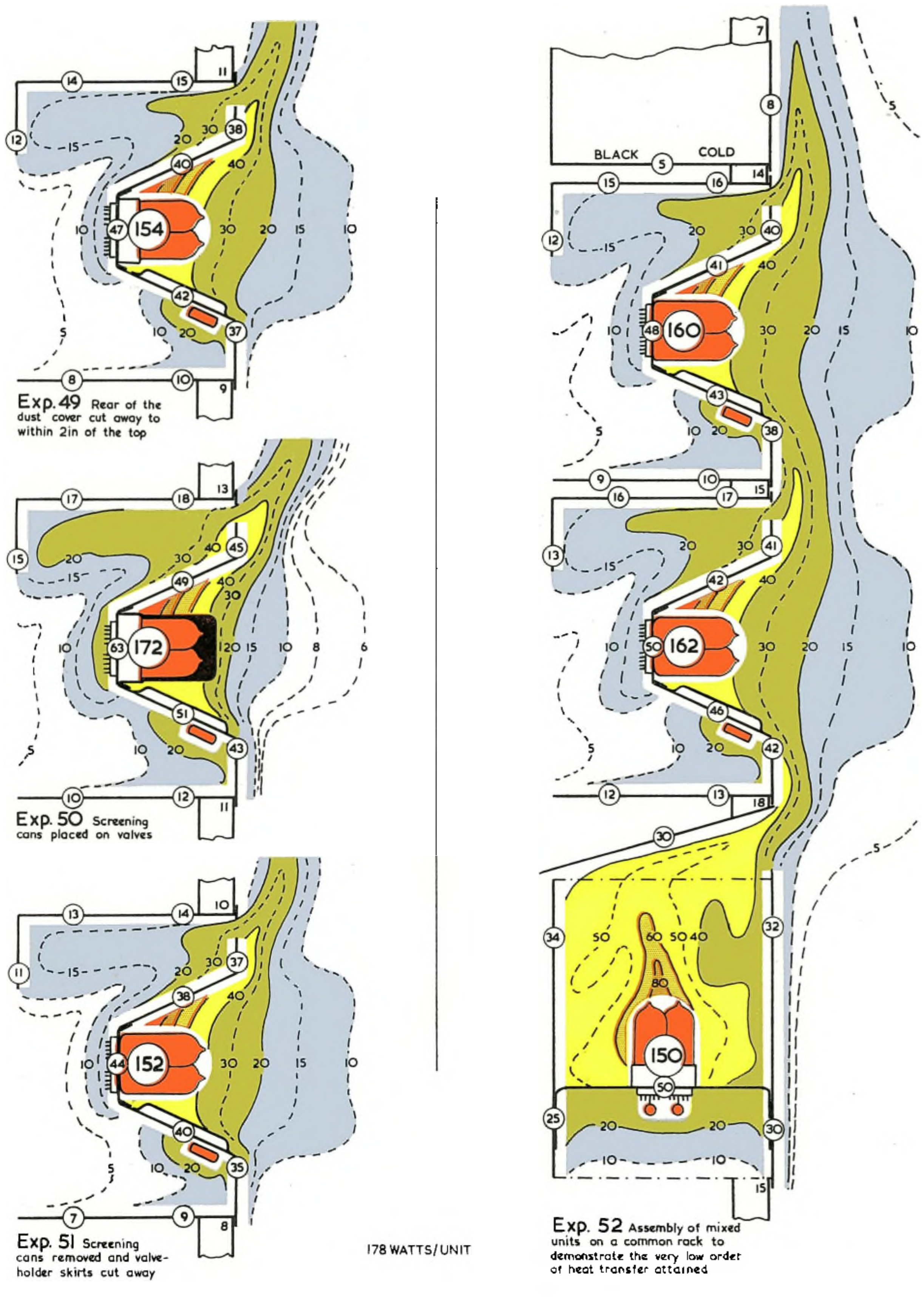


Plate 8. Final version and assembly of units on a rack

Forced Ventilation

It seems hardly necessary to state that, compared with equivalent space factors and thermal loadings in conventional units, the markedly superior heat-dissipating properties of the arrangement evolved makes forced ventilation completely unnecessary. If, for any of a variety of reasons, it becomes desirable to resort to closed circuit ventilation, the existence of a natural convection stream and the absence of frustrating horizontal surfaces will require little additional power to maintain circulation.

The Final Unit as a Bench Instrument

Placed flat on a bench, the final unit grew slightly warmer than in Exp. 51 because it was denied the rack as a heat sink, but this rise was cancelled by tilting the instrument some 5° to 10° from the horizontal to make the deflector angle more obtuse and to spill out the remaining frustrated air. Variations were so small that detailed measurements were not made, it being assumed that conditions in the interior were effectively the same as in Exp. 51.

An instrument case similar in design to the dust cover, but with the open rear more practically reproduced by large-aperture louvres, would suit this application and would benefit from being slightly tilted. With all the heat being extracted through the front, case finish would have little effect upon internal temperatures, although care would have to be taken not to obscure the rear air inlets.

A more attractive proposition would be a truly universal instrument with a dust cover louvred at the rear suitable for mounting on an open rack, in a cubicle without its dust cover, or for use as a bench instrument on a tilted bench stand.

Internal Layout

The virtual exclusion of the heat source from container interiors offers designers the opportunity of positioning components in such a manner that they are both more accessible and less subjected to thermal stresses than when assembled in the confined space below an 'inverted tray' chassis. A suggested layout is shown in Fig. 10, in which the heat-dissipating components are grouped under the louvred reflector to add their convection currents to the main ventilating stream and the thermally insensitive items such as coupling and decoupling capacitors occupy a corresponding position above the upper deflector plate. The cooler rear of the interior is reserved for temperature-sensitive elements such as crystal diodes, selenium rectifiers and all close-tolerance components affecting the equipment accuracy and stability. Adjacent components are mounted on opposite sides of a tag strip to reduce impedance to air flow and supported between the valve deck and upper and lower structural members at the rear, the diagonal mounting retaining access to both sides of the tag strips and to the valve bases, with the 'end-on' aspect of the components guarding against radiation heating from the low-emissivity, but quite hot, valve recess. The only place in which a conventional group board will not impede ventilation is above the deflector plate, where its use will also reduce radiation heating.

This form of layout should suit electronic computers and similar instruments, either along the whole length of a standard 19in panel or, where many identical units are used, divided into a number of modular plug-in sections.

In instruments requiring numerous controls on the front panel a compromise must be made, either by increasing the panel height to accommodate both the valve recess and the necessary controls, or by a further process of segregation shown pictorially in Fig. 11. This represents an actual low-frequency oscillator that was divided into three sections, heat flow control being applied to the centre portion containing the valves and other heat-dissipating components, with the 'chunky' items of the power supply on the right and the equally 'chunky' temperature-sensitive components on the left with controls appearing on the front panel. Frequency-determining items such as the 10μF capacitors were mounted on a vertical face as remote as possible from the heat source—a segregation that paid dividends in the form of an exceptionally high order of frequency stability for this type of instrument.

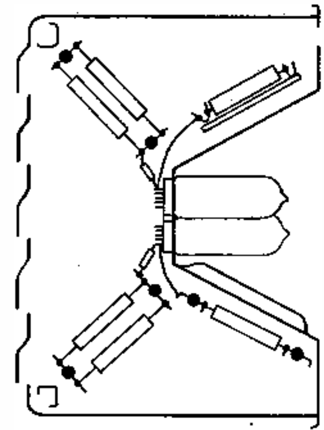


Fig. 10. Component positioning for maximum accessibility and minimum impedance to ventilation

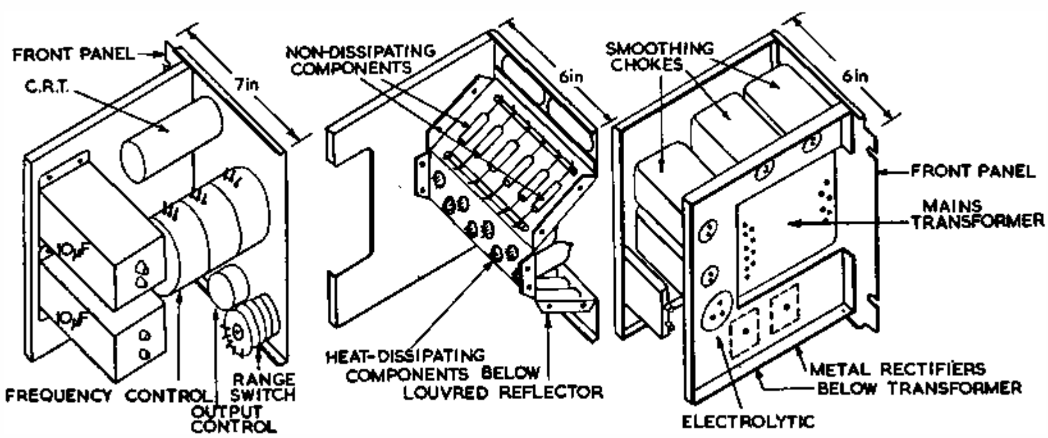


Fig. 11. Exploded view of an instrument with heat control applied to a central valve compartment from which all temperature-sensitive and bulky components have been excluded

Finally, when contemplating the application of the principles of heat control outlined in this article, the designer is advised to avoid squat containers of excessive depth with large horizontal frustrating surfaces and to concentrate upon shallow containers of greater height. Larger panel areas will aid dissipation and make easier the compromise between positioning of controls and valve recesses, while decreased depth will permit more accessible placing of components in the cool stream of air entering from the rear.

Conclusions

The accuracy, stability and life-expectancy of electronic equipment containing temperature-sensitive components is adversely affected by high ambient temperatures generated by heat-dissipating elements, especially when these are miniaturized. Conventional constructional methods make it difficult to take full advantage of natural cooling techniques which, successful in a singly-mounted unit, introduce unpredictable hazards when such a unit is mounted between others in a rack. While an interim solution reduces this hazard with units of conventional design, the more complete understanding of the mechanism of heat transfer attained during experiments suggests a new approach with the basic concept that a heat source, miniature or otherwise, be treated as a thermal menace to be segregated from temperature-sensitive components.

Applied to that most prolific dissipator of thermal energy, the thermionic valve, this concept is realized by a logical system of natural cooling designed to reduce its potentialities as a thermal menace by controlling all means of dissipation into channels that prevent heat entering the container interior or being transferred to adjacent units in a rack. Virtual exclusion of the valve dissipation allows a very accessible disposition of components to be made within the container, with streams of air cooling the more temperature-sensitive elements.

Equally effective in both bench and rack-mounting equipment, the system is particularly applicable to assemblies of units mounted one above the other in a rack, reducing inter-unit heat transfer to a minimum without any deviation from long-established mounting methods and making forced ventilation completely unnecessary.

Acknowledgments

Crown copyright reserved; reproduction by permission of H.M.S.O.

A Simple Apparatus for Measuring Circuit Capacitances

By J. C. S. Richards*, B.Sc., Ph.D.

A simple instrument is described for measuring capacitance to earth in the range 0 to 300pF with an accuracy better than ±2 per cent or ±0·3pF. Since up to three feet of cable can be used between the instrument and the capacitance under test, and since a moderate value of resistance in parallel with the capacitance does not appreciably affect the operation of the instrument, the apparatus is particularly suitable for measuring 'stray' or circuit capacitance.

CIRCUIT or 'stray' capacitance plays an important part in determining the performance of electronic instruments. In most laboratories general purpose bridges and 'Q'-meters are available, but these are rarely suitable for measuring circuit capacitances since such capacitances are (a) generally shunted by a resistance of comparable impedance, (2) must of necessity be connected to the measuring instrument by long leads and (c) have one terminal earthed. Although instruments are obtainable commercially which are ideally suited to stray capacitance measurements, they are expensive and not often available in a small laboratory. To eliminate the guesswork involved in estimating circuit capacitance the present instrument was devised. With the possible exception of a calibrated variable capacitor, all the components are of the type held in stock or easily constructed in a small laboratory.

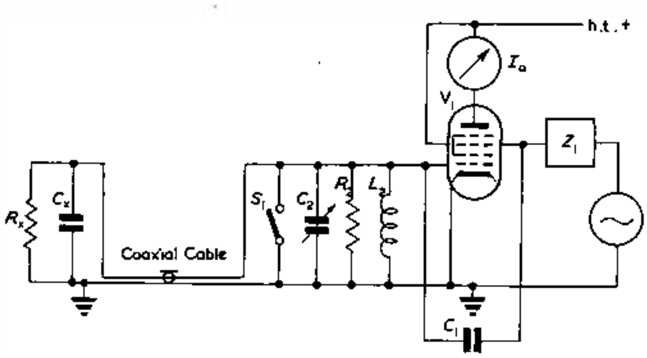


Fig. 1. Basic circuit of instrument

The instrument will measure a capacitance to earth of up to 300pF with an accuracy better than ±2 per cent or ±0·3pF. Under favourable conditions it will detect changes in capacitance of less than 0·1pF. Four operating frequencies are provided, 2Mc/s, 600, 200 and 60kc/s, but more may be added as required. The resistance in parallel with the capacitance under test may be as low as 3kΩ when the test frequency is 2Mc/s; on other ranges the minimum resistance is inversely proportional to the frequency. Three feet of coaxial cable may be used between the measuring instrument and the capacitance under test. Since the test potential is less than 1·5V r.m.s., measurements on 'live' circuits may often be made.

General Principles

The essence of the problem is to measure a small capacitance C_x which has in parallel with it a resistance R_x. The value of R_x may be determined by inspection of the actual resistors in the circuit, and it is only of relevance to the present discussion in so far as it interferes with the measurement of C_x. The obvious way to measure C_x is by substitution in a parallel resonant circuit, but when R_x is small, the Q of the circuit is small, and it is difficult to detect resonance by observing the magnitude of the impedance of the resonant circuit, as is done in most Q-meters. However, even in low Q circuits, the phase-angle of the impedance changes rapidly as the circuit is tuned through resonance, and the present instrument is based on this phenomenon.

The basic circuit is shown in Fig. 1. A sinusoidal e.m.f. is fed from a generator of internal impedance Z_1 to the third, or injection grid of a hexode (or heptode) valve V_1, and also, via a small capacitor C_1, to a parallel resonant circuit connected to the first, or signal grid of V_1. R_3 represents the losses in L_2. If the alternating potentials at the injection and signal grids are respectively $e_1 \cos \omega t$ and $e_2 \cos (\omega t + \phi)$, then when $L_2(C_1 + C_2 + C_x)\omega^2 = 1$, we have e_2/e_1 a maximum, and $\phi = \pi/2$. Because of the mixing action in the valve, the anode current I_a has an additional

* *Aberdeen University.*

steady component δI_a given by

$$\delta I_a = g_c e_2 \cos\phi \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where g_c is the conversion conductance of the valve for the given value of e_1. When ϕ is $\pi/2$, δI_a is zero. Hence to make a measurement, C_x and R_x are disconnected and C_2 is adjusted so that the anode current is the same whether the switch S_1 is open or closed; then, with C_x and R_x in circuit, C_2 is re-adjusted so that the anode current is again unaffected by opening or closing S_1. The change in C_2 gives C_x directly.

A somewhat similar circuit for detecting small changes in capacitance has been described by Alexander[1] and Hobday[2]. These authors use an octode in which the first grid is the injection grid and the third grid is the signal grid. They omit C_1, and rely on space charge coupling within the valve to induce an e.m.f. in the resonant circuit connected to the third grid. The main disadvantage of their circuit for present purposes is that the coupling to the tuned circuit is not wholly reactive, which it must be if the value of R_x is not to affect the measured value of capacitance, and the degree of coupling is not under the circuit designer's control.

Operating Frequency

The permissible range of operating frequencies is limited at the lower end by the minimum value of R_x likely to be encountered, and at the upper end by the self inductance of the cable between C_x and the instrument.

For small departures from resonance the sensitivity is given with sufficient accuracy by

$$\partial I_a/\partial C_2 = g_c e_1 C_1 \omega^2 \left(\frac{R_x + R_2}{R_x R_2}\right)^2$$

$$\simeq g_c e_1 C_1 \omega^2 R_x^2$$

when $R_x \ll R_2$. Once $g_c e_1 C_1$ has been made as large as feasible, the only way to obtain adequate sensitivity when R_x is small is to make ω large.

The errors due to self inductance of the variable capacitors etc. are in practice swamped by the errors due to self inductance of the connecting cable. Provided this cable is appreciably less than a quarter of a wavelength long, the error δC_x in measuring C_x is given with sufficient accuracy by

$$\delta C_x = (R_o^2/R_x^2)\cdot C_c + R_o^2 C_c (C_x + C_c)\,\omega^2 C_x$$

where R_o is the characteristic impedance of the cable and C_c its total capacitance. With 3ft of 75Ω cable, for which C_c is about 66pF, the first term is negligible if R_x is greater than 1kΩ. If error due to the second term is not to be appreciably more than 2 per cent when C_x is 300pF, the frequency must be less than 2Mc/s. The maximum error can be slightly reduced by including a small inductance in series with C_2, but the compensation can be made perfect for only one value of C_x.

Design Considerations

The main difficulty encountered in designing a practical circuit is to make the measured capacitance independent of R_x. Apart from its obvious effect on the 'Q' of the circuit, changing the value of R_x changes the magnitude e_2 of the alternating potential on the signal grid, and changes the loading on the generator feeding the injection grid.

If there is any appreciable even harmonic distortion or grid current flow in the valve, the mean anode current will be dependent on the alternating potentials at the grids. Now e_2 is equal to $e_1\omega C_1 R_2 R_x/(R_2 + R_x)$, and since it cannot be kept constant when R_x is changed, its maximum value, given by $e_1 R_2 C_1 \omega$, must be kept small (typically < 1 or 2V r.m.s.). To attain an adequate value of the conversion conductance g_c, the magnitude e_1 of the alternating potential at the injection grid must be large (typically > 5V r.m.s.), and current is normally drawn from this grid. In order that e_1 shall not change appreciably when R_x is changed, Z_1 must be small. In addition, the frequency of the generator must be independent of R_x. These considerations preclude the use of the triode section of a triode-heptode valve in a conventional oscillator circuit, and a separate valve must be used to generate the required alternating e.m.f.

Practical Circuit

A detailed circuit is shown in Fig. 2. Four operating frequencies (nominally 2Mc/s, 600, 200 and 60kc/s) are available, but there is, of course, no difficulty in increasing this number. The coils used in the prototype are wound on 15/32in diameter formers with iron dust cores, with the exception of L_1, L_5 and L_9, which are wound on 'Ferroxcube' pot cores and L_{10}, which consists of three small loops in the wiring. Almost any coils having 'Q'-factors > 50 may be used.

The oscillator valve V_2 is employed in a Hartley type of circuit. Since there is negligible standing bias on the valve, the mutual conductance is high, and the output impedance low. The resistors R_{11}, R_{12}, R_{13} and R_{14} serve to isolate the tuned circuit from the valve. The values given are suitable for the prototype coils; for other coils the values may need to be decreased if oscillations do not start. The crystal diode MR_1 and its associated circuit limit the amplitude of the oscillations, as measured at the cathode of V_2, to about 6V r.m.s. The oscillator circuit must be well screened from the input circuit so that C_1 and the interelectrode capacitances provide the only coupling.

With R_x infinite, the alternating potential at the signal grid should be about 1·5V r.m.s. If the coils used have a 'Q'-factor appreciably greater than 50, it may be necessary to add resistance in parallel to reduce the voltage to this figure. In order to minimize the effects of heater voltage changes etc. on the microammeter M_1 in the anode circuit of V_1, the backing-off current is taken from the screen grids of the heptode section and the anode of the triode section of the valve, and the total cathode current is stabilized by means of the relatively large cathode resistor R_9. This method of connexion serves to reduce the effect of variations in the potentials applied to the injection and signal grids. Decreasing the bias on the injection grid (and hence also the triode grid), increases the triode anode current, but has little effect on the screen grid current of the heptode. Decreasing the bias on the signal grid increases the hexode anode and screen currents, but has little effect on the triode anode current. The changes in mean current produced by even harmonic distortion are minimized by the same mechanism. R_5 is used to set the meter to zero with switch S_1 closed, and it should need only occasional adjustment. The crystal diodes MR_2 and MR_3 protect the meter from gross overloading.

The high tension supply is obtained from a conventional gas-tube stabilizer. The maximum current drain is about 25mA.

THE VARIABLE CAPACITOR

The accuracy of measurement depends almost wholly on the calibration of the variable capacitor. A single variable capacitor with a scale directly calibrated over a range of 300pF and on which increments of (say) 0·1pF can be read is relatively expensive and not readily available. In the present instrument, five steps of 50pF are provided by the switch S_2 and C_5, C_6 and C_7, which are ± 1 per cent nominal tolerance components selected to be within $\pm 0{\cdot}5$ per cent. Interpolation between the steps and the measure-

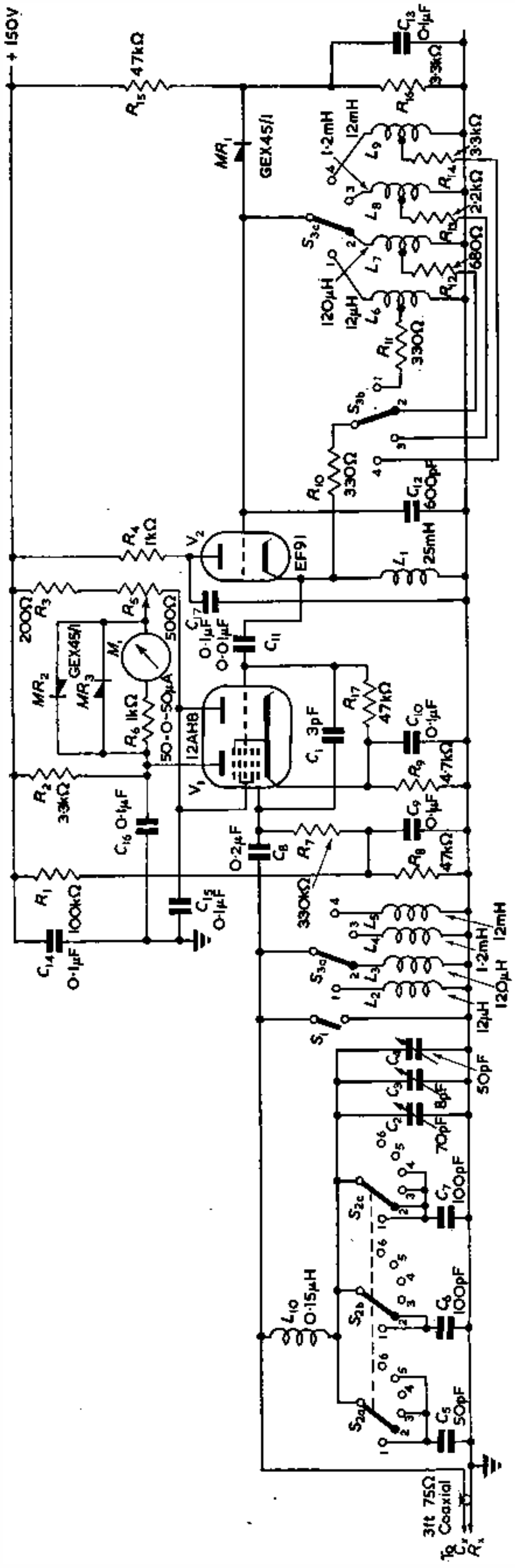


Fig. 2. Detailed diagram of practical circuit

S_{3a}, S_{3b} and S_{3c} are ganged. All resistors associated with V_1 are of the high stability (or wire-wound) variety

ment of small increments of capacitance are done by C_2 and C_3 respectively. These are linear law capacitors. If not available ready calibrated, they may be calibrated in the laboratory very easily because of the relatively low accuracy required over the limited range. All scales are marked in picofarads *removed* from circuit. C_4 is an uncalibrated capacitor used to set zero on the calibrated scales when C_x is zero.

Apart from the 0·15μH choke in series with the variable capacitance which reduces the error due to the cable inductance when measuring large capacitances on the 2Mc/s range, the lead inductances associated with the variable capacitors should be kept as small as possible. Three feet of 75Ω, 22pF/ft coaxial cable are used to connect the capacitance under test. To the free end of this cable are attached two short (2 or 3in) leads with miniature crocodile clips.

Performance

If R_x is infinite, detuning C_2 by 1pF gives a meter deflexion of about 30μA, or almost one-third full scale. The accuracy on the three lower frequency ranges is then limited by the calibration error in the variable capacitors to ±1 per cent or ±0·3pF. At a frequency of 2Mc/s, the error arising from stray inductance is less than ±2 per cent.

The smallest meter deflexion which occurs when S_1 is open or closed, and which can with certainty be ascribed to capacitance changes, is less than 3μA, so that when R_x is infinite, changes in capacitance as small as 0·1pF can be detected. When R_x is small the sensitivity is approximately proportional to $R_x^2f^2$, where f is the frequency. For the deflexion resulting from detuning the variable capacitance by 1pF to be greater than 3μA, the product R_xf must be greater than 6 000, where R is in kilohms and f in kilocycles/sec. This parameter can be conveniently taken to define the lowest value of R_x in the presence of which a reasonably accurate capacitance measurement can be made. For the frequencies provided, 2Mc/s, 600, 200 and 60kc/s, the corresponding minimum values of R_x are 3, 10, 30 and 100kΩ. Useful measurements can often be made with resistance as small as one-third of these values, with, of course, reduced accuracy.

It is possible to find the individual components of a three terminal capacitance, provided one of the terminals is earthed, by making three measurements of capacitance to earth with each of the components short-circuited in turn. The errors in the results depend on the relative values of the capacitors, and on whether the short-circuiting of one of the components can be done without disturbing the values of the other two.

Conclusions

A device for measuring capacitance to earth has been described which can readily be constructed from components available in most laboratories. The instrument is suitable for measuring capacitance in 'live' circuits. Complete with power supply, the apparatus may easily be contained in a case 7in × 14in × 9in and weighs about 25lb, so that it is readily portable.

Acknowledgments

The author's thanks are due to Mr. W. M. Stratton for constructing the prototype apparatus.

REFERENCES

1. ALEXANDER, W. An Electronic Ultramicrometer. *Electronic Engng. 23*, 479 (1951).
2. HOBDAY, S. M. An Electronic Ultramicrometer (Letters to the Editor). *Electronic Engng. 24*, 134 (1952).

A Low-Voltage D.C. Stabilizer Using a Saturated-Diode Controller

By F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and M. S. Seaman*, B.Eng.

An experimental d.c. stabilizer with an output voltage of 6·3V and capable of supplying currents up to 30A is described. It uses a saturated diode as the controlling element and is particularly suitable for supplying valve heaters. The circuit can be adapted to operate at other output voltages.

An a.c. stabilizer employing an electronically-controlled transductor circuit has been developed and described by the authors[1]. The basic arrangement adopted for it, now fairly well known in principle, is similar to that published by Helterline[2,3,4] in which a temperature-limited diode is used as the controlling element. The diode is placed in one arm of a bridge circuit, the other arms being composed of resistors. Such a bridge will balance at a certain filament voltage, but since the filament voltage controls the anode current of the diode, a small change of filament voltage gives a large unbalance voltage. The unbalance voltage is applied to the grid of a pentode, or beam tetrode, the output of which is used to control the voltage requiring stabilization via a saturable-core reactor with an associated auto-transformer.

Helterline[2,4] has shown that a circuit of this type can be adapted to control a d.c. voltage by passing the a.c. output through a rectifier and filter circuit. The diode is then actuated by the d.c. output voltage. A stabilized d.c. supply at 6·3V has been developed using this principle and its construction and performance are described here. The stabilizer will provide a current up to 30A and is particularly attractive for supplying valve heaters, instead of using secondary cells, in cases where a.c. heating is undesirable.

* *University of Sheffield.*

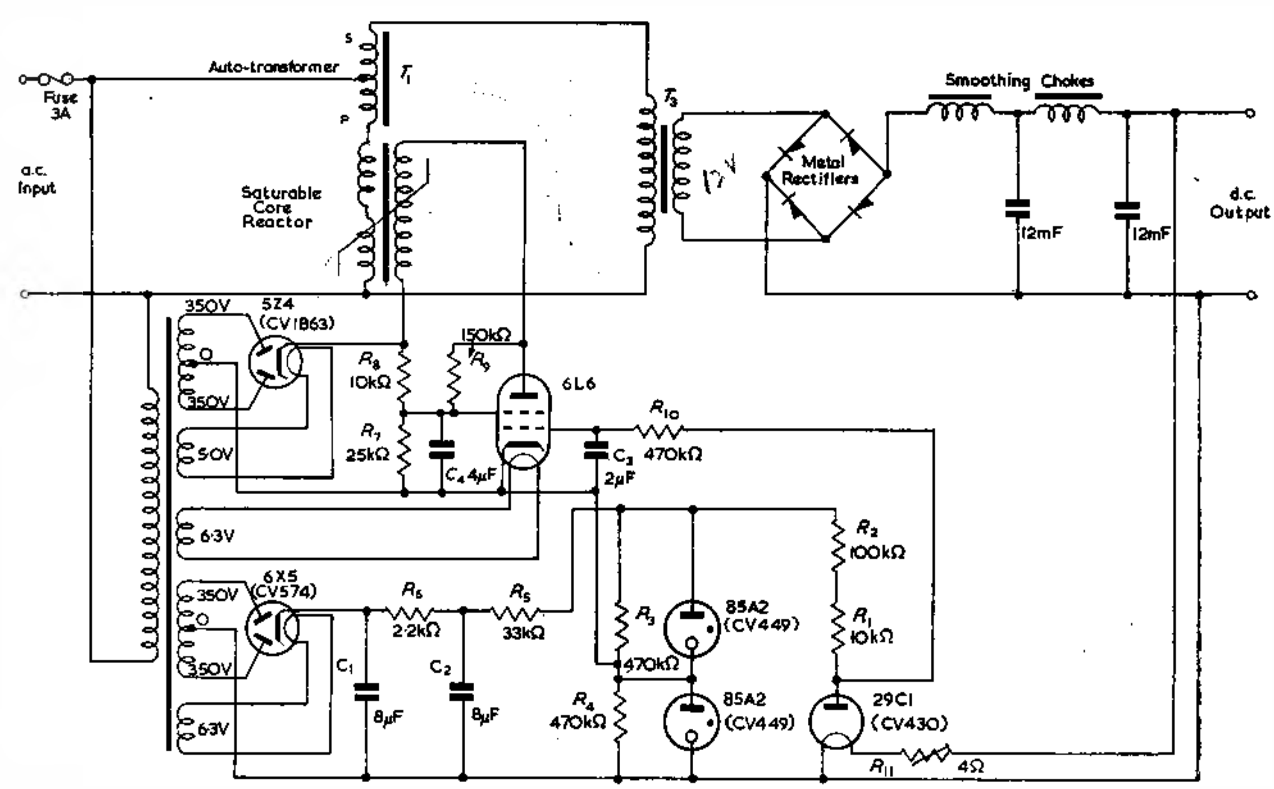
The Stabilizer and its Performance

The circuit arrangement of the stabilizer is shown in Fig. 1. It is essentially the same as the a.c. unit previously described[1] with the addition of a suitable transformer, rectifier and filter. The output voltage of the a.c. stabilizer now becomes the primary voltage of the additional transformer supplying the rectifier. Owing to the impedance of the transformer, rectifier and filter, this voltage does not remain constant, especially as the load varies, but the variation is not very great.

The designs of the diode-bridge network, saturable-core reactor and autotransformer for the a.c. stabilizer have already been briefly described[1]. The same diode bridge is used here. The optimum number of reactor turns, and the best ratios for the auto-transformer and the transformer supplying the rectifier have been found experimentally.

Suitable rectifiers for the purpose are STC Type B125-1-1CW which have a maximum current of 10A at 12V d.c. with full-wave bridge connexions. Three such units are therefore used in parallel. A two-element L-type choke-capacitor filter is employed for smoothing the output of the rectifiers. The use of a π-type filter was ruled out

Fig. 1. Circuit of stabilizer



because it gave rise to large peaky input currents, which might have limited the available output current, and because it had poor regulation which would probably have been reflected in the performance of the unit. Each capacitance element of the filter consists of six 2 000μF, 12V wkg. electrolytic capacitors in parallel.

The optimum air gap for each smoothing choke was found experimentally. The incremental impedances of each choke at a frequency of 50c/s for varying air gaps and direct currents were calculated from the experimental figures; the results are shown in Fig. 2. An air gap of 0·139in was eventually chosen giving the chokes an inductance which varied from 8·4 to 6·6mH as the load current varied from 5 to 30A. As will be seen from Fig. 2, a smaller air gap gave a larger inductance at lower currents, which decreased considerably as the load current increased, while larger air gaps gave a lower inductance which remained more nearly constant as the current varied.

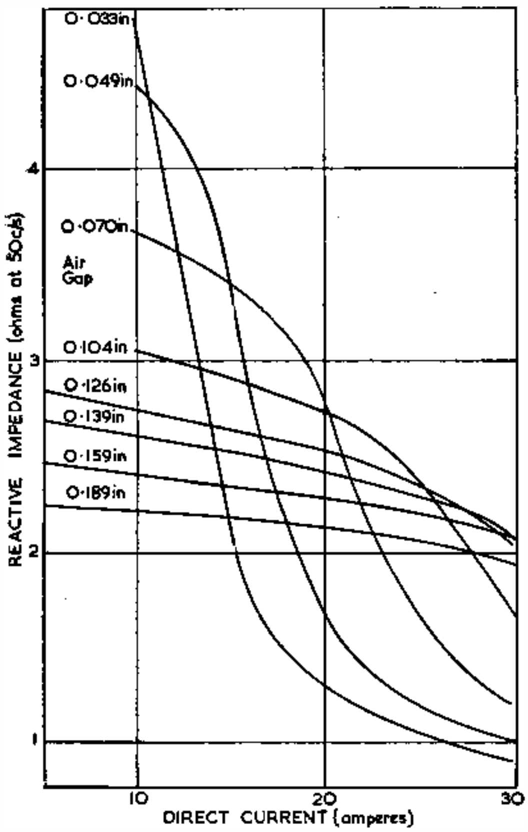


Fig. 2. Smoothing choke characteristics for different air-gaps

Tests on the a.c. stabilizer had shown that the EF55 valve gave a better performance, but a smaller maximum load current, than the 6L6 amplifier valve which is used here since it was thought that the 100c/s voltage peaks reported earlier[1] would again occur on the anode in the absence of an additional feedback circuit. The h.t. supply for the amplifier valve is obtained from a standard transformer and valve rectifier and no smoothing is employed, although a decoupling capacitor is used for the screen supply.

The diode-bridge supply is obtained from the unstabilized input to the unit via a transformer and valve rectifier and is stabilized with two 85A2 glow-discharge tubes operating at a nominal current of 6mA. The current through the 85A2 tubes varies from about 3·8mA at 190V input to about 8·2mA at 270V input; the corresponding change in the running voltage of each tube being about 1V[5]. A 2μF capacitor had to be included between the grid and cathode of the amplifier valve to prevent 'hunting'. This 'hunting' was thought to be due to the saturated-diode response being too fast for the reactor and smoothing circuit. Unfortunately, the inclusion of this capacitor caused a marked increase in the response time to about half a second.

The effect of varying the number of reactor turns in this stabilizer is the same as in the a.c. unit. Increasing the number of turns requires greater polarizing currents with the result that the maximum load current decreases. Also, larger changes in the polarizing current are required to counteract a given variation in either input voltage or load current and consequently the performance falls off. With too few turns the polarizing current is insufficient to exert adequate control, especially at low load current. The

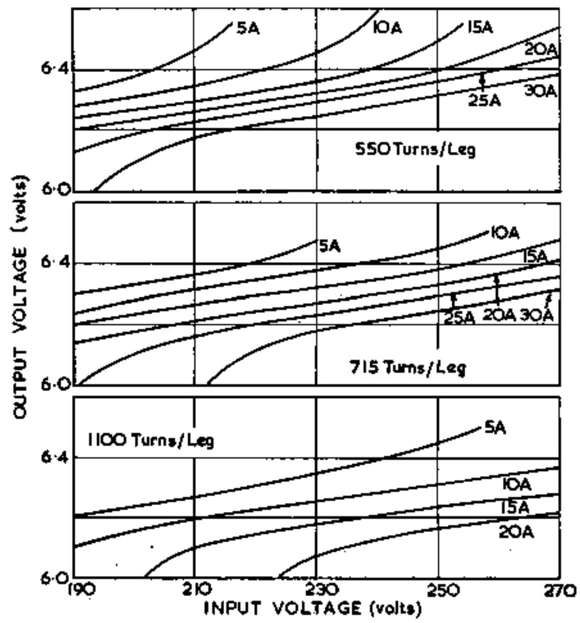


Fig. 3. Effect of varying the reactor turns
Auto-transformer ratio 1·64 : 1
Rectifier transformer ratio 800 : 41

variation of performance with the number of reactor turns is shown in Fig. 3, from which it will be seen that the main effect is to vary the regulation and the useful load-current range, the effect on the stabilization being comparatively small.

The variation of performance with various auto-transformer ratios is shown in Fig. 4. With the tapping near the output end the input power factor and efficiency improve slightly, the maximum output power increases and the unit has a poor performance at light loads. Moving the tapping towards the reactor end improves the stabilization, but makes the regulation worse since greater polarizing currents are required in the reactor.

The effect of varying the ratio of the rectifier transformer is shown in Fig. 5. With a small number of primary turns the primary voltage is low, necessitating high voltages and currents in the reactor. Since this involves relatively small polarizing currents the performance on light loads is rather poor. As the number of primary turns (and with it the primary voltage) is increased the input power factor and the efficiency improve and the input power decreases mainly due to the lower voltages on the reactor. Larger polarizing currents are needed, however, and the stabilizer is now unable to satisfactorily control heavy loads.

From the curves obtained it is found that the optimum

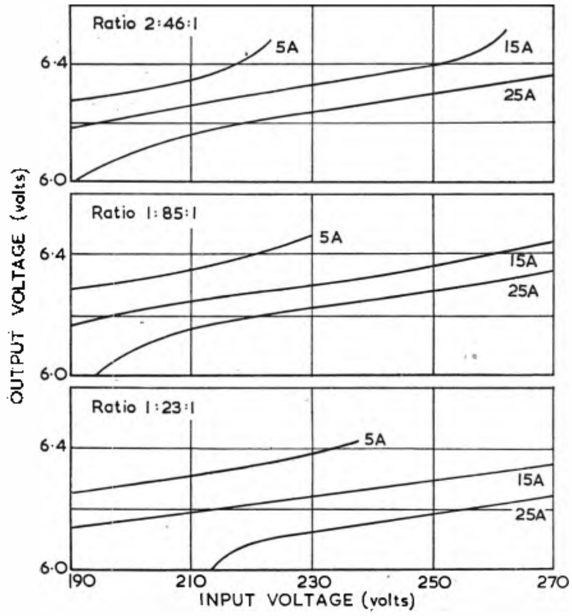


Fig. 4. Effect of varying the auto-transformer ratio
Reactor turns 715 per leg
Rectifier transformer ratio 800 : 41

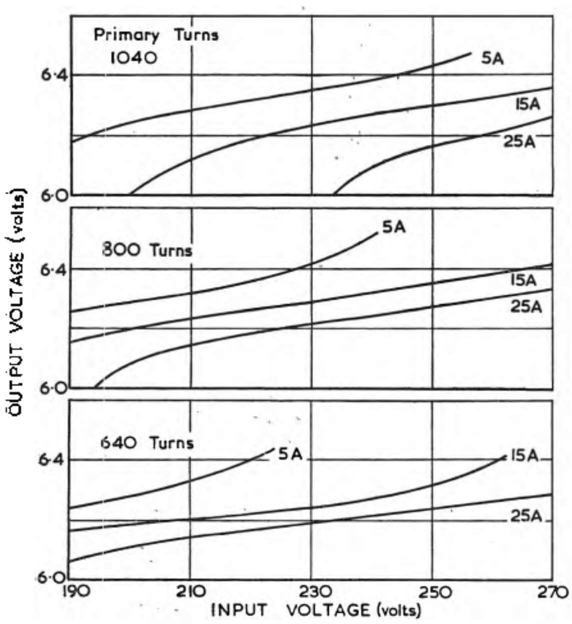
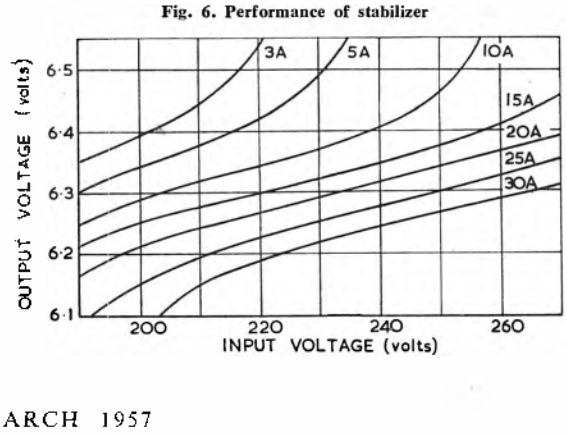


Fig. 5. Effect of varying the rectifier transformer ratio
Auto-transformer ratio 1·85 : 1
Reactor turns 715 per leg
Rectifier transformer secondary turns 41

Fig. 6. Performance of stabilizer



number of reactor turns, auto-transformer ratio and rectifier-transformer ratio are not very critical and are a compromise between stabilization, regulation and load-current range. The values finally adopted are given in the appendix.

The performance of the final stabilizer is shown in Fig. 6 and the manner in which the various internal voltages and currents vary with input voltage and load current are shown in Figs. 7 and 8 respectively.

Since in the stabilizer there is only the diode filament as a permanent load across the output, compared with all the internal supplies in the a.c. unit[1], the performance on light loads is comparatively poor, the output voltage rising

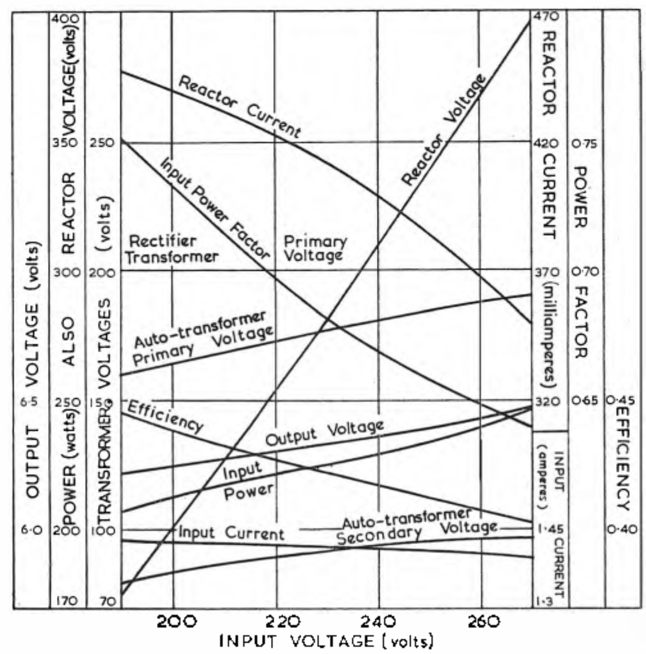


Fig. 7. Variation of internal voltages, currents, etc., with input voltage. Fixed load of 15A (approx.)

excessively. This situation is not helped by the action of the filter capacitors which perform the bulk of the smoothing under these conditions.

When switching the equipment on from cold the main h.t. supply is available about a second or two before the bridge h.t. voltage is sufficient to strike the glow-discharge tubes. If the load current is less than about 10A the polarizing current in the reactor then rises to a value higher than the load and input voltage conditions require, the result being that the output voltage rises excessively during this short interval. A good case exists therefore for having a permanent load, taking about 10A, across the output.

The efficiency of the unit is only about 45 per cent compared with 75 per cent for the a.c. stabilizer because of the additional losses in the rectifier-filter circuit. Due to the harmonic currents the input power factor is also lower than for the a.c. unit and, as before, becomes worse as the input voltage increases.

The ripple on the output voltage is about 100mV and remains fairly constant over the load-current range. A typical oscillogram is shown in Fig. 9 for a load of 15A at 230V input. The effect of varying the input frequency on the performance of the stabilizer has not been investigated, but experience with the a.c. unit suggests that the output voltage should not vary appreciably with frequency over most of the load-current range. The ripple can, however, be expected to vary widely with frequency because the attenuation of the filter network used varies as the fourth power of the frequency. Thus, at 40c/s a ripple of about 250mV can be expected while at 60c/s it would be only 50mV. The performance at light loads, which is to some extent determined by the smoothing circuit, can also be expected to vary with frequency.

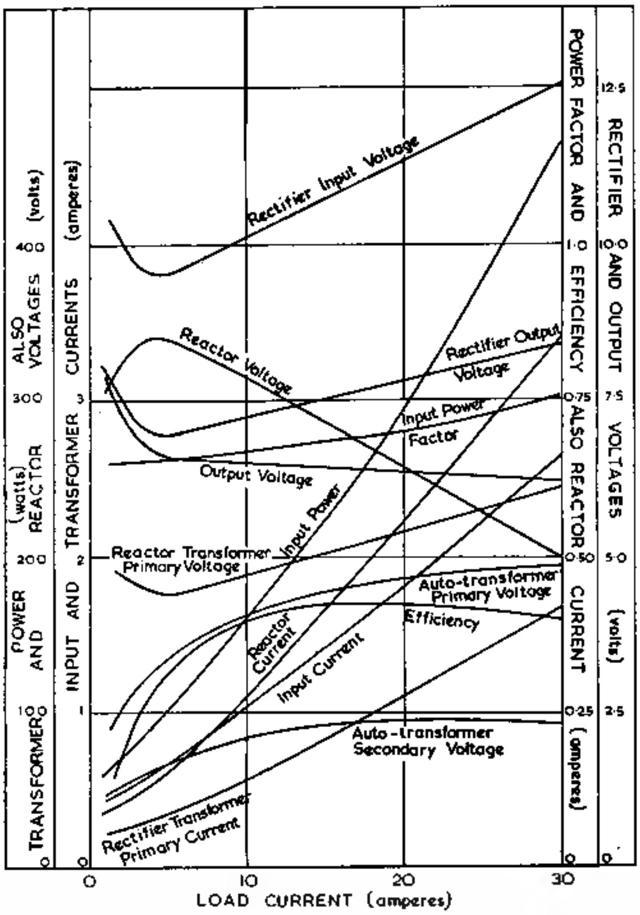


Fig. 8. Variation of internal voltages, currents, etc., with load current

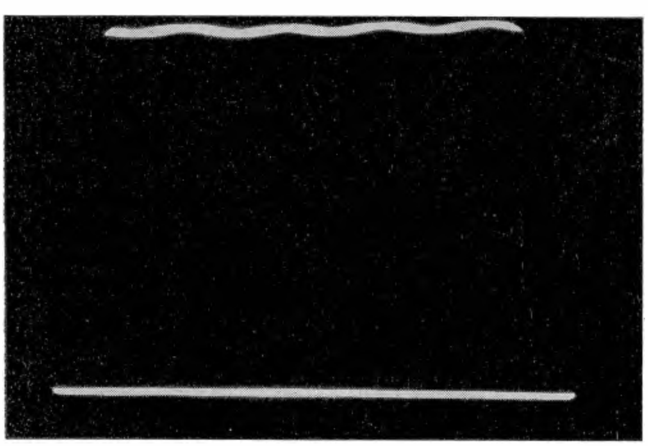


Fig. 9. Typical output voltage oscillogram for a load of 15A at 230V input

As expected, there are large 100c/s voltages on the anode of the amplifier valve, arising from the reactor. These voltages are less than the corresponding ones in the a.c. stabilizer; in this case their amplitude varies with the polarizing current and rarely exceeds 1kV, whereas in the a.c. unit they are usually of the order of 1·5kV irrespective of the polarizing current.

Conclusions

The work described here has demonstrated that the saturated diode is a suitable control element for d.c. voltage stabilizers with a low-voltage output. The circuit can obviously be adapted to operate at any other desired output voltage, the main limitation being on the loss in the diode-filament resistor. For example, for a unit with an output voltage of 200V, using a 29C1 diode the loss in the filament resistor would be about 150W.

The performance of the experimental unit constructed is rather disappointing. It was expected that a stabilization ratio greater than 15 and an effective internal resistance of less than 0·01Ω would be obtained. The use of an EF55 valve for the amplifier would be expected to improve the performance, but would probably reduce the maximum output power because of its lower ratings. With this valve it would also be advisable to reduce the voltage peaks on the anode in some way. With a 6L6 valve the h.t. supply voltage could be increased to give a rather larger output current.

The use of feedback has been tried, as in the a.c. stabilizer and gave a reduction of the peak voltages on the anode of the amplifier valve. This also allowed the delay capacitance to be reduced to 1μF with a consequent improvement in the response time. A detailed investigation of the full effects of using feedback in the d.c. stabilizer, which are not the same as on the a.c. unit because of the smoothing circuit, has not so far been carried out.

There appears to be no reason why a saturated diode should always be used as the controlling element in this type of stabilizer. For moderately high voltages a glow-discharge tube could probably be used in a suitable bridge network in which the h.t. supply is also the voltage requiring regulation.

Acknowledgments

The authors wish to thank the University of Sheffield for laboratory facilities afforded in the Department of Electrical Engineering. They are also indebted to Mr. E. F. F. Gillespie for carrying out some of the early experimental work on the stabilizer components.

Appendix

Data for the Circuit of Fig. 1

COMPONENT RATINGS

Resistors $R_{1,3,4,6,10}$ 1W
R_2 2W
$R_{5,7,8,9}$ 5W
R_{11} to carry 0·85A. Total resistance 4·05Ω, in sections of 2·0, 1·0, 0·4, 0·3, 0·2, 0·1 and 0·05Ω, with provision for short-circuiting each section as required in order to adjust the output voltage.

Resistors R_5 and R_6 should be adjusted by trial and error to give a current of 6mA in the 85A2 tubes at the nominal input voltage.

Capacitors C_{1-4} 500V working
$C_{5,6}$ 12V working (electrolytics)

SATURABLE-CORE REACTOR DESIGN

Core 2¼in stack of Sankey size 35A 'Stalloy' stampings.
D.C. winding 15 000 turns of 36 s.w.g. copper wire
A.C. windings 715 turns of 22 s.w.g. copper wire each

The d.c. winding is placed on the centre limb, and the two a.c. windings, which are connected in series, are placed on the outer limbs.

AUTO-TRANSFORMER (T_1) DESIGN

Core 1¼in stack of Sankey size 36A 'Stalloy' stampings.

Primary winding 743 turns of 22 s.w.g. copper wire
Secondary winding 402 turns of 20 s.w.g. copper wire

INTERNAL SUPPLIES TRANSFORMER (T_2) DESIGN

Core 1¼in stack f Sankey size 36A 'Stalloy' stampings
Primary Winding 24 s.w.g. copper wire
Tap at 0-38-783-860-935 turns for 10-0-200-220-240V.
H.T. windings 36 s.w.g. copper wire
Each winding is 2 700 turns, centre-tapped
L.T. windings Two for 6·3V 1A
25 turns of 20 s.w.g. copper wire each
One for 5·0V 2A
20 turns of 18 s.w.g. copper wire.

RECTIFIER TRANSFORMER (T_3) DESIGN

Core 1½in stack of Sankey size 35A 'Stalloy' stampings
Secondary winding Six 14 s.w.g. copper wires in parallel, each with 41 turns.
Primary winding 720 turns of 20 s.w.g. copper wire.

SMOOTHING CHOKE DESIGN (two required)

Core 2¼in stack of Sankey size 35A 'Stalloy' stampings.
Winding Five 14 s.w.g. copper wires in parallel, each with 108 turns.
Air-Gap 0·139in.

REFERENCES

1. BENSON, F. A., SEAMAN, M. S. An A.C. Voltage Stabilizer. *Electronic Engng.* *28*, 260 (1956) April
2. HELTERLINE, L. Diode Controlled Voltage Regulators. *Electronics 20*, 96 (June, 1947).
3. HELTERLINE, L. Voltage Stabilizers Assure Top Performance. *Elect. Manufacturing 42*, 113 (September, 1948).
4. British Patent 661 872 : 1951.
5. BENSON, F. A. The Characteristics of Some Miniature High-Stability Glow-Discharge Voltage-Regulator Tubes. *J. Sci. Instrum.* *28*, 339 (1951).

A Transistor Cardiotachometer

By L. Molyneux*, B.Sc.

Transistors may be used to construct a cardiotachometer which is small, portable, and may be used in the presence of inflammable vapours. These qualities make the instrument particularly suitable for use in the operating suite of a hospital.

THE problems of recording or displaying the human heart rate have been discussed by Boyd and Eadie[1] who, as well as reviewing previous work, present a method of overcoming the difficulties using electronic valves.

As the instrument to be described here uses transistors as the active circuit elements, it can be economically powered by a battery and thus made completely portable. The small battery needed has also the advantage that it is unable to cause ignition of the inflammable vapours normally found in an operating theatre. It was for use in the theatre and anæsthetic room that it was primarily designed, and a brief account of an earlier form of the instrument has appeared in an anæsthetic journal[2].

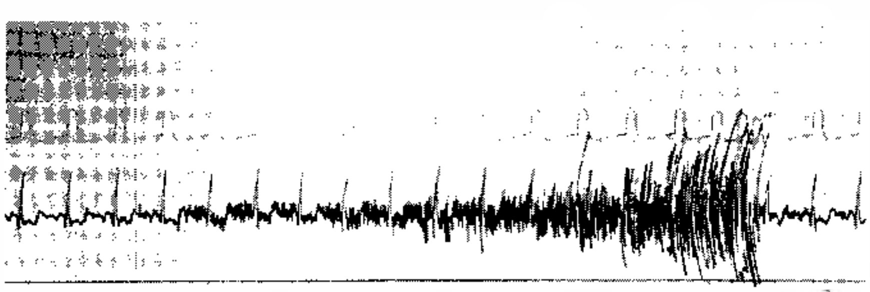


Fig. 1. Traces showing (above) the output from the pulse-forming circuit and (below) the e.c.g. and interference

The design of a cardiotachometer falls into two parts; the first part must amplify the cardiographic waveform, reject all forms of interference and produce a stimulus each time the heart beats; the second part must give an indication of the rate of occurrence of these stimulæ. Boyd and Eadie discuss the types of interference that may occur, and meet the difficulties by using a special type of tuned circuit whose resonant frequency is 6c/s.

It is difficult to translate their circuit into one using transistors, and another method is adopted here. Two filters of the 'parallel-T' feedback type are stagger-tuned to give a pass-band from 10 to 14c/s. This arrangement has proved very satisfactory for the purpose for which it was designed, but while the instrument is insensitive to most of the interference found in an operating theatre it will not work satisfactorily while the diathermy knife is actually being used on the patient.

As this system is new its performance during exercise may be of interest. The following experiments were made:

Wrist electrodes were applied to a subject, and he was asked to pick up a 5lb weight and raise it above his head. As he did so, a recording was made on a twin channel e.c.g. of his cardiograph and the pulse output from the transistor apparatus. In two tests out of three, the instrument made one 'mistake' during the lift. That is to say, it indicated one extra heart beat. In the third test there was no mistake. However, mistakes became quite frequent if the subject continued to hold the weight above his head.

In another test a subject, again with wrist electrodes, was asked to step on to and off a box 7½in high as quickly as he could. During a period of 160 heart beats, representing about 80sec, there was one mistake.

* *University of Durham.*

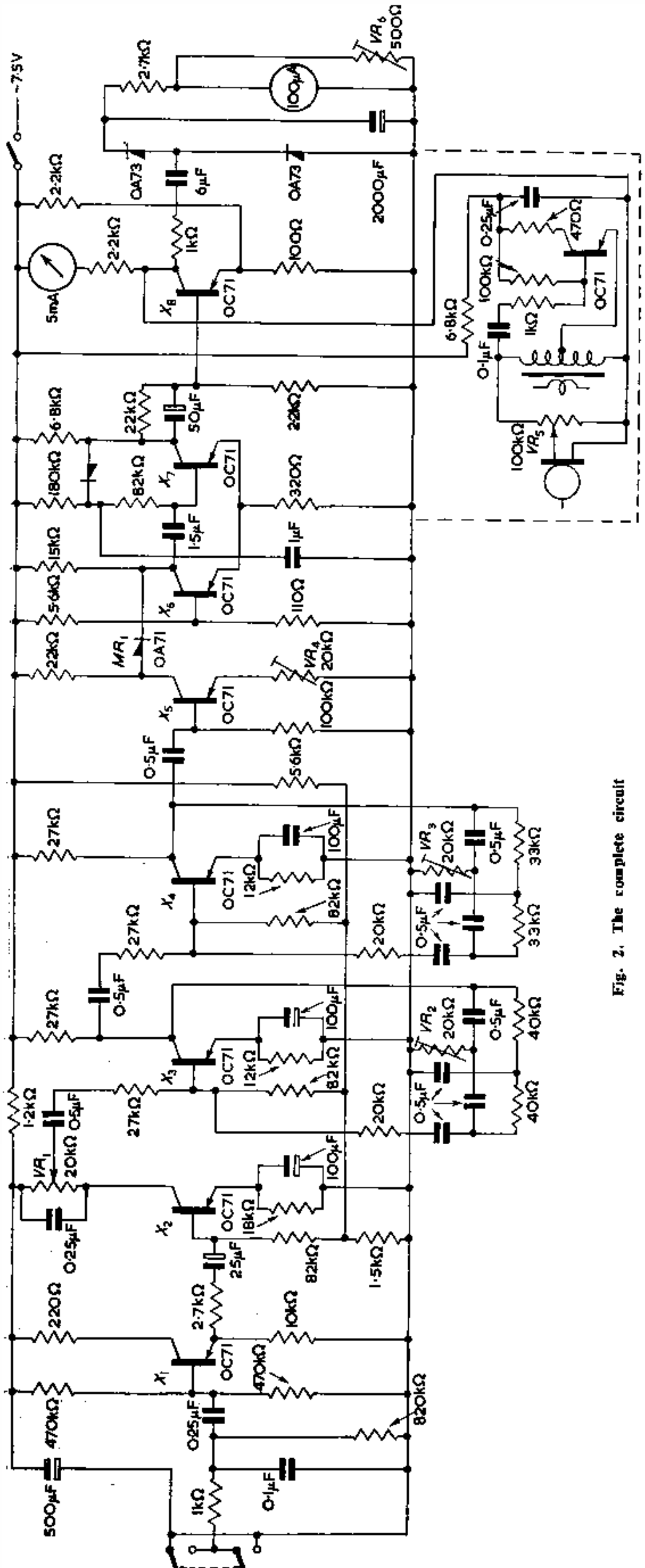


Fig. 2. The complete circuit

In a third test, a subject was asked to clench his fists slowly and was told to release them when a mistake was made. Fig. 1 shows a typical trace. It can be seen that the muscle activity reaches considerable proportions before a false reading occurs. As the pen used to record the muscle potentials had an upper frequency limit of about 80c/s, the actual muscle voltages would be larger than those shown.

The Circuit Used (Fig. 2)

Fundamentally, transistors have a low input impedance. To render the apparatus insensitive to changes in electrode resistance, it is necessary to arrange that the input impedance is fairly high. This is accomplished by X_1, where the 'grounded' collector circuit, which is the transistor equivalent to the cathode-follower, is used. Unlike the cathode-follower, however, the input impedance of this circuit depends on the load placed on its output, and with the circuit values used the input impedance measured at the input terminals of the instrument is about 80kΩ.

X_2 is a straightforward amplifier, X_3 and X_4 are tuned amplifiers. The circuit will be recognized as the transistor equivalent of the anode-follower with a parallel-T network in the negative feedback line. The Q of these circuits is intended to be low, and for this reason, only one adjustment is necessary. The frequency response of this first part of the apparatus rises to a peak at 11c/s, falls to 0·7 of this value at 9 and 15c/s and falls to 1/10 at 4 and 25c/s. Fig. 3 shows its response to the normal e.c.g. waveform. It can be seen that the R-wave produces a heavily damped oscillation, but the T-wave produces little response, so that double-triggering on a single heart beat is avoided. If the input leads are reversed, the response is similar, though of the opposite phase

The output of the tuned amplifier is then passed to the counting section of the apparatus.

X_5 is the trigger amplifier; it is biased so as to produce an asymmetrical output. Under the circuit conditions it draws little current, and its collector is near the negative line. When its base is driven negative, the collector draws current and triggers the emitter-coupled monostable circuit, composed of X_6 and X_7

The action of this circuit which is similar to the cathode coupled multivibrator is as follows:

X_7 is biased on, and the collector current flowing through the 320Ω resistor maintains the emitters at about 0·3V negative with respect to the positive line. The base of X_6 is held at about 0·15V negative, so that X_6 is cut off. When the collector of X_5 is driven in the positive direction the voltage is passed by the 1·5μF coupling capacitor to the base of X_7 which draws less current, allowing the emitter of X_6 to drop to such a value that this transistor can draw current. Thus the base of X_7 is driven yet further in the positive direction in the familiar avalanche reaction. When the collector of X_6 has bottomed, other trigger pulses are prevented from disturbing the circuit by MR_1. The coupling capacitor recharges through the 82kΩ resistor and the circuit returns to its original state after about 150msec.

In the original design, the 'diode pump' circuit[3] was driven by a point contact transistor which was triggered between its stable states by the differentiated output of X_7. It has been abandoned in favour of a junction transistor for two reasons. In portable apparatus, the voltages that can reasonably be used are not sufficient to allow adequate design margins, so that the circuit had to be adjusted individually for each transistor. Secondly, the ratings of junction transistors have been sufficiently increased to allow their use in an output stage. The output transistor X_8 which drives the diode pump circuit is biased-off between pulses. It is driven to saturation by the output of X_7. The meter indicating the integrated output of the diode pump circuit has by the nature of the circuit a slow response, so that the onset of irregular beating of the heart, which can be important in certain procedures, may easily be missed until it is of a gross nature. Irregularities may be made quite evident by providing an audible and visual indication each time the heart beats. These indications very soon become ignored if they are regular, much in the same way that the tick of a clock soon passes from consciousness but becomes evident if the rhythm changes. The audible signal in the original instrument was made by passing the high peak current from the point circuit transistor as it charged the 6μF capacitor through the speech coil of a sensitive loudspeaker. As the present output circuit does not really produce enough current in this way to make a satisfactory noise, the circuit inside the dotted line has been adopted. It is a blocking oscillator of simple design, and makes a 'peep' each time the heart beats.

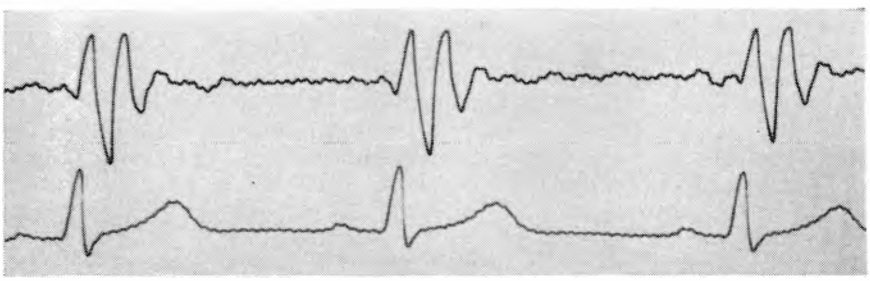


Fig. 3. Traces showing (above) the output from the tuned amplier and (below) the input waveform

Temperature has a marked effect on the performance of transistors, so that some kind of stabilization is necessary. Methods of doing this have been described[4,5,6] and are used where they are applicable. Stabilization of the monostable pulse-forming circuit is not possible by wholely conventional means. The arrangement used for this part of the circuit works as follows. In the period between pulses, the potential of the 1μF bias capacitor is determined by the potential of the collector of X_7 through MR_2. During a pulse, the capacitor changes through the base resistor of X_7, but quiescent conditions are re-established immediately after the pulse by MR_2. In this way the bias current of X_7 is determined by the potential of its collector between pulses, and stabilization takes place in a way described by Barron[6].

Setting-Up Procedure

The Tuned Amplifier

The lead is disconnected from the slider of VR_1 and connected to a source of 10c/s at 25mV. With an oscilloscope connected to the collector of X_3, VR_2 is adjusted for maximum amplitude (about 250mV). The lead is reconnected. The 0·5μF capacitor is disconnected from the collector of X_3 and 25mV injected at 14c/s with the oscilloscope connected to the collector of X_1. VR_2 is adjusted for maximum and the capacitor reconnected.

The instrument is connected to a subject whose e.c.g. is normal and whose R-wave has a value of XmV (as measured on a standard e.c.g.) with VR_1 (the main gain control) at $1/3X$ of maximum. VR_4 is adjusted so that reliable triggering just occurs. A 6·8kΩ resistor is connected in parallel with the 5·6kΩ bias resistor of X_6 which will make the circuit self-run. The frequency at which it is running is found by timing with a watch and VR_6 is adjusted until the meter (which is linearly calibrated 0 to 200 beats/min) reads this value.

This resistor with a switch may be made a permanent feature of the instrument so that its calibration may be checked at any time. The calibration is a linear function of battery voltage. The battery used (Ever Ready type AD35) has a useful life before the calibration shows a marked change, of about 40 hours.

Construction

It is unwise to use a metal case or chassis since capacitive pick-up could spoil the great freedom that battery-driven apparatus enjoys from a.c. mains interference. Apart from this the circuit may be built in practically any form. The prototype instrument weighs about 8lb and is 10in high, 7½in wide and 5in deep.

Electrodes

Needle electrodes in the skin of the chest are very suitable for anæsthetized patients. Otherwise the normal e.c.g. electrodes may be used. In most locations satisfactory operation may be obtained by the subject dipping each hand into steel bowls of tap water to which ordinary clipped leads are attached. However, this procedure is not recommended except as a demonstration of the instrument's capabilities in working under difficult conditions and is useful as a form of marginal testing.

Operation of the Instrument

The procedures for adjusting the gain control for optimum results is as follows. The gain control is advanced very slowly from zero until reliable triggering is just obtained. The setting is noticed and the gain control is advanced until the triggering becomes irregular. This second reading is noted and the control is set halfway between these limits. If the range seems restricted the leads are reversed, with the lead reversing switch and the procedure repeated. In practice the gain control is set at half maximum when the instrument usually works well. It is only under poor conditions that the full procedure has to be followed.

Finally, as the instrument has a fairly high input impedance it may be connected in parallel with a standard e.c.g. without noticeably affecting the appearance of the trace.

REFERENCES

1. Boyd, W. E., Eadie, W. R. Continuous Recording of the Human Heart Rate. *Electronic Engng. 26*, 330 (1954).
2. Molyneux, L., Pask, E. A. A Transistor Pulse Counter. *Brit. J. of Anaesthesia. 27*, 262 (1955).
3. Earnshaw, J. B. The Diode Pump Integrator. *Electronic Engng. 28*, 26 (1956).
4. Shea, R. F. Transistor Operation. Stabilization of Operating Point. *Proc. Inst. Radio Engrs. 40*, 1435 (1952).
5. Sulzer, P. G. Junction Transistors Circuit Applications. *Electronics 26*, 170 (Aug., 1953).
6. Barron. Simple Circuit Stabilizes Transistor. *Electronics 27*, 182 (1954).

An Electronic Measuring System for a Jig Borer

By C. R. Borley*, B.Sc.

The article describes a measuring system for a high accuracy machine tool. The system is designed to ease the setting of the machine table, to reduce setting errors and to improve the repeatability of setting accuracy. An improved master scale is employed.

THE measuring system has been developed jointly by the Coventry Gauge and Tool Co. Ltd and the Mullard Research Laboratories. An existing machine, the 'Matrix' No. 50 jig boring machine has been modified to receive the system together with extensive push-button controls for the table and spindle motions. In common with the optical measuring system currently used in the No. 50 machine, the electronic measuring system is a fine setting arrangement. The operator positions the table within 0·1in on a scale divided to tenths of an inch fixed to the table of the machine, and then refers to the fine setting system to set to tenths of a thousandth of an inch. The master scale has divisions 0·1in apart, a digital interpolation system being used to divide any selected interval on the master scale into 1 000 parts. Three decade switches are provided for each ordinate to allow the last three figures of the desired position to be set into the system. The error between the desired and actual position is displayed to the operator in numerical form. At present the table is driven manually to reduce the error to zero. The form of the input and display have been chosen to facilitate later development to automatic positioning in response to numerical information.

The Master Scales

The internal standards of length take the form of steel buttress scales. (Fig. 1.) Each buttress scale consists of a solid steel bar having a sawtooth form cut into it of accurate 0·1in pitch. A step, machined lengthwise, reveals the form, consistency of upright and sloping edges. The sawtooth form is obtained by cutting a sawtooth screw into the bar. The upright edges provide references at 0·1in intervals with a maximum adjacent pitch error of 3×10^{-5}in. The accumulated pitch error is such as to provide measurement to jig boring machine standards. The upright edges have been stellited to provide extra hardness and corrosion resistance. The scale is very rugged and stable, being machined from a single piece of material. It has the same coefficient of expansion as the machine.

The Interpolation System

Interpolation between the edges of the buttress scale is accomplished optically, the principle being shown in Fig. 2. The buttress scale is attached to the machine table and is parallel to the direction of motion of the table. The axis of the optical system is normal to the plane of the edges, the interpolating image being brought to a focus in this plane. The glass interpolation scale is a little over 2in long. A vertical opaque reference edge is provided at one end. The upper half of the rest of the scale is clear, the lower half having somewhat over 1 000 alternate opaque bars and transparent spaces evenly spaced. The bars and spaces are each ·001in wide. The lens images a length of 1 000 bars down to 0·1in length in the plane of the buttress scale edges with the bars parallel to the vertical edges.

The only illumination provided at the interpolation scale is a moving line of light ·001in wide, which moves across the scale, keeping parallel to the bars, light passing between the gaps one at a time. In the upper clear half of the scale, the line of light is visible throughout the scan. The line scans the scale repetitively from end to end at sensibly constant speed, there being no flyback. Two photocells are used. One photocell, in conjunction with a mirror, receives light only from the lower divided scale, and provides a train of pulses somewhat over 1 000 in number during each scan. A second photocell, placed behind the buttress scale, receives light via the reducing lens from the top clear half of the interpolation scale only. The waveform available from this cell for a few typical buttress scale positions is shown in Fig. 3 together with interpolation pulse trains. The scan is unidirectional, starting at the reference edge. Sharp rises of voltage at the buttress scale photocell only occur due to the obscuration provided by the vertical buttress scale edges. These sharp rises are detected and used as a gating signal operating upon the interpolation waveform. Since the two waveforms bear a fixed relationship in space as well as in time, the gated pulse train contains a number of pulses equal to the number of tenths of a thousandth of an inch in the fraction of 0·1in displacement of the buttress scale, independent of scanning speed.

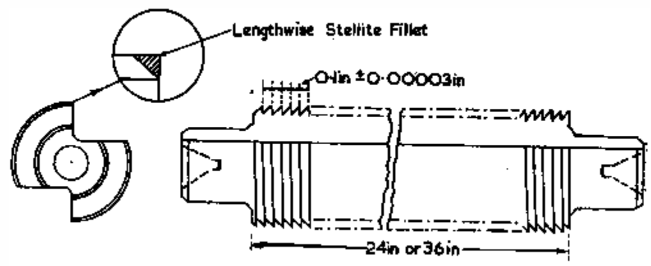


Fig. 1. Steel buttress scale

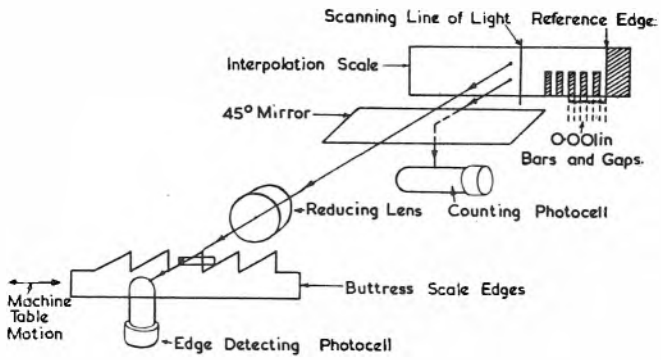


Fig. 2. Principle of interpolation

The Input, Counting and Display Arrangements

A block diagram of the system required for the measuring system is shown in Fig. 4. The scanning is repetitive at 4 scans/sec, with a dwell period of some 30 per cent

* *Mullard Research Laboratories.*

between scans. At each scan a fresh measurement of buttress scale position is made, independent of previous measurements. Three decade counters are provided, each of the four-stage hard valve binary type with feedback to reduce the radix of counting from 16 to 10. The feedback is such that the first ten consecutive states are used. Each counter may be set to any of the ten states from a 10-position switch using a matrix to convert from decimal to binary coded decimal. The last three figures of the required position are set on these switches. After the gated pulse train has been passed to the counter, the number left in it can be passed, via matrices converting from binary to decimal, to a register consisting of cold-cathode tubes.

To describe the cycle of events during and immediately after a scan, it is assumed that the counters have been set to the number set up on the switches and that the gate is open. When the scan begins, the pulse train from the interpolation scale photocell is passed through the gate to the decade counters. The connexions between the decades, between the stages in each decade, and the feedback are arranged so that the incoming pulses subtract from the total. The subtraction proceeds until the image of the line of light is obscured by a vertical edge on the buttress scale, when the sharply rising photocell waveform generates a signal which closes the gate. During the rest of the scan no signals occur which can open the gate and allow any further pulses to reach the counter. At the end of the scan, therefore, the counter will contain the difference between desired position and the actual position.

If during counting, the counter is emptied, receipt of a further pulse will cause it to return to a full count of 999, further pulses counting down from this figure. The convention is adopted that if, when counting is complete, the number stands at 0 to 499 inclusive, this will be an error of that magnitude and of positive sense. If the number stands at 500 to 999 inclusive then the error is negative and the magnitude will be the complement of the number with respect to 1 000, 999 meaning an error -1, etc. The end of the ungated pulse train provides a signal for the number in the counter to be passed to the error register where the error standing from the previous scan is now rejected in favour of the new value. When the error sampling and registration are complete, the counters are reset to the values set on the dial switches and the gate is opened to await the pulse train of the next scan. No information is carried over from scan to scan and in consequence, the system reports machine position and not displacement. The error is displayed at intervals of 10^{-4}in for 3×10^{-4}in either side of zero by individual display lamps. Larger errors are grouped to provide rough indication of error magnitude. In units of 10^{-4}in, the groups are either side of zero 4 to 6, 7 to 9, 10 to 49, 50 to 99 and 100 to 499 inclusive. In all, 17 neon lights illuminating numbered screens are provided, arranged in a horizontal line with zero in the centre.

Fig. 3 shows the waveforms produced for different positions of the buttress scale. The third example shows the condition obtained when control is about to pass from one buttress scale edge to the next. The travel of the line at the buttress scale is slightly more than 0·1in so that there is a small range of buttress scale positions in which two edges are encountered. Two gate closing signals are generated, one near the beginning and one near the end of the scan. As the length of scan from the reference edge to the first buttress scale edge is reduced, there will come a time when insufficient light is transmitted to generate a gate-closing signal at the beginning of the scan. The gate will now remain open for the whole scan. At this changeover point the gated pulse train will increase abruptly from 1 or 2 pulses to 1 001 or 1 002 pulses respectively, since the optical reduction has been adjusted to fit a length of scan equal to just 1 000 bars between two consecutive edges. The counter has a counting capacity of only 1 000 however, and the number left in the counter is the same whether x or $(1\,000+x)$ pulses are applied to it. Hand-over

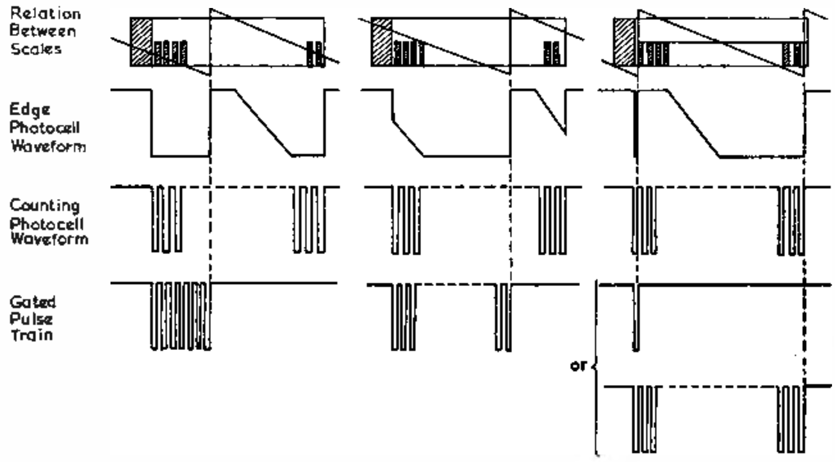


Fig. 3. Interpolation scale and waveform relationships

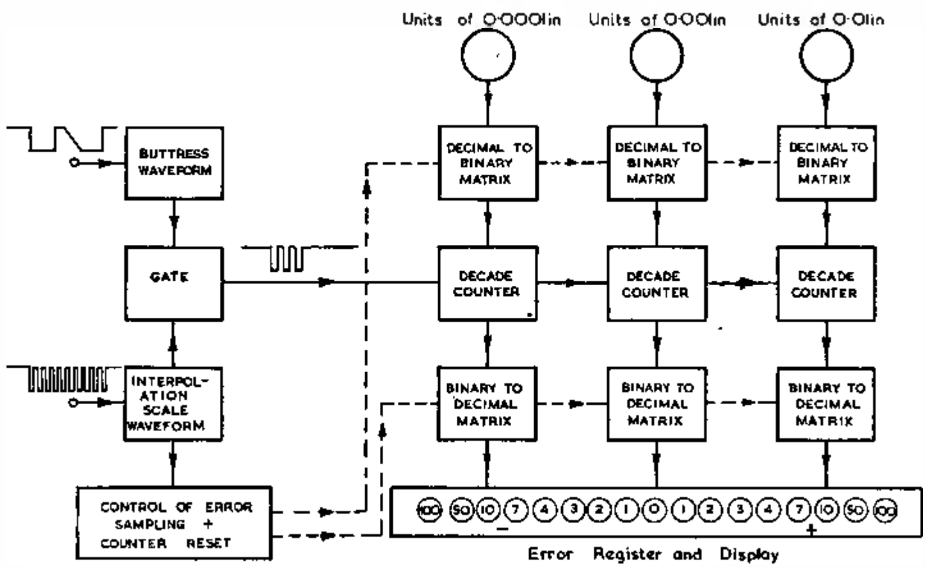


Fig. 4. System block diagram

from edge to edge occurs without a break in indication, therefore.

The Scanning Mechanism

A mechanical flying line scanner has been designed to provide the scanning line of light for the interpolation system. The scanner must, as far as possible, match the master scale in simplicity, ruggedness and long life. The scanning line must be ·001in wide as a maximum in order to resolve the bars. It must remain in focus over the whole travel and must remain parallel to the bars within ±9 minutes of arc. The direction of the emergent cone of rays after the formation of the line image at the interpolation

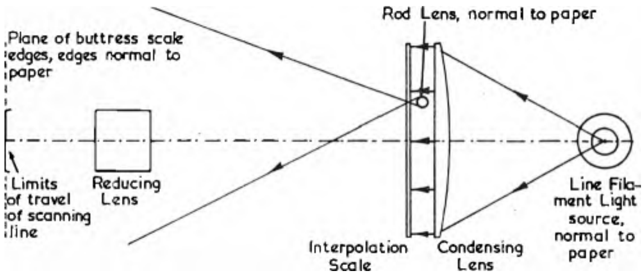


Fig. 5. Optical system of the scanner

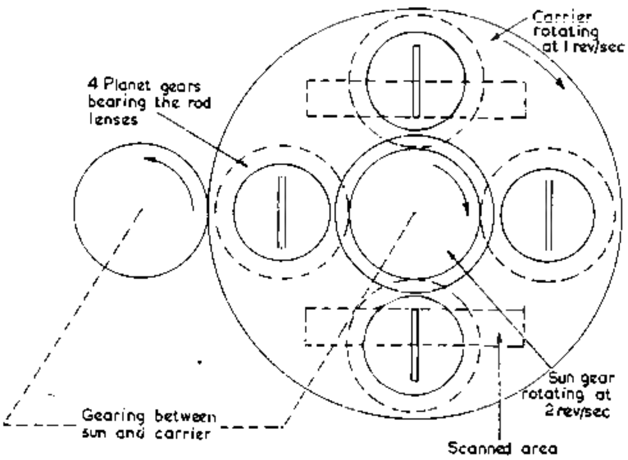


Fig. 6. Mechanical system of the scanner

scale should remain normal to the scale. This ensures that the aperture of the reducing lens is kept fully illuminated across the scan. Optical systems involving rotating lenses, mirrors and the like were examined but none was suitable.

The solution adopted was to move the line image by moving the lens producing it. Fig. 5 shows the optical arrangements in plan view, the bars on the interpolation scale being normal to the paper. The light source is a line filament prefocus 36W headlamp bulb. A condensing lens provides a beam of parallel light some 2½in wide to illuminate the scale over its entire length. A glass rod lens ⅜in in diameter, parallel to the bulb filament forms a line image parallel to the bars on the scale. The rod is stopped down to provide a line image ·001in in width. The rod is then moved across the scale at constant distance from it, the line image illuminating each of the bars in succession. A screen either side of the rod moves with it, shielding the rest of the scale.

The motion of the rod lens is provided by mounting it in the planet gear of an epicyclic gear train (Fig. 6). If the sun and carrier gears are driven at the correct relative speeds, the planet gear, while moving round the sun, does not rotate about its own axis. With large diameter ball-races for the planet bearings, the centre of a gear can be bored out and the rod lens mounted at right-angles to the gear axis and parallel to the bars on the interpolation scale. The rod lens moves in a circular path, remaining upright, and by using a rod somewhat longer than the depth of the scale and by using 60° of carrier motion for the scan, the scale is fully scanned. Four planet gears and rod lenses are set in the carrier which rotates at 1 rev/sec driven through reduction gear from a three-phase induction motor giving 4 scans/sec.

Application to Jig Borer

The Matrix No. 50 jig borer is an open-sided machine having compound table motions (Fig. 7). It is convenient therefore to mount a single scanner in the intermediate table and arrange that this co-operates with two buttress

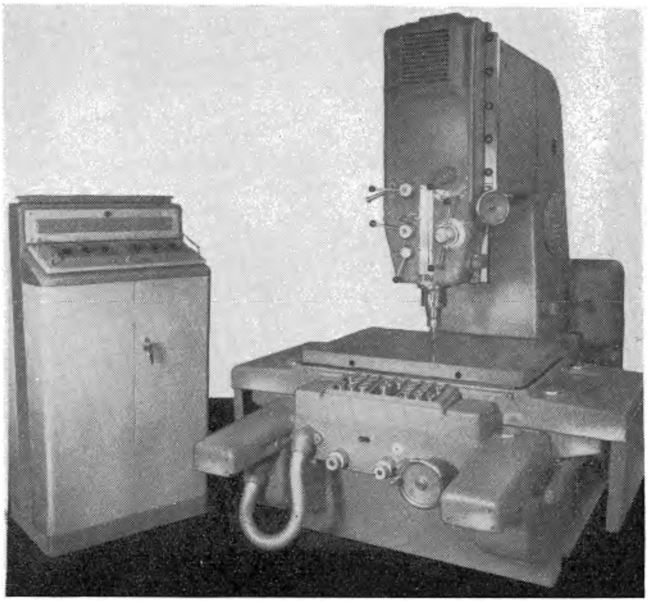


Fig. 7. 'Matrix' 50 jig boring machine, with electronic console on the left

scales, one fixed to the top table and one to the base. The arrangement is shown schematically in Fig. 8 and mechanically in Fig. 9. A common scanner can be used since the top and bottom 60° arcs of carrier motion provide a scan.

The scanner has a single 12V 36W axial filament bulb as the common light source, two 45° mirrors illuminating the upper and lower interpolation scales. The upper optical system provides interpolation for the transverse machine motion of 36in travel. The upper system has a 45° mirror reflecting light upward to the reducing lens, buttress scale and photocell, as the separation between the buttress scales is greater than that between the interpolation scales.

The photocells used are nine-stage photo-multiplier cells, necessitated by the low light level at the buttress scale after the reductions due to the apertures of the rod lenses and the reducing lenses. It is convenient to use the same type of cell at the interpolation scale owing to the large output that can be obtained. The supply for the photocell dynodes is obtained from a chain of seven 85A3 sub-miniature voltage reference tubes and three 85A2 miniature reference tubes.

This supply chain, and cathode-follower head amplifiers for all four photocells are built into the machine body.

The console (Fig. 7) houses two sets of input dials, one set for longitudinal and one set for transverse directions, with a single set of counting and display circuits which are switched to either table motion as desired. The only other controls provided at the console are the on-off buttons. The machine is controlled from a push-button panel at the front of the intermediate table.

A feature of the photo-multiplier cell is the reduction of output which occurs in the first few hours of operation. This is a fatigue phenomenon due to exposure to light, recovery occurring overnight. In the present circuit, compensation for this effect is provided by taking advantage of the fact that both photocell outputs are amplified and

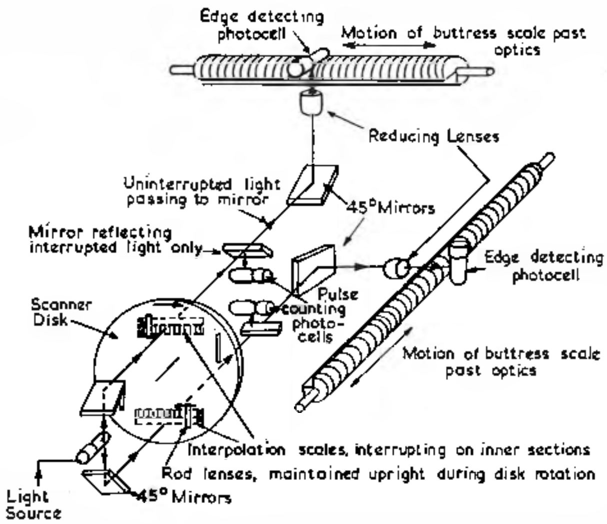


Fig. 8. Schematic arrangement of two-co-ordinate measuring system

squared to provide sharp-edged pulses. The d.c. level of the photocell waveform is caused to vary in proportion to the total amplitude in such a way that the half amplitude level of the waveform remains at the level at which squaring occurs. The total amplitude is sampled at each scan and stored in a long time-constant circuit which provides the variable level for the photocell waveform. Thus, the same point on the waveform is squared regardless of amplitude changes.

Accuracy of the System

Two factors affect the accuracy of fine setting. Table motion of nearly 10^{-4}in can occur without a change in pulse count, and hence in indicated position, occuring. The table cannot be set to better than $\pm 5 \times 10^{-5}$in on this account unless the table is always set at one end of the region of no change, where the pulse count fluctuates by one from scan to scan. This region of fluctuation is about 2×10^{-5}in wide. Secondly, the distortion introduced by the reducing lens will affect the linearity of interpolation. The optical path length is adjustable and precisely 1 000 bars can be fitted between the two adjacent buttress scale edges. The

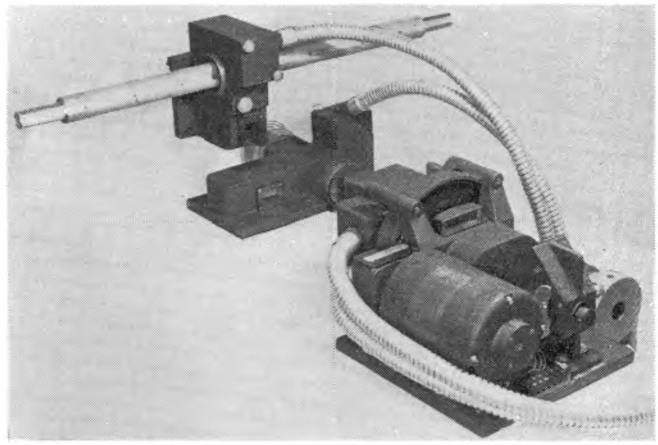


Fig. 9. Physical arrangement of two-co-ordinate measuring system

lens distortions will be symmetrical about the axis and hence interpolation distortion will have a sine-wave form, being zero at the ends and in the centre of the field of view, but having positive and negative errors either side of the centre. These errors are less than $\pm 5 \times 10^{-5}$in. In normal use a fine setting accuracy of better than $\pm 10^{-4}$in can be obtained.

Acknowledgments

The author wishes to thank the manager of Mullard Research Laboratories and the directors of the Coventry Gauge and Tool Co. Ltd and Mullard Ltd. for permission to publish this article. Also he is indebted for help received from many colleagues, especially Mr. R. W. Fenemore who initiated the project.

Telephone Repeaters Using Transistors

The rapidly increasing size and complexity of telephone networks make the use of the two-wire repeater particularly attractive. With its aid, the performance of existing cable routes is considerably improved and the call for the relatively expensive four-wire circuit diminished. Also, where new cable routes are planned, the inclusion of two-wire repeaters may well permit the use of smaller cable conductors.

An order for transistorized two-wire repeaters has recently been placed with Standard Telephones and Cables Limited by the Irish Post Office for use on loaded junction cables in their telecommunications network.

In this repeater the problems of power supplies and maintenance have been greatly simplified by the use of transistors to replace thermionic valves as the amplifying elements. Junction-type transistors are employed, the power consumption is very small and their expected long life will cut maintenance to a minimum.

The total power required by the repeater is less than ¼W and may be obtained from existing auto-exchange batteries of 24 or 50V or from mains-operated power units. Provision is made for independent adjustments of gain in either direction up to a maximum at 14dB, and the peak audio output to line is 25mW. The transmitted frequency range is 300 to 3 400c/s, conforming to C.C.I.F. recommendations.

Built-in with the repeater are the line coils, line and switchboard balancing networks, equalizer and signalling by-pass circuit, to form a robust self-contained unit.

The size of the repeater, one-half of a standard 2-unit panel, i.e. 3½in by 9½in by 7½in, is mainly dictated by the signalling and ringing facilities to be provided. In this case the signalling path is of low resistance and will transmit all dialling or ringing signals up to 50c/s; during speech a line current of 50mA may also be passed.

A total of 28 repeaters, together with testing equipment and power filter panel, will mount on one single-sided 6ft rack, making a convenient and compact assembly.

As an accompaniment, the testing equipment developed for use with the repeaters, includes a single-frequency 800c/s oscillator with an output of 1mW, which also employs a junction-type transistor.

A Foetal Pulse Rate Recorder

D. H. Smith*, M.Sc.

The article describes an apparatus which gives a continuous record of the foetal pulse rate and can be used under normal clinical conditions.

ONE of the indications of the condition of the foetus before and during labour is its pulse rate. It is usually estimated by listening with a stethoscope to the heart sounds transmitted through the maternal abdomen. This is not always easy to do, because of the very low intensity of the sounds. Moreover, during labour both the trend and variability of the pulse rate are important and it is not always convenient to make frequent checks. Since the 1920s research workers in many centres have used microphones and amplifiers to broadcast the foetal heart sounds through headphones or loudspeakers. As techniques improved, the sounds were recorded on acetate disks for demonstration purposes[1]. For research work visual records were made, first on photographic paper and later on direct writing pen recorders[2,3,4,5]. To date no paper describing apparatus which will record directly the foetal pulse rate, has been published. This is not surprising for it is apparent from a search of the literature that not until about the last five years had a good enough signal-to-noise ratio been achieved to make continuous automatic counting possible.

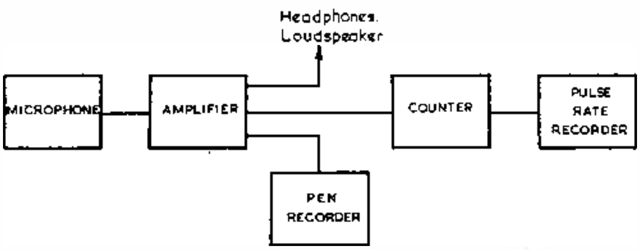


Fig. 1. Block diagram

When a continuous record of the pulse rate of a child or adult is required, it is usual to utilize the electrical pulse produced by each beat of the heart, i.e. the electrocardiac potential, for counting. This is much more convenient than using the heart sounds, for a number of reasons. But in spite of the striking advances which have been made in foetal electrocardiography in recent years, the difficulty of suppressing the maternal e.c.g. makes it impractical at present to use the foetal e.c.g. for pulse counting, so attention has been concentrated on the heart sounds.

In this work, the aim was to produce an instrument which would be useful in day-to-day clinical routine. Though this aim has not yet been fully achieved, the experimental equipment described here has given very encouraging results.

General Description

The block diagram is shown in Fig. 1. A sensitive microphone is pressed firmly into the maternal abdomen and held by a strap. Screened cable feeds the microphone output into a high-gain, narrow band amplifier. The amplifier signal operates a specially designed counter which gives an output current that depends on the pulse rate. This current is normally displayed on a 0 to 1mA Esterline Angus recorder which is calibrated directly in pulses per minute.

Two additional outputs from the amplifier can be used for (1) continuous listening to the sounds by headphones or loudspeaker (2) making records on a direct-writing pen recorder.

Microphone

The main requirement is that the microphone should be sensitive enough to produce a usable signal from the foetal heart sounds, and yet be very insensitive to all airborne sound. It seems obvious that a sensitive vibration pick-up should be suitable, and the Brush type DPIH[6] has been used.

The microphone is located on the area where the sounds are loudest, and is held in position by a 4in wide Elastoplast strapping (elasticized abdominal belts were found to be more efficient, but are unsuitable for use during labour). There is no interference from the maternal heart sounds, and in the twin pregnancies so far encountered, it has been easy to pick up a signal from each foetus, without interference from its twin.

Amplifier

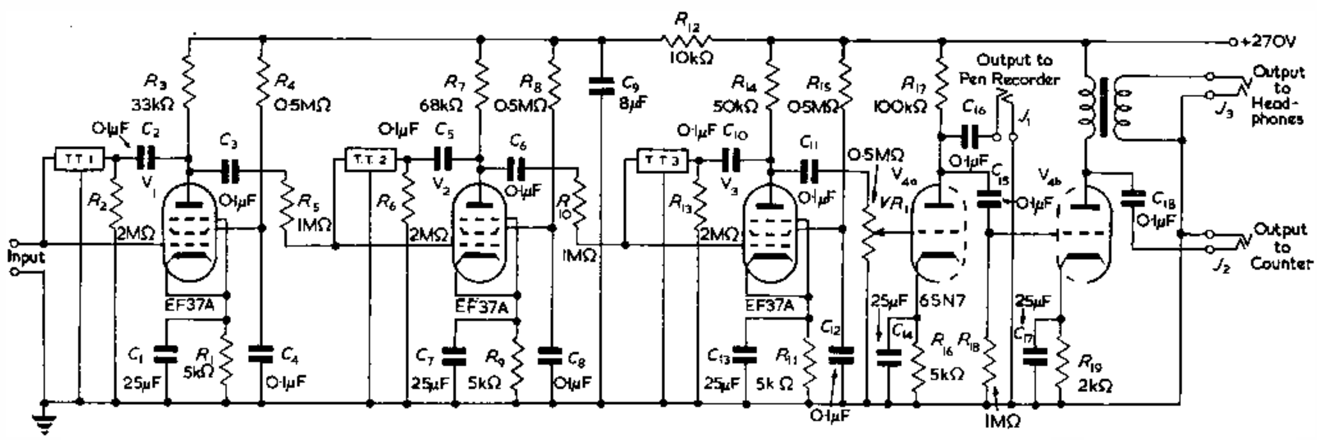
Normal heart sounds have two main components. The first, or systolic, has a fundamental frequency in the range 60 to 100c/s, and the second, or diastolic, has a fundamental in the range 40 to 80c/s[6]. The foetal heart can be expected to produce similar sounds, and, in the absence of information on the transmission characteristics of the intervening layers of fluid and tissue, these figures were taken as a guide. The design was influenced also by the desirability of avoiding interference from the mains frequency, 50c/s, and its harmonics. Hence a pass-band of 60 to 80c/s was chosen, and in order to reduce still further the effect of airborne noise, and to discriminate against interference from digestion noises, uterine souffles, etc. (which have wide frequency spectra) the amplifier was designed to have a steep-sided, flat-topped characteristic.

Fig. 2 shows the circuit diagram. The first three stages form a band-pass filter, with a frequency response as given in Fig. 3. This response is achieved by using three stagger-tuned feedback stages, each feedback circuit containing a twin-T *RC* network. The components of these networks have been chosen so that in each case there is one frequency at which the feedback signal is in phase with the input. Provided the stage gain is just too low for oscillation to occur, sharply tuned circuits are obtained[7,8].

The 1MΩ resistors in series with the coupling capacitors result in a loss of about 50 per cent of the signal at each stage, but they are necessary to prevent the twin-T networks being loaded by the relatively low anode loads of the preceding stages. Such loading would cause a serious loss of selectivity.

Two triode stages complete the amplifier. The voltage gain available (within the pass-band) at the anode of the final stage is $>10^6$, and it has been found adequate for all cases so far encountered. As the response curve would lead one to expect, impulsive sounds, such as heart beats, cause the amplifier to 'ring' and the type of signal obtained at the amplifier output is shown in Fig. 4. There is no more difficulty about counting such signals, than there would be in counting undistorted signals from the foetal heart.

* *The University, Manchester, formerly D.S.I.R., New Zealand.*



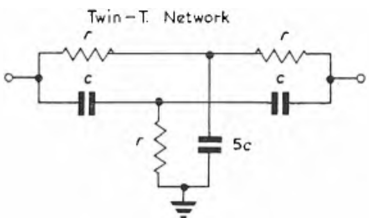


Fig. 2. Amplifier

Twin-T components

T.T.1	*tunes to 68c/s.*	$R = 1M\Omega$.	$C = \cdot 00154\mu F$
T.T.2	*tunes to 59c/s.*	$R = 1M\Omega$.	$C = \cdot 00177\mu F$
T.T.3	*tunes to 78c/s.*	$R = 1M\Omega$.	$C = \cdot 00133\mu F$

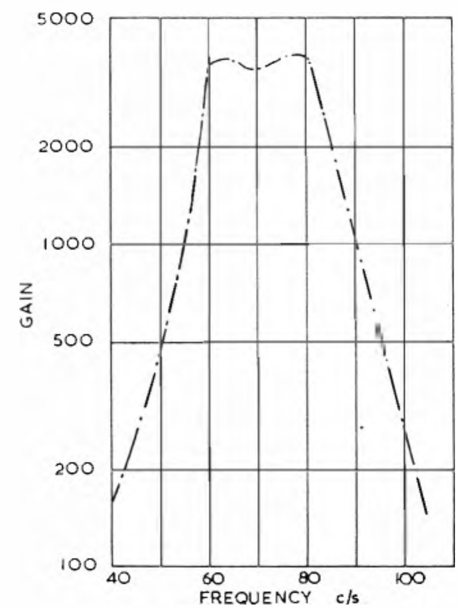
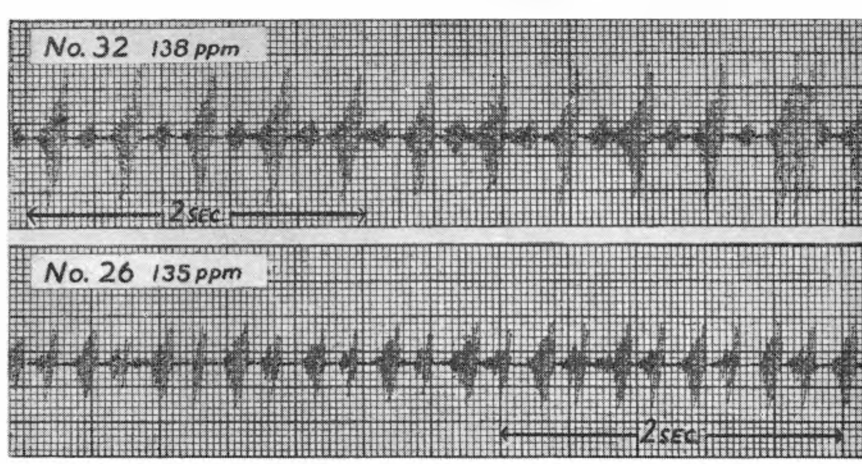


Fig. 3. Frequency response curve of band-pass filter

Fig. 4. Foetal heart sounds recorded on a high-speed pen recorder

In No. 32 one sound is of much greater amplitude than the other

In No. 26 the two sounds are of approximately equal amplitude



Pulse Rate Counter

The foetal heart averages about 140 pulses per minute, but rates in the range 110 to 170 pulses/min are quite normal. During labour, the obstetrician is particularly interested in rates below 100 as these are regarded as sure signs of foetal distress. On rare occasions, rates of 180 and above may occur and these, too, are indicative of possible distress. So it is necessary for the counter to be able to handle a range of an octave or more. Now, as Fig. 4 shows, it is possible for the two components of the heart sound to be of equal amplitudes, and therefore the possibility of double-counting must be guarded against.

The circuit of the counter is shown in Fig. 5. The operation is as follows: If there is no input signal, C_{26} (and hence C_{29}) charges up negatively to a potential sufficient to reduce the anode current of V_8 to zero. Each heart beat causes a single pulse of current to flow through the thyratron V_7. The 'normally-open' contacts on the relay in the cathode of V_7 close momentarily, discharging C_{26} very rapidly. C_{29}, C_{28} and C_{27}, through their respective series resistors, begin to share their charges with C_{26}, so that the potential at the grid of V_8 becomes more positive. Meanwhile, as soon as the relay contacts open C_{26} begins to charge up negatively again. It is clear that when a train of regularly spaced pulses is arriving at the grid of V_7, the potential on C_{29} and hence the anode current of V_8, will reach an equilibrium value dependent on the repetition rate of the pulses. A recording milliammeter in the anode circuit of V_8 can then be calibrated directly in pulses per minute.

The upper and lower limits of the scale can conveniently be set by adjustment of the potentiometer VR_1 and capacitor C_{26} respectively. The scale which was normally in use was calibrated from 75 to 180 pulses/min. Switch S_2 provides a choice of response time, since the time for the potential across C_{29} to reach a new equilibrium value, after a change in pulse rate, will

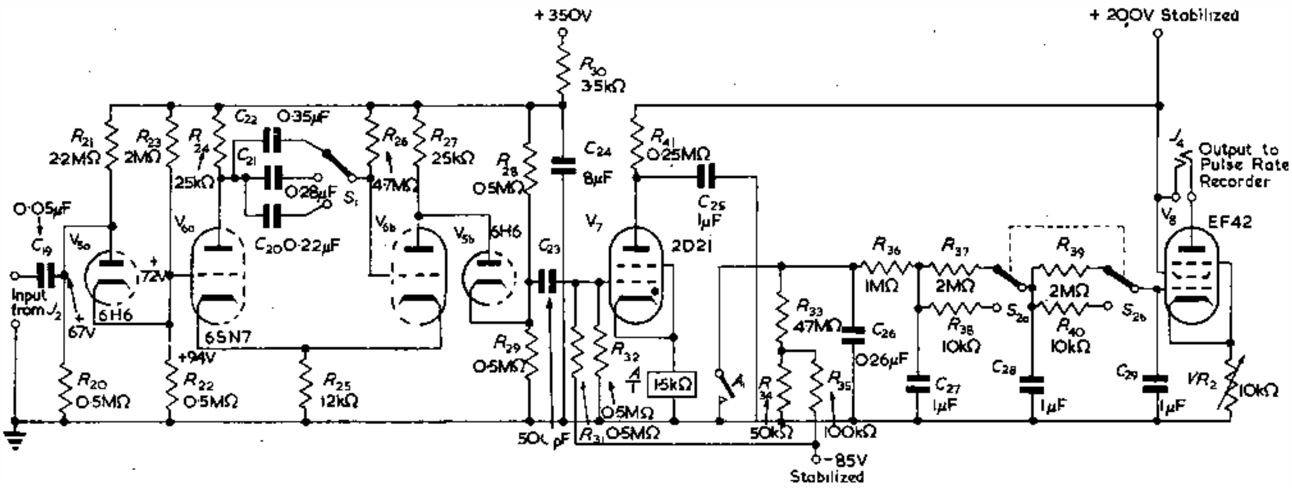


Fig. 5. Counter circuit

be dependent on the value of the series resistors R_{36}, R_{37} . . . etc. The components shown provide response times with a ratio of about 5 to 1.

The univibrator circuit (V_5 and V_6) provides a pulse of constant shape and amplitude for operating V_7. It was found unnecessary to stabilize the 350V supply, but a constant voltage transformer was used to minimize the effect of fluctuating mains voltages. The action of the univibrator is well known[9]. It is insensitive to any pulses arriving during the time in which it is in the 'quasi-stable' state (i.e. when V_{6a} is conducting, V_{6b} cut off). This state is initiated by the arrival of a positive pulse at the grid of V_{6a}, and its duration is controlled by the time-constant of R_{26} and the coupling capacitor C_{22}. Thus the circuit is completely insensitive for a pre-determined period after the arrival of a pulse. In order to cover the required range, with adequate margins of safety so that no pulses were missed, and no double-counting occurred, three coupling capacitors were used. Their ranges of usefulness are: C_{22}, 75 to 115 pulses/min; C_{21}, 100 to 155 pulses/min; C_{20}, 130 to 190 pulses/min. Switch S_1 allows the appropriate one to be selected when recording commences.

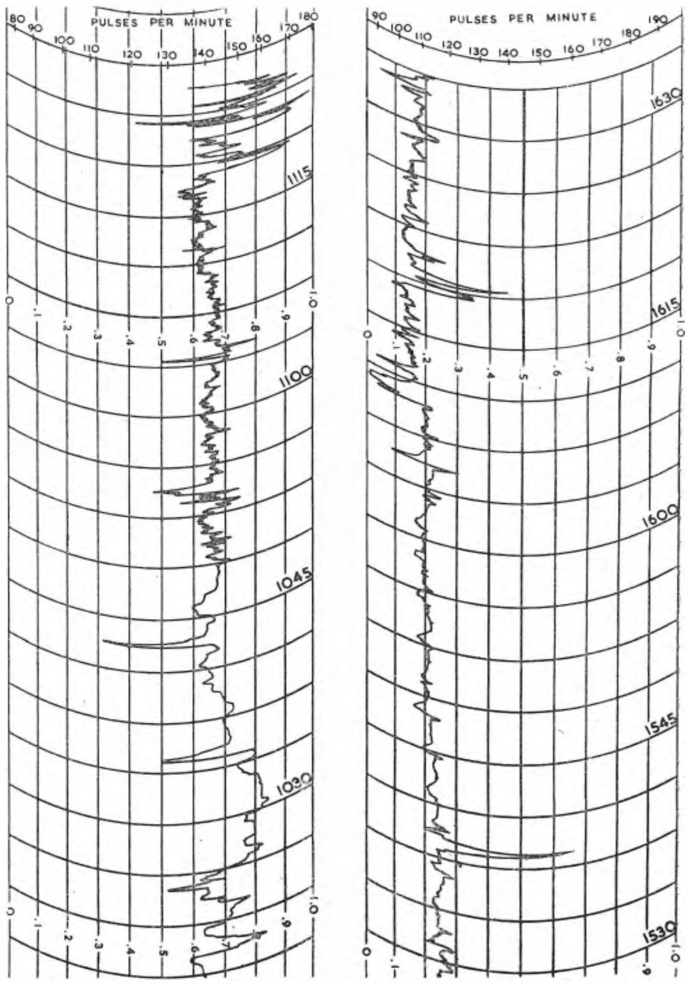
The counter has been found very stable in operation and has performed satisfactorily. It may be used with an Esterline Angus recorder or it can be connected directly to one channel of an Evershed & Vignoles double pen recorder to enable the foetal pulse rate to be recorded simultaneously with uterine pressure. Useful information on the progress of labour can then be obtained.

Results

On pre-natal patients good results are always obtained. The only serious source of interference is from the rubbing of clothes, blankets, etc., against the back of the microphone. Intermittent interference of this kind does not have any significant effect on the pulse rate record. Spurious pulses from foetal movements are of too short a duration to cause trouble. Fig. 6(a) shows a typical record. The effect of switching from 'slow' to 'fast' responses at 1050, is very marked. The rapidly fluctuating pulse rate during the last few minutes of the record was caused by violent foetal activity. A rather less violent period of activity occurred at the beginning, but the effect is masked by the slow response.

During the course of a prolonged labour the instrument may be working under extremely difficult conditions. If the patient is in acute discomfort, the interference from incessant movements may spoil the record. If the patient turns into a nearly prone position the pressure of the microphone on the abdomen may be very much reduced, and the amplifier output may be too low to operate the counter. In spite

Fig. 6(a) (left). A typical record of foetal pulse rate
(The sudden drops at 1035 and 1043 were caused by instrumental artefacts)
(b) (right). Pulse rate record made during labour



of these difficulties it is possible to obtain records which give an accurate indication of the trend of the pulse rate. Moreover these records may, as shown in Fig. 6(b), contain sections of an hour's duration which are quite free from the effects of interference.

Conclusion

The instrument represents an advance in that it gives, for the first time, a continuous record of the foetal pulse rate. It can be used under normal clinical conditions, and is not affected by the high noise level usually encountered in hospital wards and labour rooms. The only limitation is that careful supervision is necessary when the instrument is used during labour, so that erroneous variations in rate caused by maternal movement can be identified. Improved performance depends on the possibility of modifying the microphone assembly so that it becomes less sensitive to the noise and vibration generated by maternal movement.

Acknowledgments

The author wishes to acknowledge the help and encouragement of Professor H. M. Carey, Post-graduate School of Obstetrics and Gynaecology, Auckland University College, Medical Director of the National Women's Hospital, Auckland, N.Z., for whom this instrument was developed. He also thanks his colleagues at the Auckland Industrial Development Laboratories, for their help and advice, especially Mr. J. B. Cornwall who designed the thyratron counting circuit.

REFERENCES

1. POMMERENKE, W. T., BISHOP, F. W. Amplification of Fetal Heart Sounds. *Amer. J. Obstet. Gynaec. 35*, 851 (1938).
2. SMITH, A. L., HERVERT, W. J. A Method for Recording and Reproducing Fetal Heart Sounds. *Amer. J. Obstet. Gynaec. 40*, 102 (1940).
3. STEER, C. M., HERTSCH, G. J. Continuous Observation of The Fetal Heart. *Amer. J. Obstet. Gynaec. 62*, 1139 (1951).
4. WOOD, M. C. The Amplification and Recording of Foetal Heart Sounds. *Electronic Engng. 25*, 90 (1953).
5. SOUTHERN, E. M. Electrocardiography and Phonocardiography of the Foetal Heart. *J. Obstet. Gynaec. Brit. Emp. 61*, 231
6. OLSEN, H. F. Elements of Acoustical Engineering. 2nd edition. 335 (D. Van Nostrand, New York, 1947).
7. HYDE, F. J. Highly Selective Amplification at Low Frequencies. *Wireless Engr. 29*, 85 (1952).
8. SMITH, D. H. The Characteristics of Parallel-T RC Networks. *Electronic Engng. 29*, 348 (1957).
9. SAYRE, D. Waveforms. M.I.T. Radiation Laboratory Series. 1st edition 19, (McGraw Hill, New York, 1949).

Demonstration of Wideband Radiocommunication

A demonstration and display of wideband radio communications systems and techniques was recently given by Marconi's Wireless Telegraph Company Ltd at their Development Laboratories at Writtle. In addition there were manufacturers of such specialist apparatus as multichannel telephone equipment, diesel alternators, aerials, aerial-supporting towers, electronic test gear, travelling-wave tubes and other thermionic devices, thereby collectively providing for every possible requirement on wideband telecommunications systems.

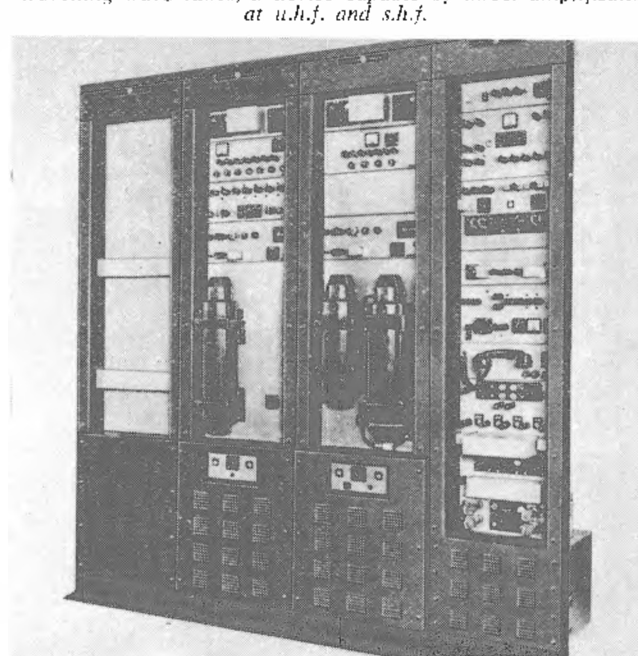
The demonstration was arranged in two main sections; firstly, the working display and, secondly, a static display of equipment and accessories.

In the working display the v.h.f., u.h.f. and s.h.f. multichannel equipment was shown under actual operational conditions, in which the signals were routed over cross-country paths, with repeater stations, before returning to the terminals. Of particular interest was the Marconi type HM.200/250 travelling-wave tube wideband radio link, which was being used to relay colour and monochrome television pictures over a radio-path of some 30 miles, as a practical means of showing the equipment's capabilities in the matter of handling complex wideband signals. A mobile s.h.f. multi-channel system with military applications and a v.h.f. single-channel subscriber's system were also focal points of interest. The latter equipment, manufactured by Automatic Telephone and Electric (Bridgenorth) Ltd served to show how a subscriber located in a remote area where no telephone lines exist, can be connected to the nearest exchange by a simple radio link which provides the subscriber with a telephone service equivalent to that expected from a direct physical connexion.

The static display, besides featuring radio and carrier equipment, included a selection of test instruments used in the development, installation and maintenance of multichannel links, while other stands were devoted to displays illustrating aerial towers and aerials designed for use with radio links and suitable sources of power for unattended stations. A selection of travelling-wave tubes, voltage stabilizers and rectifiers was also on show.

Today, radio telecommunication offers a very attractive alternative to cable or open-wire systems, particularly across undeveloped areas where the terrain is rugged and inhospitable. A modern wideband radio link can, for example, provide a service of up to 600 simultaneous telephone conversations or a much greater number of telegraph or teleprinter channels over distances of up to 2 500km; alternatively, television signals can be carried. A radio system can usually be expected to cost less than a comparable cable system, and at the same time is far less vulnerable to storm damage or sabotage, while providing an equal degree of privacy in the transmission.

The Marconi type H.M.200 u.h.f., f.m. wide-band multi-channel terminal equipment for point-to-point telecommunications. This operates in the 1 750-2 500Mc/s frequency band and offers advantages of broad modulation bandwidth and high aerial gain. The H.M.200 is designed to provide main trunk routes, with an ultimate capacity of up 600 telephone channels or one high definition television channel. It utilizes travelling wave tubes, a device capable of direct amplification at u.h.f. and s.h.f.



The Junction Transistor as a Computing Element

(Part 3)

By E. Wolfendale*, B.Sc., A.M.I.E.E., L. P. Morgan*, B.A., and W. L. Stephenson*, B.Sc.

Examples of the Use of Transistors in Computing Circuits

A Decade Counter

One type of decade counter, which has been investigated, uses a binary followed by a 5 digit ring counter, Fig. 25. The count is indicated directly on a meter which sums the currents from each bistable circuit. Using junction transistors with an α cut-off of approximately 500kc/s,

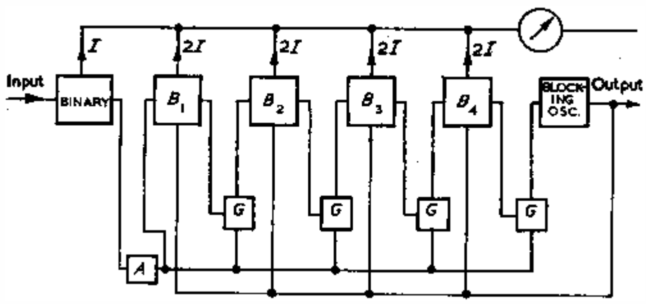


Fig. 25. Block diagram of decade counter

TABLE 4
Counting sequence for the decade counter.

NUMBER	BINARY	B_1	B_2	B_3	B_4	TOTAL
0	0	0	0	0	0	0
1	*I*	0	0	0	0	*I*
2	0	2*I*	0	0	0	2*I*
3	*I*	2*I*	0	0	0	3*I*
4	0	2*I*	2*I*	0	0	4*I*
5	*I*	2*I*	2*I*	0	0	5*I*
6	0	2*I*	2*I*	2*I*	0	6*I*
7	*I*	2*I*	2*I*	2*I*	0	7*I*
8	0	2*I*	2*I*	2*I*	2*I*	8*I*
9	*I*	2*I*	2*I*	2*I*	2*I*	9*I*
10	0	0	0	0	0	0

a minimum resolution time of better than 10μsec has been achieved, and it is expected that with h.f. transistors (α cut-off 5Mc/s) a minimum resolution time of 1μsec may be achieved.

The sequence of the count is shown in Table 4. The binary provides an output pulse for every other input pulse. This output pulse is amplified and used to drive the ring. This drive pulse is fed to all four bistable circuits via gates which will only allow the pulse to pass and switch the circuit on, if the preceding circuit is already on. The fifth circuit is a blocking oscillator, which is triggered by the fifth pulse to the ring, corresponding to the tenth input pulse, and provides both a switch-off pulse resetting the ring to zero and an output pulse.

There were two main problems with this counter, the first was to design a binary which would operate reliably at 100kc/s, and the second was to develop a suitable gating system for the ring.

In the binary circuit it was found necessary to use catching diodes to prevent the transistors bottoming; this improved both the high frequency performance and the

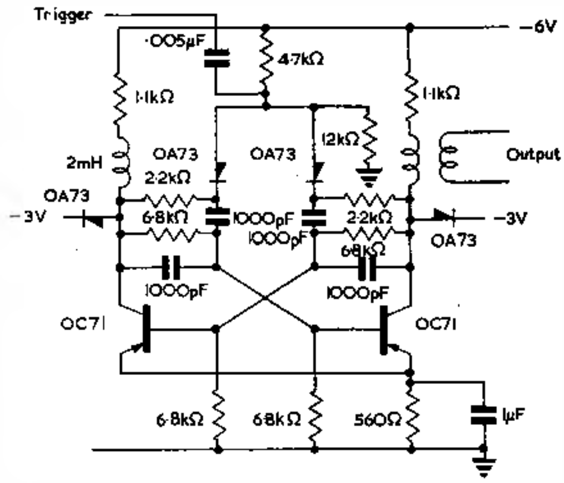


Fig. 26. Complete circuit of the binary counter

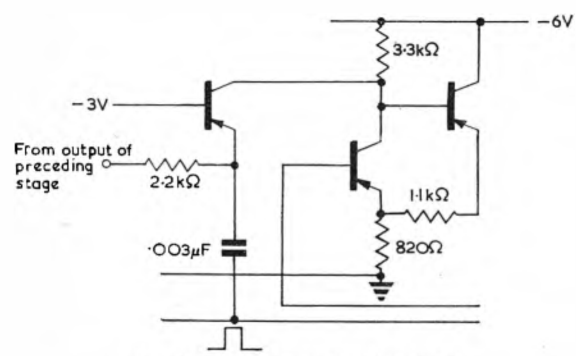


Fig. 27. Ring gate and bistable circuit

trigger sensitivity of the circuit. Also a somewhat complicated binary gating system controlled by the collector potentials was required to ensure that the circuit operated reliably. The system used only allows a pulse to reach the base of the transistor that has to be switched off. Other simpler systems tend to switch both transistors off and rely on the charge on the cross-coupling capacitors to ensure that the circuit returns to the correct state.

The complete circuit of the binary is given in Fig. 26. Up to 10kc/s the trigger pulse required is approximately 1V, but this increases to about 3V at 100kc/s. Therefore in the complete counter a drive circuit is used to the first decade with slightly modified input to the binary, but each succeeding decade is coupled direct from the output of the preceding decade.

In the ring it was decided to use the asymmetrical bistable circuit since this had a low impedance output to operate the gate. The requirements for the gate are that

* Mullard Research Laboratories

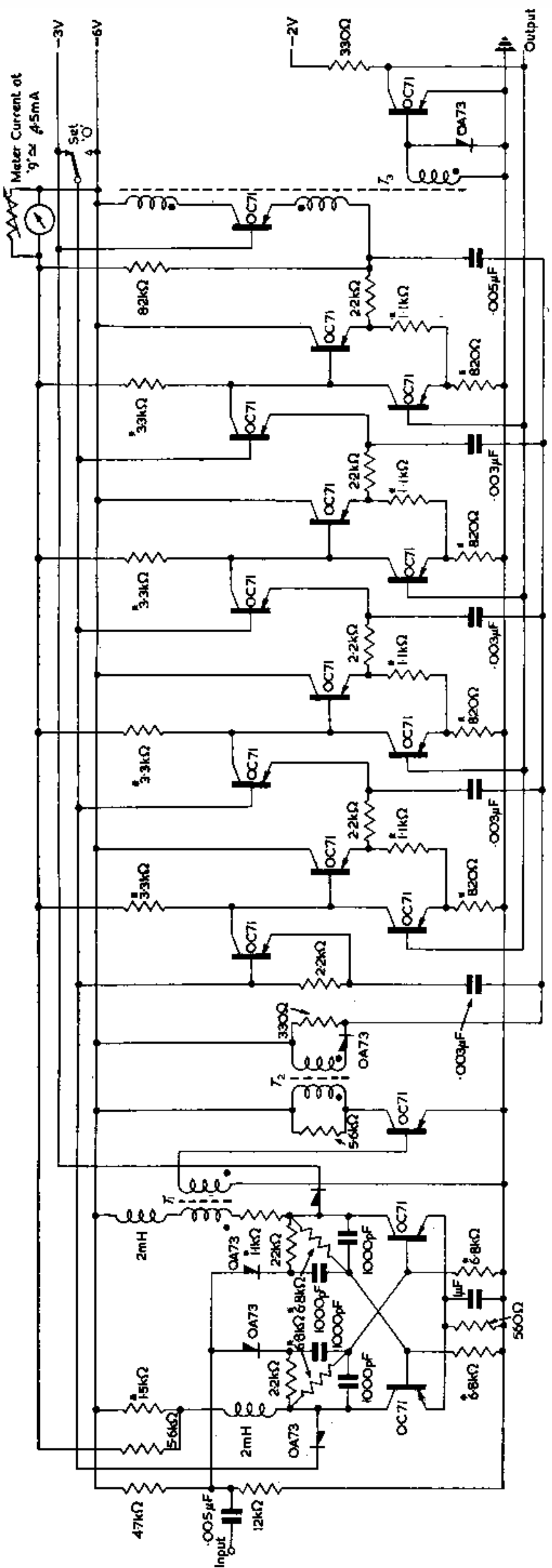


Fig. 28. Complete circuit of a decade counter

TRANSFORMER DETAILS: *FX1011 Pot Cores.* T_1: *Pri.* 40 *turns* 38 *s.w.g. En.; Sec.* 30 *turns* 38 *s.w.g. En.* T_2: *Pri.* 80 *turns* 38 *s.w.g. En.; Sec.* 30 *turns* 38 *s.w.g. En.*
T_3: *Pri.* 200 *turns* 42 *s.w.g. En.; Sec.* 40 *turns* 42 *s.w.g. En.; Tert.* 40 *turns* 42 *s.w.g. En. Resistor values marked * are* ±1 *per cent tolerance.*

it must only pass the trigger pulse when the preceding circuit is on, and also that there is a delay of approximately 10μsec before it operates, otherwise when the first circuit is switched on the gate will open and allow the second circuit to be switched on by the same pulse. A simple transistor amplifying gate was finally decided on (Fig. 27). When the preceding circuit is off its output is at −5·5V and this cuts-off the gating transistor. When the preceding circuit is on, its output is at −3V and the emitter of the gating transistor is then just in conduction and a trigger pulse will be passed. The 1·8kΩ resistor and ·003μF capacitor provide the necessary delay. Also the collector diode of the gating transistor acts as a catching diode for the bistable circuit preventing the first transistor from bottoming. The complete circuit of one decade is shown in Fig. 28. The intermediate supply lines of −2V and −3V must be stabilized in such a manner that they vary in direct proportion to the input supply. This can easily be achieved with simple transistor circuits.

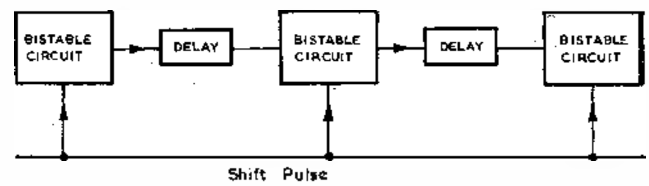


Fig. 29. Block diagram of shifting register

The complete counter has been tested and found to operate reliably over the temperature range 0°-40°C with a supply voltage between 9V and 5·5V.

A Shifting Register

Transistor versions of the valve shifting register, using Eccles-Jordan bistable circuits, have been made using low frequency transistors (α cut-off ≃ 500kc/s). But the operating speed of these circuits is limited by the gating systems employed. The switching time of the bistable circuit is about 5μsec and therefore if an ideal gating system could be devised the register should operate at 200kc/s, but in practice speeds between 50 and 100kc/s are the best that can be obtained. In an attempt to reach higher operating speeds a circuit using delay line coupling between stages was developed. In this register, (Fig. 29) the shift pulse switches all the stages to zero and any stage that was on emits a pulse into the delay line. This pulse, after the delay period, switches on the following stage, thus transferring the information one place to the right. In this register a bistable circuit may be required to switch off and then on again in the pulse repetition period. The switching time of the bistable circuit is approximately 5μsec, therefore the shortest pulse repetition period is 10μsec. In practice it is found that this circuit will operate at this repetition period, which corresponds to 100kc/s. The asymmetrical bistable circuit was used since it had a low impedance output to feed the delay line. The complete circuit of one stage is shown in Fig. 30.

The grounded collector stage terminating the delay line acts as a buffer amplifier and also allows correct termination of the delay line, thus preventing spurious pulses appearing on the delay line which cause unreliable operation. Each stage requires a shift pulse of 4μsec duration and amplitude 2V 2mA.

It is hoped that using h.f. transistors (α cut-off 5Mc/s) the maximum clock pulse rate may be increased to 1Mc/s.

Transistors with Ferrite Stores

The drive requirement to switch a 2mm square loop ferrite toroid is a magnetic force equivalent to 700mAT.

For ease of construction of matrix stores, only one turn per core is used so that the pulse generators are required to provide a pulse of 700mA for full switching or a pulse of 350mA on each of two lines for single core selection.

Two methods of operation will be considered:

(1) Transistors as gating elements for current pulses.

(2) Transistors as pulse generators.

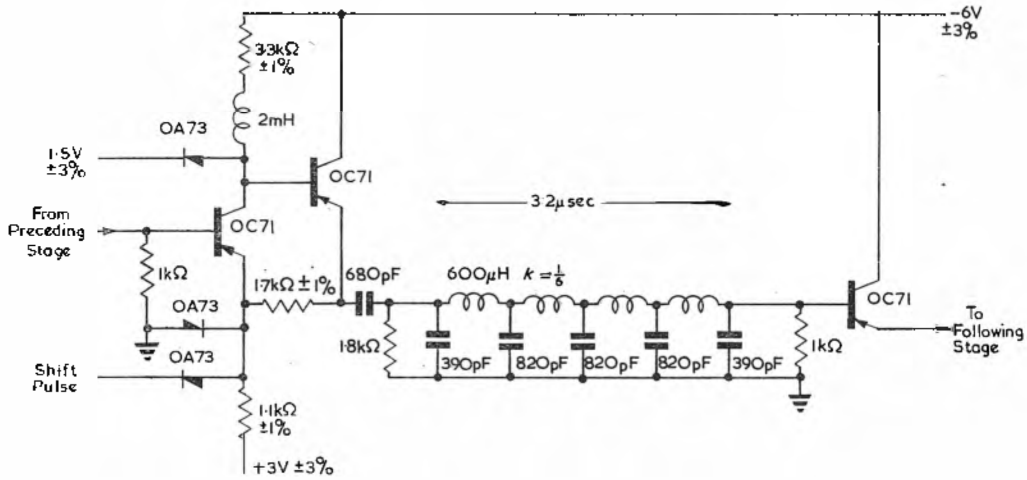


Fig. 30. Complete circuit of shifting register stage

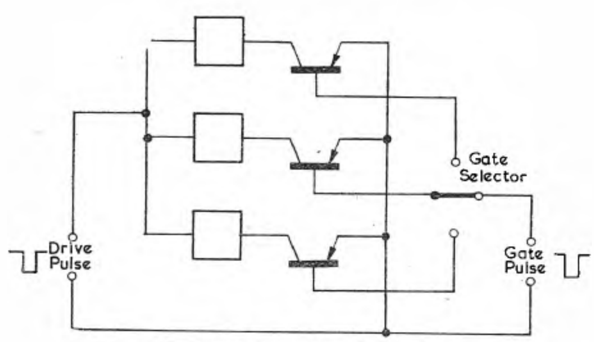


Fig. 31. Drive pulse gating system

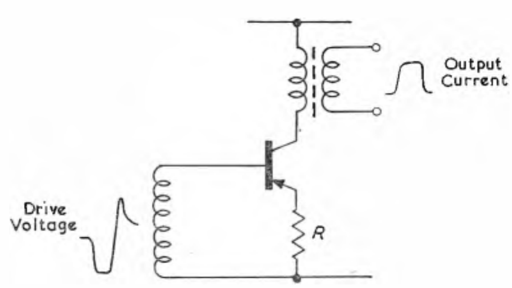


Fig. 32. Amplifier for high-current pulses

Gating Current Pulses

As a gating element the transistor may be used to select which of a number of parallel loads is operative for a current source. Thus in Fig. 31 current from the generator will pass only through the load which has its transistor gate open. It is possible with this system for the transistor to pass a current pulse of any shape without deterioration of the rise time.

The limitation on the current passed by the gate is one of heat dissipation. If the transistor is held bottomed ($V_c < 0{\cdot}1$V) during the drive pulse period, the collector current may have a peak value of several hundred milliamperes, but the peak dissipation will not be more than a few tens of milliwatts. The gate pulse current required is at least $1/\alpha'$ of the peak collector current, and since in most types of transistor α' is low at high values of collector current, the overall gain in the transistor gate may be very low.

Another effect which must be taken into account is the hole storage in the transistor. This has the effect in a normal amplifier that if a pulse is applied to the base, the collector current does not reach its final value until a short time has elapsed. In the gate circuit the effect is that the transistor will not pass a collector current pulse and remain in the bottomed condition until some time after the pulse has been applied to the base. This delay time may be reduced by increasing the amplitude of the gate pulse, but this reduces the gain of the gating transistor. A further effect of hole storage is that following the current pulse a further period must elapse before the transistor will no longer pass a current pulse and the gate can be considered closed.

Since the delays involved in operating the gate may be comparatively large, this system is not amenable to operation at high repetition frequencies, using low frequency transistors. For collector pulse currents of the order of 120mA with the OC71 transistors the delays involved are 5μsec between the application of the base gate pulse and the collector current pulse, and 10μsec must then elapse before the drive pulse is again applied to the system. With high frequency transistors these delays would be reduced considerably, but since these transistors are capable of providing sufficiently fast current pulses on their own, no advantage is to be gained by using them in this system.

Pulse Generators

The transistor may be used as an amplifier driven from a blocking oscillator (Fig. 32) to provide a high current pulse. Since the blocking oscillator provides a constant voltage pulse, a resistor is included in the emitter of the amplifier to stabilize the output current. The rate of rise of this current is determined mainly by the frequency response of the transistor and the value of r'_{bb}.

The circuit of Fig. 32 with $R = 15\Omega$ gave a collector current pulse of 100mA for an input voltage pulse of 2V from a blocking oscillator. The initial surge of base current was 30mA falling to 5mA during the pulse. The rise times were:

	OC72	*OC45*
Base voltage	1μsec	0·2μsec
Collector current	1·5μsec	0·2μsec

Read Pulse Amplifiers

The pulse output obtained when a 2mm ferrite core is switched is triangular in shape, the total duration being of the order of 2μsec. The amplitude of the output pulse

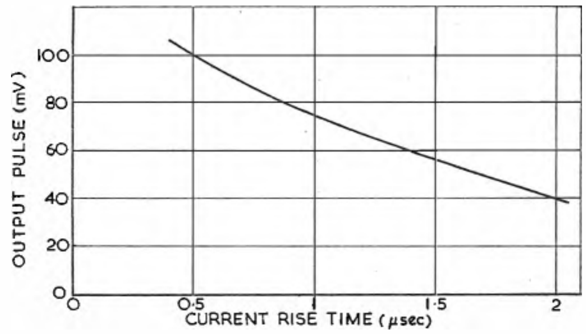


Fig. 33. The variation of output pulse amplitude with drive pulse rise time for a particular 2mm ferrite core

depends on the rise-time of the current in the driving pulse (Fig. 33), but is generally of the order of 50 to 100mV. The source impedance is very low.

With high frequency transistors a normal grounded emitter stage may be used to amplify this pulse, a single stage giving an output of more than 1V in 1kΩ.

With low frequency transistors, the gain of a single stage is comparatively low due to the short length of the pulse involved. However, if the input pulse is stepped up by a factor of 5 in a transformer, a grounded emitter transistor will amplify this pulse to give an output of about 2V in 1kΩ. The rise-time of the output pulse is approximately 1μsec and the fall time 1·5μsec.

Conclusions

This paper has described the operation of junction transistors under pulse conditions, and has shown some of the types of circuit that may be used in digital computers. Those familiar with computing techniques will no doubt see many other possibilities in using transistors to perform various functions, and in fact many circuits have been published since this paper was written.

Although some of the circuits described bear some resemblance to known valve circuits, it is felt that better results can be obtained by designing transistor circuits to suit the particular characteristics of the transistors, rather than to consider the transistor as another type of valve and modify known circuits accordingly.

Acknowledgments

The authors wish to thank the manager of Mullard Research Laboratories and the directors of Mullard Ltd for permission to publish this paper.

Extension of Telephone Weather Service

The G.P.O.'s telephone weather forecasting service has proved very popular in London and it is now being extended to Birmingham, Manchester, Liverpool, Edinburgh, Glasgow, Cardiff and Belfast.

For the extension of this service to these seven provincial centres, new recording and replay equipment has been adopted by the Post Office. Basically, this is the Emidicta office dictating machine, chosen for its reliability, its ease of handling by non-technical staff, and its adaptability to the specialized P.O. technical and operational requirements.

The Emidicta is a magnetic disk recording and relay instrument using an 11in plastic disk coated with high coercivity magnetic material, over the centre of which is placed a spiral tracking disk. Attached to the underside of the arm carrying the record/replay head is a toothed wheel which engages with the grooves on the tracking disk and so tracks the head across the magnetic disk. It also steps the head back as required to repeat part or all of the recording when it is caused to rotate by means of pulses supplied to it from a repeater mechanism. Erasure of the recording is made by lowering a special multi-pole permanent magnet into close proximity with the disk.

The special Emidictas supplied to the Post Office incorporate certain modifications which have been carried out by E.M.I. in co-operation with the Post Office Engineering Staff to enable it to produce automatically a continuous repetition of the recorded message.

An adjustable rod, sliding in a tube and capable of being locked in any desired position by the operator, is mounted on the arm carrying the head. Micro-switches are fitted at the outer and inner limits of the travel of the arm. The outer micro-switch is actuated when the head is in the correct position at the start of the disk and it initiates the recording or the replay; the inner micro-switch is actuated by the adjustable rod when the head arm reaches the end of the recording; this operates the relay providing the step-back pulses which return the head to its outer position. The pulses are produced by a cam driven from the motor.

Operation is simple. The machine is first switched to 'Record' and the head arm is positioned so that the outer switch is just closed. At the end of the recording the machine is stopped and switched to 'Replay'; the movable rod on the head arm is then adjusted so that the inner micro-switch just closes and initiates the step-back pulses. When the head arm reaches the 'start' position again, the outer micro-switch closes, the pulses are cut off and the recording is reproduced in the normal way, after which the cycle of operations is repeated automatically and continuously.

During the stepping-back process the amplifier is muted, the muting being removed by the operation of the outer micro-switch. The step-back pulses are designed to return the head to the start of the disk in not more than ten seconds even with the maximum recording time of six minutes. The outer micro-switch is provided with a fine adjustment so that the starting position of the head can be precisely located at the beginning of the recorded message.

Since the recordings will be made by many different operators, a system of delayed a.g.c. has been incorporated to ensure that the output level will remain reasonably constant. Thus changes in input levels of up to 16dB produce no more than a 3dB change in output.

The head has been designed to use replaceable pole tips, which are easy to renew and have a life of many hundred hours. To increase by as much as four times the normal life of the pole tips a highly polished disk has been developed, while to minimize the destructive effect on the pole tips of dust and grit particles which, if allowed to settle on the recording disk, become embedded in its surface, a special plastic cover has been made to fit completely over the turntable deck.

The effective front gap of the record/replay head is 0·0005in, which provides an adequate high-frequency response with an average track velocity of 6in/sec. The width of the recording track is 0·01in and the spacing between adjacent tracks is 0·025in.

A Transistor Demodulator

By H. Sutcliffe*, M.A., A.M.I.E.E.

This article describes a demodulator or phase-sensitive rectifier which employs junction transistors. An economical design can be achieved since the transistors function both as linear signal amplifiers and as non-linear rectifying elements.

A RECENT article by Moody[1] stimulated interest in the possibility of a small, compact, rugged d.c. amplifier using transistors. The d.c. stability of transistors with temperature is poor and it is necessary to use a modulator with stable non-linear circuit elements followed by an a.c. amplifier and a demodulator. Amplification of an alternating signal by transistors can be performed by a variety of well-established techniques[2], but the design of a demodulator for use with transistor circuits has not received such extensive study. A transistor chopper as described by Bright and Kruper[3] may be used and can be extremely effective in such an application, since the transistors are used as on-off switches and can demodulate at power levels considerably greater than their own power dissipation. They do not, however, make any contribution to signal amplification.

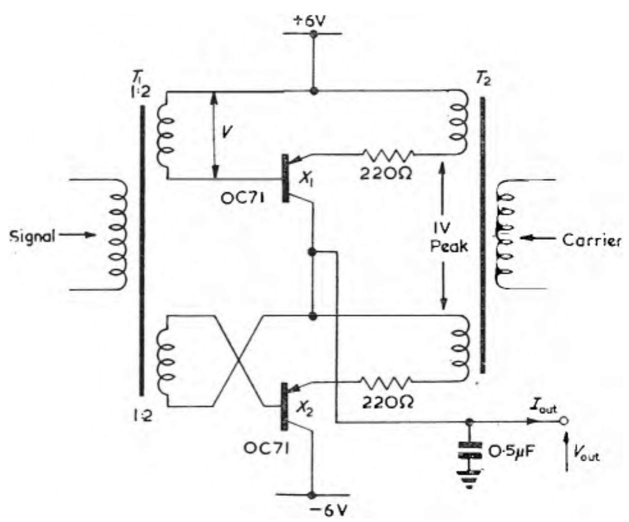


Fig. 1. Circuit diagram of a half-wave demodulator for a 2kc/s carrier

The demodulator to be described in this present article is for signals carrying information rather than power. The d.c. output requirement was typically 0·5mA into 10kΩ. This permits an arrangement in which the transistors have considerable amplification in addition to their behaviour as phase-sensitive rectifying elements and a very economical circuit can be designed.

Circuit Description of Demodulator

A half-wave circuit is shown in Fig. 1. In general terms the circuit operation is as follows. The carrier reference voltage from transformer T_2 turns the emitter currents on and off in alternate half periods of the carrier waveform. If the input signal from T_1 were zero, the emitter to base voltages applied to X_1 and X_2 would be equal and, apart from a possible difference in the characteristics of the two transistors, there would be no output current. Now suppose a signal to be fed from transformer T_1 in such a sense that, during the half period in which the transistors are conducting, a negative voltage appears on the base of X_1 and a positive voltage on the base of X_2. The emitter to base voltage of X_1 is thus greater than that of X_2, causing the former to conduct more strongly, and a positive value of the output current I_{out} appears.

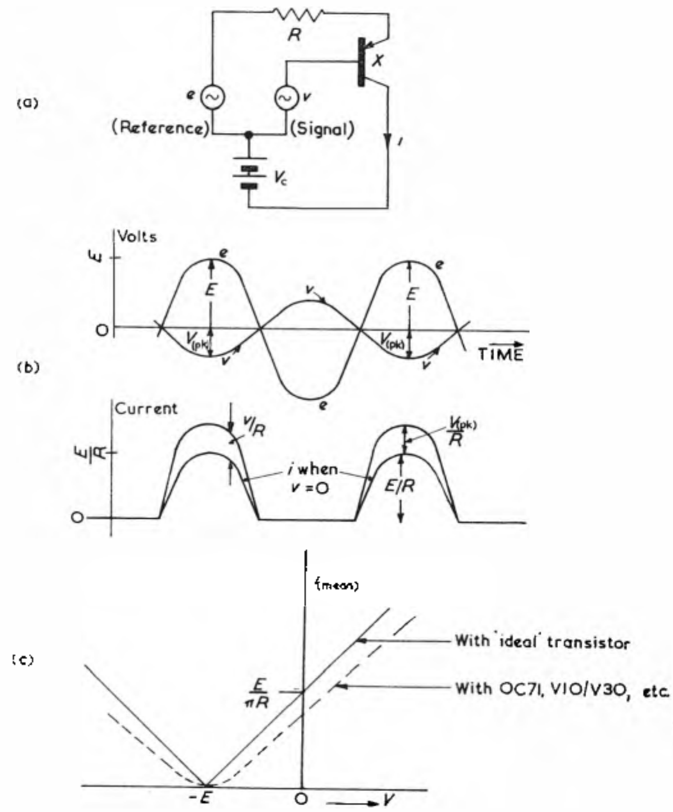


Fig. 2. (a) The equivalent circuit, (b) waveforms, (c) output characteristic, of one transistor demodulation element

A more quantitative assessment of the performance of the demodulator can be obtained by considering Fig. 2, which is concerned with one only of the demodulator transistors and its associated circuit. It is convenient to consider first the case of an ideal transistor, that is a transistor whose base current is zero and whose base to emitter p.d. is zero when conducting in the forward direction. These idealizations are realized approximately when circuit component values and signal values are as shown in the figure. In this ideal case then, the value of collector current i would be given by:

$$e > v,\ i = (e - v)/R;$$
$$e < v,\ i = 0;$$

where symbols have meanings as indicated in Fig. 2.

* *University of Bristol, formerly Smith's Aircraft Instruments Ltd.*

Current waveforms would be as illustrated in Fig. 2(b), and the resulting mean value of i as a function of signal and reference voltage amplitudes is as shown in Fig. 2(c). The actual output current is less than that in the ideal case for two reasons. The emitter current is slightly less than $(e - v)/R$, because of the finite value of the emitter to base p.d. of the transistor, and the collector current is slightly less than the emitter current because a small portion of the emitter current flows in the base circuit. The result of these two

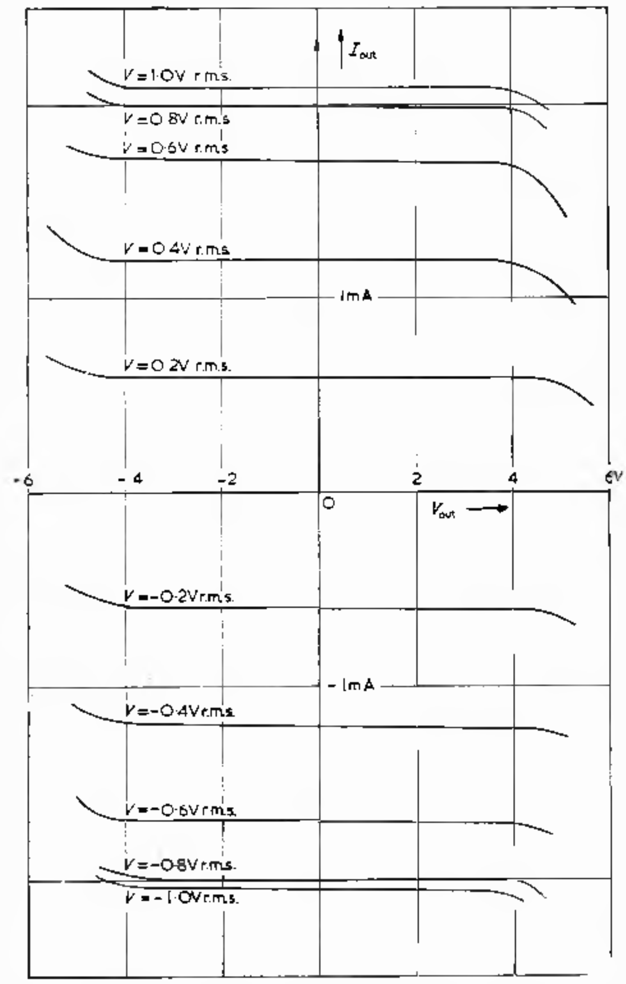


Fig. 3. The measured output characteristic of the circuit shown in Fig. 1

effects is to modify the ideal output characteristic to the dashed line in Fig. 2(c). The circuit arrangement is such that, with normal junction transistors, the value of V_c has a negligible effect on the current output characteristics, provided that V_c is adequate to maintain a collector voltage greater than, say, 0·5V.

The reversal of slope of the characteristic as $V_{(pk)}$ passes through the value $-E$ is because the transistor begins to conduct on the wrong half cycle of the carrier waveform. This effect limits the linear range of the demodulator when supplying current into a low impedance load. Limitation of the linear regime when the load has a high impedance will be due to collapse of transistor action because of inadequate collector voltage.

The half-wave demodulator of Fig. 1 employs a pair of the elementary circuits of Fig. 2(a). A full wave demodulator would employ four such elements in conjunction with extra secondary windings on transformers T_1 and T_2.

Performance of the Demodulator

Fig. 3 shows the measured output characteristic. The curves show the high degree of linearity within the region ± 1mA and ± 4V.

It is of interest to compare the measured transconductance of the demodulator with the theoretical value assuming ideal transistors.

From the curves of Fig. 3, the transconductance g referred to the secondary windings of T_1 is:

$$g = (\Delta I_{out}/\Delta V_{(pk)}) = (1{\cdot}2/0{\cdot}4\sqrt{2}) = 2{\cdot}12\text{mA/V}$$

The theoretical value for the transconductance is:

$$g = 2/\pi R = 2{\cdot}27\text{mA}/V, \ (R = 280\Omega)$$

where R is the total resistance (including transformer winding resistance) in each emitter circuit.

The demodulator presents a time-varying impedance to the input signal. The effective value, measured at the primary winding of T_1, is in the region of 1·1kΩ and it is desirable that the impedance of the source feeding the demodulator is appreciably smaller than this. A transistor used in the common collector arrangement is suitable.

The drain on the battery power supply is approximately 1mA. It is proportional to the carrier voltage and inversely proportional to the resistance R. The load on the carrier source is approximately $R/2$ for half cycles only of the waveform.

The presence of the pair of emitter series resistors has a pronounced stabilizing effect on the performance of the demodulator. The output characteristic is to a large extent independent of changes in battery voltage, carrier voltage and transistor characteristics.

Acknowledgments

This article is based on a programme in the Research Laboratories of Smiths Aircraft Instruments Ltd., and is published by permission of the management. The constructional and experimental work was performed by G. M. Willdey, B.Sc.

REFERENCES

1. Moody, N. F. A Silicon Junction Diode Modulator. *Electronic Engng. 28*, 94 (1956).
2. Shea, R. F. Transistor Audio Amplifiers. (John Wiley, New York).
3. Bright, R. L., Kruper, A. P. Transistor Choppers for Stable D.C. Amplifier. *Electronics 28*, 135 (April 1955).

Indication Equipment for Remote Control Systems

A d.c. multiplexing equipment has been developed by The General Electric Co. Ltd., for the continuous indication, of the operating state of apparatus situated in remote locations. The information from up to twelve such items of equipment can be relayed to the control along a single pair of wires up to five miles in length.

The equipment, which can be applied to installations such as hydro-electric schemes or mines, uses transistors and consists of up to twelve transmitter units and a cabinet containing the corresponding receiver sub-units.

Each of the transmitter units is an audio-frequency oscillator tuned to a different frequency and distributed along the wires so that one is adjacent to each point to be monitored.

The receiver sub-units comprise a band-pass filter, amplifier, detector and relay, each band-pass filter being tuned to a particular transmitter frequency.

Any operation of the equipment being monitored can be arranged to switch the adjacent monitoring transmitter. At the presence of a particular oscillator tone on the wires, the relay in the corresponding receiver operates and can be used to show which transmitter is working.

Each transmitter unit is enclosed in a hermetically-sealed can and a case can be provided at the side to house a supply battery, which has a life on continuous load of approximately six months before replacement becomes necessary.

The receiver cabinet will house twelve working receiver sub-units and twelve spare sub-units.

Short News Items

An exhibition will be held in the College of Science and Technology, Manchester, on 26 and 27 March under the auspices of the Radio and Telecommunications Group of the North Western Centre of the Institution of Electrical Engineers. The exhibition will be open from 10 a.m. to 9 p.m. on the above dates.

The Television Society's Annual Exhibition will be held at The Royal Hotel, Woburn Place, London, W.C.1, from 5-7 March inclusive. The times of opening will be 11.30 a.m.-8 p.m. on Tuesday, 5 March, and 12 noon-8 p.m. on Wednesday and Thursday, 6 and 7 March. Tickets are available free from The Television Society, 164 Shaftesbury Avenue, London, W.C.2.

The Radio Industry Council has announced the award of five premiums of 25 guineas each for articles published in the public technical press during 1956. One premium has been awarded for each of the following.

"Weather Avoidance with Airborne Radar" by P. L. Stride (Ekco Electronics Ltd), *British Communications and Electronics,* April, 1956.

"Particle Accelerators and their Applications" by D. R. Chick (Research Laboratories, A.E.I.) and C. W. Miller (Research Department, Metropolitan-Vickers Electrical Co Ltd), *British Communications and Electronics,* October / November, 1956.

"Klystron Control System" by R. J. D. Reeves (E. K. Cole Ltd), *Wireless Engineer,* June, July and August, 1956.

"Two-Channel Stereophonic Sound Systems" by F. H. Brittain and D. M. Leakey (G.E.C. Research Laboratories), *Wireless World,* May and July, 1956.

"TRIDAC—A Research Flight Simulator" by J. J. Gait and J. C. Nutter (Elliott Bros Ltd), *Electronic Engineering,* September/October, 1956.

Forty articles were submitted in all. The awards will be presented at a luncheon to be held by the Public Relations Committee of the Radio Industry Council on Thursday, 14 March.

Radio and Electronic Component Exhibition. The R.E.C.M.F. has discontinued the lapel badge system of admission for visitors to the 1957 exhibition to be held from 8-11 April. This year double tear-off tickets will be issued which will admit to both sections of the exhibition, at Grosvenor House and Park Lane House. These tickets will be issued from the R.E.C.M.F. offices and all prospective bona fide visitors should apply in writing direct to the Secretary, Radio and Electronic Component Manufacturers' Federation, 21 Tothill Street, Westminster, London, S.W.1.

Airmec Ltd are holding an exhibition of their electronic instruments at the Napier Hall, Vincent Street, London, S.W.1, from 25-28 March. The complete range of Airmec electronic equipment, including several entirely new instruments, will be shown. Tickets for free admission will be forwarded on request to Airmec Ltd, High Wycombe, Bucks.

A Conference on Physics of the Solid State, organized by Professor L. B. Bates, will be held on 8, 9 and 10 April in the Physics Department, the University of Nottingham. Requests for application forms, programmes and all enquiries should be addressed to the Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7, marked for the attention of Miss Miles.

A Decca Navigator mobile chain will be used in connexion with the nuclear tests which were announced by the Prime Minister in the House of Commons and which will be held on Christmas Island this year. Combined with an instrument known as Flight Log this navigation aid will show pilots of participating RAF planes their positions at a glance.

Electronics in Automation is the theme of the British Institution of Radio Engineers' Convention to be held in the University of Cambridge from 27 June-1 July. Arrangements for the Convention have been completed and some thirty papers are being presented to cover six sessions. Office Machinery and Information Processing; Machine Tool Control; Chemical and other Processes; Simulators; Automation in the Electronics Industry; Automatic Measurement and Inspection. Papers will be read and discussed in the Clerk Maxwell and Green lecture theatres of the Cavendish Laboratory. Institution members and other delegates who are attending the whole of the Convention will be accommodated in King's College. Further information may be obtained from the Institution's offices, 9 Bedford Square, London, W.C.1.

The Council of the Radio Communication and Electronic Engineering Association has approved the formation of a section to deal with electronic data processing, to be known as the Data Processing Section. The new section is intended to serve the interests of the many firms who are concerned in the new field of electronic computer equipment, both digital and analogue, and also the manufacturers of input and output devices and allied equipment.

The Annual Report of the Radio Communication and Electronic Engineering Association for 1956 was presented at the annual general meeting on 1 February. It is stated in the report that the foremost work of the Marine Committee during the year has been the revision of the Marine Radar Performance Specification in conjunction with the Ministry of Transport and Civil Aviation and the Radio Advisory Service. Liaison with bodies outside the Association has continued to be an important part of its work; contact was also maintained with the Ministry of Supply on various matters. In the Chairman's Report last year reference was made to the proposals of the Frequency Advisory Committee for improved Government machinery to deal with frequency planning. During the past year this matter has been pursued; while there has not yet been any official reaction from the Postmaster-General, discussions clarifying the industry position have been held with G.P.O. officials. A booklet for general circulation setting out the activities of the Association is in an advanced state of production. Two other publications in course of preparation by the Association are "A.C. Synchro Systems for Civil Aircraft" and "Guide to the Specification and use of Quartz Oscillator Crystals". The Publicity Committee is planning a display for the Association which will be part of the exhibit of the British Electrical and Allied Manufacturers Association in the Brussels International Exhibition in 1958.

Sir Hugh Beaver has resigned from the Council for Scientific and Industrial Research. He has now become Deputy President of the Federation of British Industries and has been obliged to limit his other activities.

Exports of British radio equipment of all kinds in 1956 totalled more than £40 m, a new record. This was recently announced by the Radio Industry Council.

The annual National Exhibition of the Manufacturers of Radio and Television Components, Accessories, Electronic Tubes and Electronic Measuring Instruments will be held from 29 March-2 April at the Parc des Expositions, Porte de Versailles. Further information may be obtained from the Salon National de la Pièce Detachée Radio, 23 rue de Lubeck, Paris XVI.

Mr. J. F. Richardson has been appointed Technical Secretary of the Radio Communication and Electronic Engineering Association. He will work under Mr. H. E. F. Taylor whose appointment as Executive Secretary of the Association was announced recently.

LETTERS TO THE EDITOR

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

Relay Adding Circuits

DEAR SIR,—In binary adding relay circuits described in previous correspondence the addition of two binary digits has required the initial setting of the digits in relays, the 'add' circuit being formed by contacts of these relays in a multicontact chain.

The 'add' circuit described here may interest other readers. Its purpose is to accept pulses, coincident in time, direct from the binary digit sources and to provide a temporary 'sum' store for the duration of the original pulse time. The relays used are the single contact, high speed type with duo-coils.

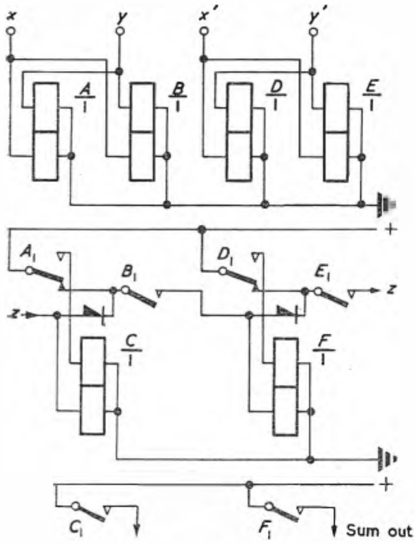


Fig. 1. Binary adding circuit

In Fig. 1 a binary stage has three relays, A, B and C. The two coils each of A and C are wired in opposition, giving mutual cancellation with both excited. The two coils of B provide alternative pick-ups. Relay B will be called when either or both x and y are pulsed, relay A will operate with one pulse only at x or y.

With both pulses accepted by the stage, B only is called and its contact is transferred to provide a carry-out pulse to the next stage, relay C (sum) is not called. A carry pulse z, into the stage at this time, will call relay C to register sum 1.

With one pulse at either x or y both relays A and B are called, A, contact transferring to call relay C and prevent a carry out pulse via relay B. A carry pulse into the stage will excite the second coil of relay C, so to suppress the sum, and will pass, via relay B, as a carry-out.

A carry pulse into an empty stage calls relay C to give a sum 1.

In Fig. 1 the rectifiers in the carry line limit the feeding of a long carry pulse to four higher stages. An alternative sum/carry circuit, Fig. 2, would have no such limit but requires an additional A contact provided by an extra relay, A', whose coils are paralleled with those of relay A.

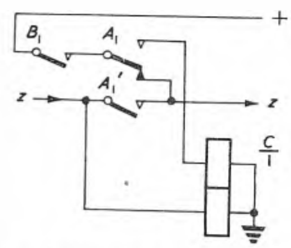


Fig. 2. Alternative sum/carry circuit

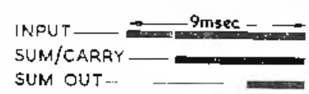


Fig. Typical time sequence

Using relays of 1 to 2msec operating time a suitably controlled time sequence governing the sum-out pulse could be in order of that shown in Fig. 3.

Yours faithfully,

A. J. KEEN,
The British Tabulating Machine Co. Ltd.,
Letchworth.

Transistor Stabilizing Circuits

DEAR SIR,—It may be of interest to quote some practical figures obtained from simple transistor stabilizing circuits.

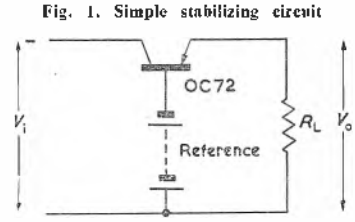
For the simple circuit (Fig. 1) the output impedance $R_o = r_e + R_b/\beta$, where R_b includes the reference impedance, and the stabilization ratio $F =$

$$\frac{R_b R_L}{r_c (R_L + r_e + R_b/\beta)}$$

Using an OC72 to provide a stabilized 6V supply at 60mA, we have:

$$R_L = 100\Omega$$
$$r_e = 2{\cdot}5\Omega$$
$$r_b = 150\Omega$$
$$r_c = 500\text{k}\Omega$$
$$\beta = 50$$

Fig. 1. Simple stabilizing circuit



These give $R_o = 5$

and $F = 1/3\,300$

It can be seen that the stabilization factor is quite good but the output impedance is rather high. However, by introducing an amplifier to feed back an antiphase error signal from the output to the base, R_o is reduced by the gain of the amplifier and F is improved by roughly the same factor. To obtain a reasonable gain from the amplifier, and to keep the R_b low, it is necessary to interpose an impedance transformer, such as an emitter-follower, as in Fig. 2. With this circuit an output impedance of less than 0·3 was obtained, and a stabilization factor of better than 20 000.

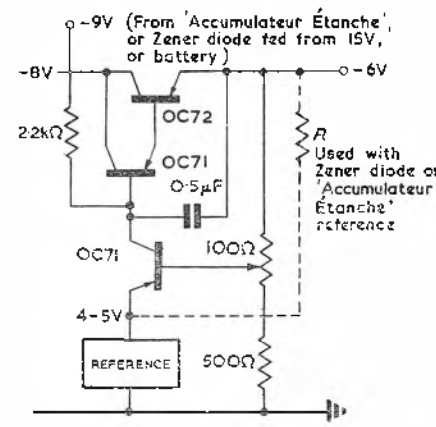


Fig. 2. Practical circuit

The reference used could be either battery, zener diode or an 'Accumulateur Etanche'. This last device is a form of sealed accumulator (as, of course, the name implies), and has been available for some months in this country.

A comparison of the three references may be of value.

For a $4\frac{1}{2}$V reference, the slope impedances are all about the same value (10-15Ω) and the temperature coefficient also is very similar for each type, i.e. $-0{\cdot}04$ to $+0{\cdot}01$/°C for zener diodes[1] and $+0.02$/°C for Accumulateur Etanches and batteries.

If zener diodes are not available, then the Etanche is preferred to the battery because of a longer life expectancy.

Yours faithfully,

C. S. EVANS,
J. L. CARROLL,
Ferranti Ltd.,
Wythenshawe,
Manchester 22.

REFERENCES

1. SMITH, D. H. The Suitability of the Silicon Alloy Junction Diode as a Reference Standard in Regulated Metallic Rectifier Circuits. *Trans. Amer. Inst. Elect. Engrs.* 73, Pt. 1, 645 (1954).

BOOK REVIEWS

Handbook of Basic Circuits

By Mathew Mandl. 365 pp. 80 figs. Demy 8vo. The Macmillan Co. Ltd. 1956. Price 52s. 6d.

THIS is a difficult book to review as it is not at all clear for whom it was written. It is said to contain 136 circuits used in television, f.m. and a.m. It deals with a wide variety of things such as klystrons, transistors, component measurement, filters, colour transmitters and receivers, and it is stated that the circuits and descriptions are applicable to "industrial and commercial electronics and to high fidelity and public address systems." In fact, it turns out to be a collection of circuits and brief descriptions of them with no sequence except that they are arranged alphabetically.

Many basic circuits commonly met with in the field of television and radio are covered by this book but the student may find the descriptions very confusing. For instance, on page T-17-2 there appears the explanation of the working of a t.a.t.g. oscillator. Following a rambling account of the behaviour of a parallel resonant circuit when different frequencies are applied, the author states that the circuit will not oscillate when both grid and anode circuits are tuned to the same frequency because "only the resistive component will be present in the inductance. Since the lumped components now represent only resistance and capacitance, oscillations (and flywheel effect) will not occur." Later he goes on to say that if the anode circuit is tuned to a lower frequency "the oscillator can generate a frequency, because the lumped constants comprise inductance and capacitance as well as resistance." Still later he continues "oscillations are procured for any setting of variable capacitor but the desired frequency is produced only under the conditions outlined above."

On page D-6 it is stated that a d.c. setter improves the low frequency response of a video amplifier.

On page V-8 there appears the explanation of the working of a resistance coupled voltage amplifier for audio use. This is very confusing. Typical examples are: "During the interval when high potential positive alternations of the signal appear at the grid, peaks of signal may drive the grid positive and cause grid current to flow. This will result in electrons gathering on the capacitor (negative with respect to grid) and driving the grid highly negative." The author goes on to say that this stops the valve amplifying and explains that the grid capacitor must not be too big or its leakage resistance will make the grid positive and the larger the capacitor the smaller must be the grid leak. Later on he says that in normal resistance coupled amplifiers the grid should not be driven into the positive region by the grid signal.

On page S-7 when describing the action of a sync separator the following sentence appears: "Because the sync tip places a positive charge on the grid with respect to a negative charge on the cathode. . . .". Such explanations are unlikely to be of much help to students.

If the book was intended as a reference book for engineers wanting to copy a circuit for a certain purpose without having to work one out, the book may be useful, but in the circuits errors appear and component values are hardly ever given. On page S-5 for instance, there is no h.t. supply for the oscillator valve.

In my opinion the book is a hodgepodge of circuits, the explanations are not clear and it is difficult to see the purpose of the book.

C. H. BANTHORPE.

Fixed Capacitors (Volume 3. Radio and Electronic Components)

By G. W. A. Dummer. 260 pp. 40 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1956. Price 45s.

THIS 260-page book is undoubtedly a worthy successor to the two volumes in the series on components already published by the author. Here, as before, he has taken Service type components, in this case fixed capacitors as representative of high standards of manufacture built to withstand more stringent climatic conditions than other basically similar commercial types.

The book makes extremely interesting reading and is a virtual mine of information on practically every aspect of fixed capacitor design.

The author has gathered together under one cover much material which is often only to be found by searching through many separate references.

The two opening chapters deal with Service Specifications, requirements, colour codes, tables of insulating materials and their properties, etc: Chapter III on measurements has a number of interesting methods described for working at v.h.f. such as the Hartshorn and Ward method of measuring loss angle of flat solid disk dielectrics and liquid dielectrics, also resonant-cavity methods for frequencies from 3 000Mc/s upwards.

Paper, mica, ceramic, glass and glaze dielectric capacitors receive attention in Chapter IV, while chapter V gives information on construction and the effects on design of temperature and frequency. Following the latter is much useful data concerning capacitors with plastic and loaded plastic dielectrics, electrolytic capacitors and those of air-filled, gas-filled and vacuum construction. Experimental capacitors are also included. The final chapters concern faults that occur and future developments.

In fact, it is only by dipping into this bibliological cornucopia that one realizes how much information is at the disposal of the user.

R. H. MAPPLEBECK

Quartz Crystals as Oscillators and Resonators

By D. Fairweather and R. C. Richards. 54 pp. 52 figs. Demy 8vo. Marconi's Wireless Telegraph Co. Ltd. 1956. Price 7s. 6d.

THIS paper, the second in the Marconi Review Monograph Series, has been compiled by the Crystal Production department of Marconi's Wireless Telegraph Co. Ltd., to provide information for potential users of quartz crystals. The treatment given is not intended to be highly theoretical.

Both the authors have had many years of experience in providing crystals of all types for a variety of requirements and it is mainly from the practical point of view that the paper has been written.

An Approach to Modern Physics

By E. N. da C. Andrade. 232 pp. 40 figs. Demy 8vo. G. Bell & Sons Ltd. 1956. Price 25s.

THE author's earlier book, *Mechanism of Nature*, originally appeared in 1930, and the fifth edition in 1936. Since that date discoveries and developments of the first importance have been made in physics. With this in view the book has been completely revised and so considerably extended that it now retains little of the text of its predecessor in unaltered form and is nearly twice as long.

The Physics of Nuclear Reactors

110 pp. 50 figs. Demy 4to. The Institute of Physics. 1956. Price 25s.

THE papers discussed in this supplement to the *British Journal of Applied Physics*, with the discussion on them, comprise the proceedings of a Conference arranged by The Institute of Physics in collaboration with the British Nuclear Energy Conference and held in London from 3 to 6 July 1956.

The purpose of the Conference was to keep members of the Institute of Physics and its sister Institutions abreast of scientific problems needing solution.

Electronic Computers Principles and Applications

Edited by T. E. Ivall. 167 pp. 40 figs. Demy 8vo. Iliffe & Sons Ltd. 1956. Price 25s.

THIS non-mathematical introduction to the principles and applications of computers employing valves and other electronic devices is primarily written for technicians, engineers and students with a knowledge of electricity or electronics.

Both digital and analogue computers are covered, and comparisons are made between the two. The main part of the book, is devoted to describing their circuits and construction; their rapidly developing applications in industry, commerce and science are also outlined.

An Introduction to Cybernetics

By W. Ross Ashby. 295 pp. 40 figs. Demy 8vo. Chapman & Hall Ltd. 1956. Price 36s.

THERE exists a great need for introductory textbooks written in such a manner that the author leads the reader into the subject, rather than attempting to present him with an overall picture, which must, of necessity, be superficial. Such a book has the enormous advantage that in addition to providing a formal introduction to a particular field, it also displays, by example, the approach of the scientist himself to the subject, the manner in which he selects from the available data those parts which appear relevant, the way in which he introduces those concepts in terms of which his theory is to be constructed. At a time when the uninitiated must wonder at the way the scientist produces mathematical formulae purporting to describe such apparently un-mathematical subjects as military strategy and biology, it is more than ever imperative that the layman should have some idea of the power and the range of validity of the scientist's methods.

In many respects Dr. Ashby's new book, *An Introduction to Cybernetics,* fulfils such an ideal. The book is well written and lucid, and contains many detailed exercises, together with the answers, by which the reader can test his understanding of the main body of the text as he proceeds. The differential calculus is not required and nothing more than elementary algebra is demanded of the reader as Dr. Ashby develops his subject from familiar everyday ideas.

Nevertheless, one is forced to doubt whether this book is, in fact, an introduction to cybernetics, as that word is understood by other specialists. What Dr. Ashby offers might perhaps be described as an abstract theory of machines, or of systems displaying machine-like behaviour. He discusses in general terms such concepts as equilibrium, regulation, coupling, and amplification (although he relies on the reader buying his previous book *Design for a Brain* to obtain illumination on his apparently central concept of "ultrastability"). At several points Dr. Ashby's terminology is at variance with that used in the other literature of the subject. While such variations are common in a young and fast developing subject, the ambitious reader may well be confused when he attempts to explore the subject further. On this point, it should perhaps be mentioned that a large proportion of those books given in the list of references are liable to prove difficult reading, even after thoroughly digesting this book.

In his preface Dr. Ashby states that his book is intended to help those students of the biological sciences who wish to acquire some understanding of cybernetics. While the reviewer feels that such students should find the book of interest for its own sake, he cannot recommend it to those with a training in the physical sciences as the introduction to cybernetics which it is claimed to be.

D. L. HURD

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

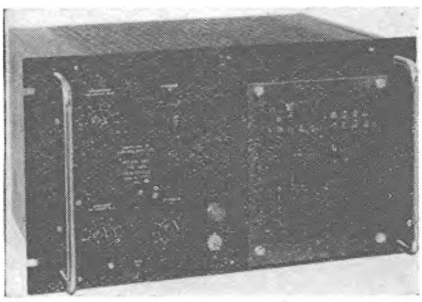
PULSE AMPLIFIER

(Illustrated below)

Dynatron Radio Ltd, The Firs, Castle Hill, Maidenhead, Berkshire

TYPE 1430A is a wide band linear amplifier designed for use with fast ionization chambers, proportional and scintillation counters. This equipment comprises three units, the main amplifier and stabilized power supplies, h.f. head amplifier for ionization chambers and proportional counters and an alternative cathode-follower head for scintillation counters.

The main amplifier has a maximum gain of 86dB reduceable in steps of 2dB to a gain of 46dB with an input impedance of 100Ω. It will deliver output pulses of +50V for not greater than 1 per cent non-linearity. Integrating and differentiating time-constants are selected by controls over the ranges of i.t.c. 0·08µsec to 8µsec in 7 steps and d.t.c. 0·08µsec to 8µsec in 7 steps plus 250µsec step. The h.f. head amplifier has a fixed



gain of 31·5dB and provision for input of either polarity. The alternative cathode-follower head unit has an insertion loss of 23dB and also has provision for input of either polarity. Both these head units incorporate e.h.t. filter circuits.

H.T. and l.t. supplies for the whole equipment are stabilized and the amplifier circuits incorporate negative feedback enabling a gain stability of the order of 1 per cent.

HUMIDITY SENSING ELEMENT

Distributed by B & K Laboratories Ltd, 57 Union Street, London, S.E.1

THE El-Tronics humidity sensing element is capable of accurately measuring relative humidity over a wide range of temperatures. It employs a conducting plastic whose electrical resistance changes in proportion to the relative humidity of the surrounding atmosphere.

The element embodies printed electrodes on each face and measures $\frac{7}{8}$in, wide, $1\frac{1}{2}$in long and 1/32in thick. Tests have proved it effective for measuring and control applications.

Adsorption and not absorption is one of the properties of the element and reaction time is very short, the standard production element having a reaction time of 30sec to reach 67 per cent of a given humidity change. Reaction time can be reduced to one second for special applications.

Lack of degradation at elevated temperatures allows the use of the element to measure humidity at much higher temperatures than previously possible. This is due to a negligible temperature coefficient in the range from 50°C to 80°C (23°F to 176°F).

Long term stability and the maintenance of calibration without need for adjustment under adverse conditions are other features of this element. Effects of hysteresis are at a minimum and the element will repeat its reading within ±3 per cent.

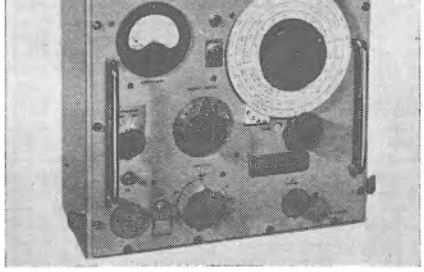
V.H.F. SIGNAL GENERATOR

(Illustrated below)

Advance Components Ltd, Roebuck Road, Hainault, Essex

THE Type D1P/2 is a special version of the v.h.f. signal generator model D1/D which has been produced with frequency ranges and performance most suitable for the alignment of narrow band communication receivers. A Colpitts type oscillator is employed with a variable capacitor driven by a slow motion drive with a 50:1 reduction. The frequency calibration is displayed on a 6in diameter dial and a linear scale and vernier are provided to ensure maximum resetting accuracy. The attenuator system consists of a continuously variable control and a five-position step attenuator. To eliminate attenuator reaction on the oscillator frequency, a buffer stage separates the oscillator from the attenuator. The output signal may be amplitude modulated 30 per cent at 1kc/s using a crystal modulator at the output, thus preventing incidental frequency modulation. To provide an accurate check of the frequency calibration, a 2Mc/s oscillator is incorporated which when in use ejects a 2Mc/s signal with the r.f. output providing marker signals at multiples of 2Mc/s, which can be detected by the receiver under test.

Frequency: 2 to 190Mc/s in six ranges. Accuracy ±0·01 per cent.
R.F. output: Continuously variable



from 0·1µV to 10mV. Accuracy ±(3dB+1µV).
Output Impedance: 75Ω unterminated, 37·5Ω terminated.

DIGITAL MICROSECOND CHRONOMETER

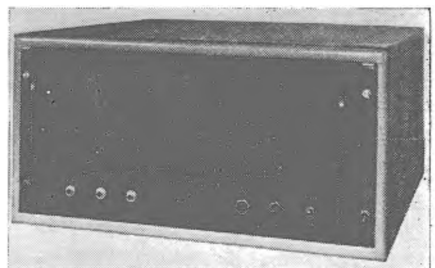
(Illustrated below)

Racal Engineering Ltd, Western Road, Bracknell, Berkshire

MICROSECOND chronometer type SA.45 is designed to measure the time interval between two events with an accuracy of ±1µsec.

Direct reading visual display of the elapsed time interval is provided and the reading is held until the instrument is manually reset.

The instrument employs digital techniques throughout, thus reducing to a minimum any possibility of ambiguity of result and its range extends from 3µsec to 1sec. Longer time intervals can be measured on the standard equipment and, by the addition of a mechanical counter



the display can be extended by as many as six digits.

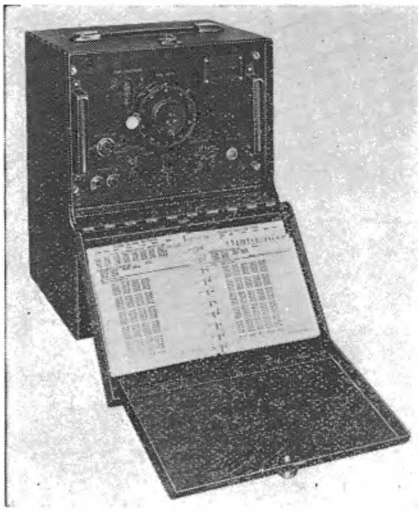
Two inputs are provided; a single line start/stop input for use when the electrical impulses which represent the beginning and end of the period to be measured are derived from a common source, and a two-line input for applications where two separate electrical sources provide the starting and stopping pulses.

The instrument may also be used as a counter and totalizer. Random or regular pulses or a sine wave at frequencies up to 1Mc/s will be counted and indicated on the six decade display. Furthermore, such external generated pulses or frequencies can be counted in the period between any two externally derived pulses.

Indication.
6 digit illuminated figure.

Input required, single line start/stop.
+15V pulse, 0·5µsec rise time, 1µsec duration. 2µsec (min) between pulses. Impedance 47kΩ and 15pF.

Input required double line start/stop.
+15V pulse, 0·5µsec rise time, 3µsec duration (max). 2µsec (min) between pulses. Impedance 47kΩ and 15pF.



FREQUENCY METERS

(Illustrated above)

Telemechanics Ltd, 3 Newman Yard, Newman Street, London, W.1

TYPES 7474 and 7475 frequency meters are used for laboratory or field testing and measuring, pulsed, modulated or c.w. r.f. transmitters, receivers and signal-generators. They are produced in the ranges 20 to 250Mc/s and 85 to 1 000 Mc/s respectively.

Signals which are picked up by the antenna beat against the output of a variable frequency heterodyne oscillator, which is tuned until the beat tone heard in the head telephone approaches zero beat. The dial reading of the heterodyne oscillator is then converted into frequency by means of the calibration chart provided. A crystal-controlled oscillator of fixed frequency is included to provide a calibration standard. Signals are radiated by the heterodyne oscillator, enabling the instrument to be used as a signal generator. Power required for the operation of this equipment is supplied by standard l.t. and h.t. cells and a mains-operated power unit can be supplied as an optional extra. This is designed to fit into the battery compartment.

The instrument is completely enclosed in a two-section case. The top section of the case contains the instrument chassis; the other section, which opens at the bottom rear, accommodates the batteries.

Power range: Input 20mV to 2V.
Input impedance: Adjustable through antenna coupling.
Signal output: Variable audio level to headphones.
Accuracy: ·002 per cent at 25°C, 0·01 per cent between −20°C and +70°C.
Fundamental frequency: Type 7474, 20 to 40Mc/s.
Type 7475, 85 to 200Mc/s.
Internal Modulation: 1kc/s.
Temperature range: −40 to +70°C.

U.H.F. TRIODE

The General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

AN indirectly heated, high slope, low noise, triode, the A2521, especially suitable as an r.f. amplifier for use at frequencies in the region 500 to 1 000Mc/s has been developed.

The valve has a slope of 12mA/V and an anode dissipation of 2·5W. In normal receiver applications it is free from microphony. At 500 and 900Mc/s the noise factors are 9 and 12dB respectively.

When operating in a u.h.f. grounded-grid amplifier circuit, the power gain and bandwidth may be adjusted by altering the coupling between the anode line and the output loop. At 900Mc/s, for power gains of 6 and 16dB, the available bandwidths are 80 and 4Mc/s respectively.

TELEGRAPH DISTORTION MEASURING EQUIPMENT

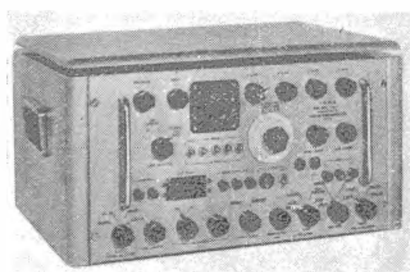
(Illustrated below)

Winston Electronics Ltd, Shepperton, Middlesex

ALL the facilities for the complete setting and testing of telegraph equipment are contained in the one cabinet of Type 603.

It will transmit, when required, several kinds of teleprinter signals, including a pre-set code character at varying speeds, which can be accurately measured.

Test signals can be sent and a known amount of start signal distortion can be introduced when testing teleprinter apparatus for margin. The equipment will monitor any telegraph signal on working lines with only a negligible insertion loss and will monitor its own transmitted signal after the signal has been transmitted over a line and back.



It will accurately set high speed telegraph relays via a special socket. The relays can be set for neutrality and transit time, which can be measured on the screen of the instrument.

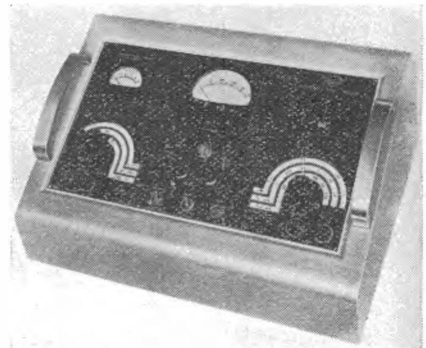
TRANSISTOR TESTER

(Illustrated above right)

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

TO cater specifically for transistorized equipment production and service workshops the simplified transistor tester type L.264 has been produced. Compact, and as easy to use as a conventional valve tester, the instrument affords a ready means of checking three of the more important junction transistor parameters: base-collector short-circuit current gain; d.c. collector current for zero base current; and collector turnover voltage. All measurements are presented as direct meter readings.

Collector voltage range: 0-20V, metered. Collector current range: 0-2·5A, metered. a' range: 0-100 Accuracy ±5 per cent. $I'_{c(o)}$ range: 0-2·5A, in six ranges. Accuracy ±5 per cent. Turnover



voltage range: dependent on permissible collector dissipation of transistor: 0-18V on 2·5mW range; 0-85V on 25 and 250mW ranges.

WAFER SWITCH KITS

Radiospares Ltd, 4-8 Maple Street, London, W.1

WITH the 'Maka-Switch' kits any switch combination up to six banks can be quickly built. They comprise a shafting assembly of control unit, side struts, locknuts and washers; spacers, for separating the wafers; and wafers, available in the following combinations:

1 pole 12-way, 2 pole 6-way, 3 pole 4-way, 4 pole 3-way, 6 pole 2-way.

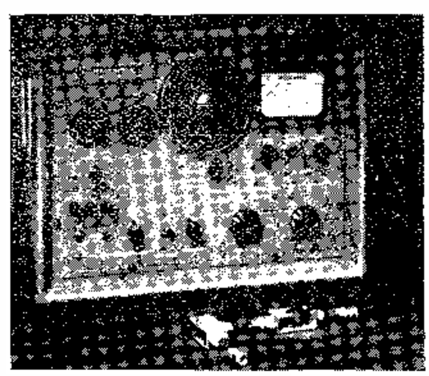
F.M./A.M. SIGNAL GENERATOR

(Illustrated below)

Marconi Instruments Ltd, St. Albans, Hertfordshire

THE f.m./a.m. signal generator TF 995A/2 is the latest of the TF 995 series. The frequency range of this instrument extends from 1·5 to 220Mc/s in five bands and there is a built-in crystal calibrator for use above 13·5Mc/s. A precision slow-motion mechanism is employed for the main tuning drive, and a directly-calibrated incremental tuning control for making bandwidth measurements has been incorporated.

The open-circuit output level is variable, in 1dB steps, from a minimum of 0·1μV to a maximum of 100mV at 52Ω and 200mV at 75Ω. The output may be c.w., frequency modulated, amplitude modulated, or simultaneously both frequency and amplitude modulated. The modulation, obtained either from an internal 1kc/s oscillator or from an external source, is variable to maximum frequency deviations ranging from 25 to 600kc/s for f.m., and to depths up to 50 per cent for a.m.



Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 27 March. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Disc Recording.
By: G. F. Dutton.

North Western Section

Date: 7 March. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester, 1.
Lecture: Electronic Musical Instruments.
By: Alan Douglas.

Merseyside Section

Date: 6 March. Time: 7 p.m.
Held at: The Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool 3.
Lecture: The Preparation of Service and Technical Data Sheets.
By: E. A. W. Spreadbury.

Scottish Section

Date: 15 March. Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University of Edinburgh.
Lecture: Some Applications of Nucleonics in Medicine.
By: E. W. Pulsford and N. Veall.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Radio and Telecommunication Section

Date: 4 March.
Informal evening on Electronics in Administration—a Survey.
Talk by: D. C. Espley.
Date: 20 March.
Lecture: Recent Developments in X-Ray and Electron Microscopy with some Applications to Radio and Electronics.
By: V. E. Cosslett and C. W. Oatley.

Ordinary Meetings

Date: 7 March.
Lecture: Cathodic Protection.
By: L. B. Hobgen, A. Spencer and P. W. Heselgrave.
Date: 21 March.
Lecture: A Cyclotron for Medical Research.
By: J. W. Gallop, D. D. Vonberg, R. J. Post, W. B. Powell, J. Sharp, and P. J. Waterton.

Measurement and Supply Sections

Date: 12 March.
Lectures: An Analogue Computer for Nuclear Power Studies.
and: The Application of Analogue Methods to Compute and Predict Xenon Poisoning in a High-Flux Nuclear Reactor.
By: G. J. R. MacLusky.

Informal Meeting

Date: 18 March.
Discussion: What is Limiting the Application of Servomechanisms in the Electrical Industry.
Opened by: C. Ryder.

Supply Section

Date: 27 March.
Lecture: Earth Electrode Systems for Large Electric Stations.
By: J. D. Humphries.

East Midland Centre

Date: 5 March. Time: 6.30 p.m.
Held at: E.M.E.B. Service Centre, Derby.
Lecture: Electrolytic Processes for Surface Conditioning of Metals.
By: J. W. Cuthbertson.
Date: 14 March. Time: 7.30 p.m.
Held at: George Hotel, Kettering.
Lecture: Research in Industry.
By: D. R. Hardy.
Date: 19 March. Time: 6.30 p.m.
Held at: The College of Arts and Crafts, Waverley Street, Nottingham.
Lecture: The Control and Instrumentation of a Nuclear Reactor.
By: A. R. Gillespie.

East Anglian Sub-Centre

Date: 5 March. Time 8 p.m.
Held at: The Cavendish Laboratory, Cambridge.
Lecture: Cathodic Protection.
By: L. B. Hobgen, K. A. Spencer and P. W. Heselgrave.

North-Eastern Centre

Date: 8 March. Time: 6.15 p.m.
Held at: Neville Hall, Newcastle-on-Tyne.
Lecture: The Work of the Royal Navy Electrical Branch at Sea.
By: J. C. Turnbull.
Date: 11 March. Time: 6.15 p.m.
Held at: Neville Hall, Newcastle-on-Tyne.
Lecture: Mechanical Strength of Power Transformers in Service.
By: E. T. Norris.
Date: 25 March. Time: 6.15 p.m.
Held at: Neville Hall, Newcastle-on-Tyne.
Lecture: Earth Electrode Systems for Large Electric Stations.
By: J. D. Humphries.

North-Eastern Radio and Measurements Group

Date: 4 March. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: An Experimental Study of High-Permeability Nickel-Iron Alloys.
By: C. E. Richards, E. V. Walker and A. C. Lynch.
Date: 18 March. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Power System Protection, with particular reference to the Application of Junction Transistors to Distance Relays and A Dual-Comparator Mho-Type Distance Relay Utilizing Transistors.
By: C. Adamson and L. M. Wedepohl.

North Midland Centre

Date: 25 March. Time: 6.30 p.m.
Held at: The Town Hall, Leeds.
Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.

North-Western Centre

Date: 5 March. Time: 6.15 p.m.
Held at: Engineers' Club, 17 Albert Square, Manchester.
Lecture: D. C. Winder Drives using Mercury Arc Rectifier/Inverters.
By: L. Abram, J. B. McBreen and J. Sherlock.

North-Western Radio and Telecommunication Group

Date: 13 March. Time: 6.45 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: The Electronic Age.
By: R. C. G. Williams.

North Lancashire Sub-Centre

Date: 13 March. Time: 7.15 p.m.
Held at: North Western Electricity Board Demonstration Theatre, 19 Friargate, Preston.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.

Northern Ireland Centre

Date: 5 March. Time: 7.30 p.m.
Held at: Whitla Hall, Queen's University, Belfast.
Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.
Date: 12 March. Time: 6.30 p.m.
Held at: Queen's University, Belfast.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.

North Scotland Sub-Centre

Date: 13 March. Time: 7.30 p.m.
Held at: Caledonian Hotel, Aberdeen.
Lecture: Automatic Circuit Reclosers.
By: G. F. Peirson, A. H. Pollard and N. Carr.
Date: 14 March. Time 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Repeat of above lecture.

South-East Scotland Sub-Centre

Date: 19 March. Time: 7 p.m.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.
Date: 28 March. Time: 7 p.m.
Held at: Central Hall, Edinbugh.
Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.

South-West Scotland Sub-Centre

Date: 27 March. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: The BBC Sound Broadcasting Service on Very-High Frequencies.
By: E. W. Hayes and H. Page.

South Midland Centre

Date: 11 March. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Lectures: Vacuum Techniques.
By: D. J. Taylor.
Temperature Control.
By: J. P. King.
Cutting of Semi-Conductors.
By: J. R. Walford (at Malvern Winter Gardens).

South Midland Radio and Telecommunication Group

Date: 25 March. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: The BBC Sound Broadcasting Service on Very-High Frequencies.
By: E. W. Hayes and H. Page.

Rugby Sub-Centre

Date: 12 March. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts, Rugby.
Lecture: Ferrites.
By: F. Brailsford.

Southern Centre

Date: 1 March. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Analogue Computers.
By: R. Postlethwaite.
Date: 6 March. Time: 6.30 p.m.
Held at: Central Electricity Authority, 111 High Street, Portsmouth.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.
Date: 13 March. Time: 6.30 p.m.
Held at: Brighton Technical College.
Lecture: Heat Pumps.
By: Miriam V. Griffith.
Date: 27 March. Time: 7.30 p.m.
Held at: R.A.E. Technical College, Farnborough.
Lecture: Frequency-Modulation Radar for Use in the Mercantile Marine.
By: D. N. Keep.
Date: 29 March. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Nuclear Power.
By: H. A. Roberts.

Western Centre

Date: 11 March. Time: 6 p.m.
Held at: South Wales Institute of Engineers, Park Place, Cardiff.
Lectures: Transistor Power Amplifiers.
By: R. A. Hilbourne and D. D. Jones.
and: Transistor d.c. Convertors.
By: L. H. Light and P. M. Hooker.

Western Supply Group

Date: 18 March. Time: 6 p.m.
Held at: Demonstration Theatre, South Western Electricity Board, Colston Avenue, Bristol.
Lecture: Power Stations.
By: P. L. Lutte.
and: Distribution.
By: A. G. Milne.

Irish Branch

Date: 21 March. Time: 6 p.m.
Held at: The Physical Laboratory, Trinity College, Dublin.
Lecture: Variable Speed a.c. Drives.
By: J. A. V. McEvoy.

Reading District

Date: 18 March. Time: 7.15 p.m.
Held at: George Hotel, King Street, Reading.
Lecture: The Transatlantic Telephone Cable.
By: R. J. Halsey.

THE RADAR ASSOCIATION

Date: 13 March. Time: 7.30 p.m.
Held at: The Anatomy Theatre, University College, Gower Street, London, W.C.1.
Lecture: Radar Techniques in Science and Industry.
By: Sir John Cockcroft.

PUBLICATIONS RECEIVED

HI-FI FROM MICROPHONE TO EAR by G. Slot is a book which has been written with the purpose of meeting the demands for advice from various parts of the world on the best means of recording and reproducing speech and music. Philips Technical Library, Holland. Cleaver-Hume Press Ltd., 31 Wright's Lane, London, W.8. Price 17s. 6d.

NOTES ON APPLIED SCIENCE No. 15—APPLICATION OF SPRING STRIPS TO INSTRUMENT DESIGN gives designers a knowledge of the subject which will enable them to make use of spring strips in solving certain problems in instrument design. Spring strips of tempered steel or other elastic materials have many uses in the design of all types of measuring instruments. Her Majesty's Stationery Office (for Department of Scientific and Industrial Research), price 2s.

INTRODUCTION TO PRINTED CIRCUITS by R. L. Swiggett and PICTURE BOOK OF TV TROUBLES VOL. 7 are two recent additions to the publications of John F. Rider Publisher, Inc. 480 Canal Street, New York. Prices \$2.70 and \$1.50 respectively.