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Commentary

THE 44th Physical Society Exhibition will be in full swing, at the Old and New Halls of the Royal Horticultural Society at the time that this issue is going to press. A review of the Exhibition will be included in next month's issue. This is not only the first exhibition of the year, but the first exhibition of the 1960's and it will, no doubt, providing our perception is sufficiently good, provide us with a small inkling of what we may expect from the coming decade. It would, however, take a brave, and perhaps foolhardy, man to make a forecast of scientific development in the next decade, as indeed it would have done in 1950 for, although the signs may be there, as likely as not they will appear as insignificant trivialities to all but the most gifted of crystal-ball gazers.

The most dramatic scientific advances during the last ten years have, no doubt, been the harnessing of nuclear power for the generation of electricity, the advent of supersonic flight and the launching of vehicles into outer space. As far as electronic engineering is concerned the many advances have had, perhaps, less impact on the layman but to those engaged in the profession they have been just as great and their effects equally far reaching.

The greatest single advance in electronic components with which the 1950's provided us is undoubtedly the evolution of the transistor from a somewhat refractory laboratory 'toy' to a mass-produced and thoroughly reliable component. In connexion with the transistor it is interesting to recall that at the 1950 Physical Society Exhibition, which was of course held at Imperial College, one firm exhibited examples of germanium and germanium oxide. They also had on exhibition an enlarged model showing the constructional details of an improved type 'crystal triode' in which the two 'catswhiskers' were replaced by two plates separated by a known and accurately controlled spacing. At the Exhibition now in progress diffused base germanium transistors are being shown with alpha cut-off frequencies up to 600Mc/s and diffused base silicon transistors with alpha cut-off frequencies up to 100Mc/s, while germanium power transistors are being shown having maximum collector currents up to 25A and voltage ratings up to 80V. Also being shown are microminiature semiconductor solid circuits. These are formed by the selection and shaping of conduction paths on and through a semiconductor wafer. One example exhibited is a multivibrator circuit measuring some ¼in by ⅛in by 1/32in with a performance identical to that of a conventional circuit containing eight resistors, two capacitors and two transistors; altogether representing an equivalent component packing density of 30 000 000 parts per cubic foot! These semiconductor solid circuits use single crystal semiconductors as the basic material, the circuit being formed by diffusion, metallic evaporation, alloying and chemical forming. Much of the evaporation and etching is controlled by photographic masks and these can readily be changed, allowing the economical production of even small quantities of circuits specifically designed for a given function. It may well be that these semiconductor solid circuits hold the same place at the 1960 Exhibition as did the jars of germanium dust and the point contact model in 1950.

Turning now from the smallest of components to some of the largest of equipments, the digital computer readily springs to mind and it is, indeed, a field in which the 1950's have seen tremendous developments. At the 1950 Exhibition part of the Pilot Model of ACE was shown. A year later the Manchester University Automatic Digital Computer was inaugurated: this occupied two bays 16ft long, 8ft high and 4ft wide; it contained 3 500 valves, used twelve cathode-ray tube stores and a magnetic drum and had a total power consumption of some 27kW. It was not until 1954 that the first commercial 'office' installation was made in this country, the machine then used employing some 5 000 valves and having a power consumption of 30kW. At this year's Exhibition a fully transistorized digital computer for general office use is shown; it is little larger than an office desk, has a 1 000 word main store with provision for extension to 10 000 words, a basic bit frequency of 500kc/s and it is stated that it may be plugged into any 230V 5A socket. To draw such a comparison is, of course, to grossly exaggerate the present position; the capacity and speeds of the machines being of somewhat different orders. The point to note is that the mercury delay line and the c.r.t. memory have largely given way to the magnetic core matrix and magnetic tape while the valve has largely succumbed to the transistor, but, and perhaps above all in importance is the fact that the computer, which in 1950 was a mystery to all but a few, and little more than a subject for music hall jokes for the layman, is now an accepted part of modern office and factory equipment.

And so, as one goes round the 1960 Physical Society Exhibition one is struck by the vast progress which the last ten years has seen and, if one gives thought to the matter, the enormity of the problem of conjecturing on what the next ten years may bring forth.

But to finish on a lighter note, probably the device which has made the great visual difference to the Physical Society's Exhibition is the Dekatron*: in 1950 they were virtually unknown—today they wink at one from every corner of the hall!

* *First described in* ELECTRONIC ENGINEERING *in May 1950.*

A Review of Colour Television in the U.K.

By R. D. A. Maurice*, Ing. Dr., Ing. E.S.E., A.M.I.E.E.

Since 1953 the British Broadcasting Corporation have taken an active interest in colour television and in the intervening years they have carried out a considerable amount of work in determining the adaptability of the American N.T.S.C. standards to the British 405 line system. A number of reports and a BBC Engineering Division Monograph have been published on this work: the present article provides a concise summary for those already versed in colour television technology

(Voir page 126 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 131)

IN 1953 the BBC took an active interest in colour television and from then until 1958 a great deal of work was done with the object of investigating the suitability of the American N.T.S.C. colour television adapted to the 405-line standard. A number of reports were written culminating in a BBC Engineering Division Monograph[1]. A number of lectures on colour television have been delivered to various polytechnic institutes and other bodies, and for these a need was felt to summarize the results in a manner suitable for readers already versed in colour television technology. This article is intended to fill the role of such a summary.

Compatibility

In order to achieve complete compatibility the colour signal must be capable of using the same channel allocation, be receivable in black-and-white on existing receivers and preferably be capable of using existing monochrome links and transmitters. The colour receiver should be capable of receiving a standard monochrome transmission.

Buzz on Sound

The monochrome receiver sound-channel selectivity at 840kc/s above sound carrier frequency must be adequate.

In a field test involving an Alexandra Palace transmitter radiating 5kW e.r.p. of vision and 420 members of the public (BBC Audience Research viewing panel) the figures shown in Table 1 emerged.

TABLE 1

PRESENCE OF CHROMINANCE SIGNAL MADE BUZZ ON SOUND	PERCENTAGE OF VIEWER-LISTENERS
no different	95
slightly louder	4
moderately louder	1
very much louder	0

Pattern Interference Due to Beat Between Sound and Chrominance Carriers

The difference between the sound and chrominance carriers is 3 500-2 660 = 840kc/s (approx.) and an intermodulation product at 840kc/s, if produced somewhere in the r.f. or i.f. portions of a black-and-white receiver, could give rise to a relatively coarse-structure interference pattern. Steps were taken before any field trials were undertaken to minimize such interference by ensuring that the frequency of it would be, like that of the chrominance signal, an odd multiple of half the line scan frequency. Thus the difference between the vision and sound carrier frequencies would have to be an even multiple of half the line scan frequency or an integer multiple of the line scan frequency. This was arranged to take place at the transmitter itself by ensuring that

$$f_v - f_s = 4/3\ (f_v - f_c)$$

Table 2 shows four possible schemes, the fourth being the one adopted during the field trials. The first scheme shown in the table does not use integral division ($2m + 1$ not an integer) and does not have the advantage of a 'locked' relationship between chrominance and sound carrier frequencies, but it is simpler to instrument since it relies on the accuracy and constancy of crystal-generated frequencies.

The results of this particular scheme (No. 4) is to change, very slightly, the difference between sound and vision carriers from the standard value of 3 500kc/s to 3 543·75kc/s. The sound carrier frequency is left undisturbed and the vision carrier frequency becomes increased by 44kc/s—a negligible amount.

In a field test involving 484 questionnaires completed by engineer-viewers and using a vision e.r.p. of 5kW, only 1

TABLE 2
Various Means of Minimizing Interference between Chrominance, Vision and Sound Carriers

f_f (c/s)	f_L (kc/s)	f_c-f_s (kc/s)	f_v-f_s (kc/s)	f_v-f_c (kc/s)	$2m+1$	$2n+1$	q	$\frac{f_v-f_s}{f_v-f_c}$	SCHEME NUMBER
50·0984076	10·14492754	836·9575	3500	2663·0425	165·0001927	525	345	46/35	1
,,	,,	836·956522	,,	2663·043478	165	,,	,,	46/35	2
50	10·125	835·3125	3493·125	2657·8125	,,	,,	,,	46/35	3
,,	,,	885·9375	3543·75	,,	175	,,	350	4/3	4

f_s=sound carrier frequency
f_v=vision carrier frequency
f_f=field frequency
$f_c-f_s=(2m+1)\,f_L/2,\ f_v-f_s=q\,f_l,\ f_v-f_c=(2n+1)\,f_L/2$

f_c=chrominance carrier frequency
f_L=line frequency
m, n and q should be integers

* *The British Broadcasting Corporation*

per cent found the beat pattern intolerable, while 90 per cent found it negligible (0·9 per cent in 'tolerable' class).

Vertical Bars Due to Presence of Colour Burst in the Line Blanking Interval

Few receivers are fitted with line retrace black-out arrangements so it might be thought that the burst would appear as a curtain with vertical folds in it on the extreme left-hand side of the image. The burst peaks correspond to 15 per cent/0·7 = 21 per cent of peak-white signal (neglecting picture 'sit' level). In a laboratory test using 60 observers half of whom were not technically trained and 24 domestic monochrome receivers of various makes, two observers found the 'curtain' effect unacceptable. The scale of criteria used was:—

invisible
just visible
easily visible
unacceptable

Many of the observers disagreed as to which receivers showed the effect and although it can be seen on some receivers when a white caption on a black background is transmitted, it is not considered to have significance during a normal transmission.

In a field test involving 424 members of the public, no viewers found the curtain effect 'moderately annoying' or 'very annoying'.

Synchronizing Trouble Due to the Burst

The presence of the burst in the post line-synchronizing blanking interval can cause loss of synchronism in some receivers; however, in a field test using 424 members of the public 94 per cent observed no trouble.

Hum Bars Due to Asynchronous Working

The standard N.T.S.C. system operates with a crystal controlled sub-carrier and since its frequency must be an odd multiple of half the line scan frequency (to minimize the dot-pattern interference created in monochrome receivers by the presence of the chrominance signal) it follows that the latter cannot be allowed to vary in synchronism with the frequency of the electricity supply mains. When the latter departs appreciably from 50c/s, hum bars appear on the received image due to inadequate smoothing, etc., in receivers and, to a lesser extent, in transmission equipment.

At the time when the field trials were in progress, the stability of frequency of the electricity supply was inadequate to satisfy the requirements of the N.T.S.C. system—a departure of ±0·2c/s from the nominal 50c/s could scarcely be tolerated and unless mains frequency stability improves, it will be necessary to adopt one of two possible means of overcoming this difficulty. One means* involves changing the 'count-down', from sub-carrier frequency to field frequency, in steps when the mains frequency departs from the field frequency by a sufficient amount. For example, if f_f is the field frequency, f_L the line frequency and $f_v - f_c$ the chrominance sub-carrier frequency

$$f_f = f_L/202\tfrac{1}{2}$$
$$= \frac{f_L/2}{101\frac{1}{4}}$$

but $f_L/2 = (f_v - f_c)/(2n + 1)$

$$\text{so}\quad f_f = \frac{(f_v - f_c)/(2n + 1)}{101\frac{1}{4}}$$
$$= \frac{f_v - f_c}{405(2n + 1)/4}$$

now let $f_v - f_c = 2\,657{\cdot}8125$kc/s.

* *Patent Nos. 794908 and 798083.*

Then Table 3 may be constructed.

TABLE 3

f_F	$2n+1$
50·1912	523
50	525
49·8102	527

Thus, as soon as the mains frequency exceeds 50c/s by 0·19c/s the 'count-down' factor may be changed from 525 to 523 and so forth.

Another method† involves some additional complication in the colour receiver and will not be discussed here.

Lack of 'Panchromaticity' Owing to the Use of Gamma-Corrected Primary-Colour Signals Rather than a Gamma-Corrected Luminance Signal

The luminance signal in the N.T.S.C. system is

$$E_Y' = 0{\cdot}3E_R^{1/\gamma} + 0{\cdot}59E_G^{1/\gamma} + 0{\cdot}11E_B^{1/\gamma}$$

instead of $(0{\cdot}3E_R + 0{\cdot}59E_G + 0{\cdot}11E_B)^{1/\gamma}$. The latter signal, if received by a monochrome receiver, would be transformed into display-tube beam current and hence brightness as

$$[(0{\cdot}3E_R + 0{\cdot}59E_G + 0{\cdot}11E_B)^{1/\gamma}]^\gamma = 0{\cdot}3E_R + 0{\cdot}59E_G + 0{\cdot}11E_B$$

which truly represents brightness or luminance; however, the N.T.S.C. luminance signal E_Y' when raised to the gamma by the monochrome display will become

$$(E_Y')^\gamma = (0{\cdot}3E_R^{1/\gamma} + 0{\cdot}59E_G^{1/\gamma} + 0{\cdot}11E_B^{1/\gamma})^\gamma$$

Whenever a grey colour is being transmitted all is well because

$$E_R = E_G = E_B = K \leqslant 1$$

whence $\quad (E_Y') = [K^{1/\gamma}\,.\,(1)]^\gamma = K$

because $0{\cdot}3 + 0{\cdot}59 + 0{\cdot}11 = 1$ and the true luminance is obtained.

Consider, now, the transmission of a saturated blue scene for which $E_R = E_G = 0$, $E_B = 1$

$$(E_Y') = (0{\cdot}11)^\gamma$$
$$= 0{\cdot}004 \text{ if } \gamma = 2\tfrac{1}{2}$$

instead of 0·11 had the true luminance been reproduced. This is 1/36th of the correct value and it is seen that the reproduced black-and-white image is no longer panchromatic.

If the monochrome receiver is fully responsive to signals at sub-carrier frequency, however, it is found that the presence of the chrominance signal forms an effective compensation for this error. The curvature of the beam-current-to-grid-voltage characteristic of the display tube rectifies the chrominance signal, thus raising the mean brightness of coloured patches (in the original scene). This compensation can reduce the error considerably and, in the case of a saturated blue patch, the brightness can be raised from 1/36th to ½ of the correct value. Thus the presence of the dot structure due to the chrominance signal can be of use in restoring 'panchromaticity' to the monochrome picture.

In a field trial of picture 'quality' involving 424 members of the public a short colour film was transmitted twice, once with chrominance signal present and once without it. Sixty-one per cent of the viewers graded the film the same for the two transmissions and the net result of the change in gradings for the remainder of the viewers was that 11 per cent preferred to have the chrominance signal present. The grades were 'excellent', 'good', 'moderate', 'poor', 'very poor'.

† *Provisional Patent No. 17462/55.*

INTERFERENCE CAUSED BY THE CHROMINANCE SIGNAL

The dot pattern which appears on monochrome receiver screens in areas of the televised scene which are coloured is the least acceptable feature of the compatible picture. In other words, band-sharing is only moderately successful. The chrominance signal may be regarded as an interfering carrier at 2·66Mc/s below vision carrier frequency and for saturated red it is only 7·3dB below vision carrier amplitude. (Ratio of greatest chrominance carrier amplitude to peak-white vision carrier amplitude is −13·2dB not forgetting that the chrominance carrier is a single sideband of the vision carrier.) The arrangement whereby the chrominance signal frequency is made equal to an odd multiple of half the line scan frequency would be completely effective for any ratio of chrominance to luminance if the eye integrated completely over a twelfth of a second and if the display tube had a gamma of unity. Neither of these conditions is completely fulfilled and the result is that the 'odd-multiple' technique becomes decreasingly effective as the chrominance-to-luminance ratio increases.

In a field test involving 113 questionnaires completed by engineer-viewers, 10 per cent found the dot pattern moderately or very annoying, while of 424 members of the public, 5 per cent gave opinions in the same categories. Slides and films were used in these tests and if the most unfavourable opinions were noted for each slide and not averaged over the several slides, then the percentages of opinions in the 'annoying' classifications were higher. It is considered reasonable, however, to discount these maximum gradings in favour of averaged ones since it is probably not usual to classify an entire programme on the basis of the quality of one short portion of it. Some earlier tests involving 524 engineers gave somewhat more favourable answers; a total of between 3 and 4 per cent in grades: 'somewhat objectionable', 'definitely objectionable'.

The dot pattern interference is, of course, more visible on 21in receiver screens than on smaller ones and it may be desirable to fit 'notch' or band-eliminated filters in the video circuits of 21in receivers if colour broadcasting is adopted.

There are two reasons why this form of interference may be less obtrusive in the U.S.A. than in the U.K. First, the picture frequency is 30 a second and so flicker integration by the eye may be more effective (if brightness of American screens has not been raised to the flicker value as it has been in the U.K.). Secondly, it may be that American receivers are less responsive to signals at 3·59Mc/s (the frequency of the American chrominance sub-carrier) than are British receivers to signals at 2·66Mc/s.

The adoption of the system known as 'chrominance out of band' would completely eliminate dot-pattern interference, but it is scarcely feasible for use in Bands I and III as there is virtually no spare channel space.

CONCLUSIONS WITH REGARD TO COMPATIBILITY

Although the percentages of observers who found the dot-pattern interference in the 'annoying' category were somewhat high, the percentage obtained by taking all undesirable features together and grouping them in accordance with the subjective grades given by the observers were as shown in Table 4.

TABLE 4

TYPE OF VIEWER	MODERATELY ANNOYING (per cent)	VERY ANNOYING (per cent)	SOMEWHAT OBJECTIONABLE (per cent)	DEFINITELY OBJECTIONABLE (per cent)	UNUSABLE (per cent)
Engineers	8·8	2·6	4·8	1·8	0
Members of Public	5	1·2	—	—	—

This table is based upon classifying questionnaires in accordance with the least favourable average grade given by a viewer when a number of different still pictures were presented to him, or the actual least favourable grade given by an observer when viewing a motion picture programme.

It should be remembered that the presence of the dot pattern almost completely restores panchromaticity to the black-and-white picture and this may be why there was a surplus of 11 per cent of viewers who preferred the monochrome pictures when the chrominance signal was present.

From these figures and others, the BBC has concluded that the N.T.S.C. system is adequately compatible for transmission in Band I.

Features of Reception in Colour

PRESENCE OF DOT PATTERN AS IN COMPATIBLE RECEPTION AND THE EFFECT OF ITS REDUCTION ON THE SHARPNESS OF THE PICTURE

Colour receivers should contain a 'notch' or band-elimination filter to remove the chrominance signal from the luminance channel at a point after the composite signal channel has been divided into the two separate chrominance and luminance channels. The presence of chrominance signal in the display-tube beam currents will, in the colour receiver, not only cause a dot pattern to be visible, but will also raise the luminance of coloured areas by rectification as already explained. In a colour receiver, however, the individual primary-colour signals, each carrying its own correct luminance, are fed to each primary-colour display-tube gun and any increase of luminance due to rectification of chrominance signal will cause desaturation of coloured patches—a far more serious error than the loss of panchromaticity suffered in low-bandwidth monochrome receivers due to the reception of a luminance signal E_Y' composed of gamma-corrected primary colour signals instead of the reception of a gamma-corrected luminance signal.

In a number of different tests forming part of the field trials during which slides, film and television cameras were used, 11 per cent of opinions in the 'somewhat objectionable' or 'definitely objectionable' classes was the least favourable percentage. This figure resulted from closed-circuit laboratory tests representing high-field-strength and low-noise conditions. It is interesting to note that whereas motion pictures caused the dot crawl to be more apparent on monochrome compatible reception, this was by no means always the case during colour reception. Camera pictures were no worse than film or slides in causing dot-pattern interference. Of course, the dot-pattern interference varies in accordance with the saturation of the colours being transmitted. For some high-power (55kW e.r.p.) field trials during which film as well as camera 'shots' were chosen for programme interest rather than technical features (such as high saturation), the percentage of opinions in the 'objectional' categories did not exceed 1.

COLOUR SYNCHRONIZING DIFFICULTIES

Synchronous detection was rather feared by engineers who had had no experience of it, but although colour synchronizing difficulties were occasionally encountered, these were invariably due either to poor receiver design or to definite receiver faults. It can even be stated that, as Richman implies[3], there is more colour synchronizing information transmitted than is required and more than is transmitted in monochrome television for the scanning circuits.

Colour Trailing Due to Movement in the Scene

It was thought, at one time, that the red and green fluorescing phosphors in the display tube would have rather long afterglows, but in practice, a slight blue afterglow is visible exactly as in modern monochrome tubes. These produce white by a combination of blue and yellow and the latter has a shorter afterglow than the former.

Multiple-Image Effects Due to Multi-Path Propagation

It was feared that this form of interference might be more serious with a colour transmission with its chrominance information situated 2·66Mc/s away from the vision carrier frequency. This was found not to be the case. Trouble due to multi-path reception was completely negligible in channel 1, except during some tests involving a low-power transmitter at Kingswood, Surrey, and a mobile laboratory which undertook reception tests in the very hilly terrain in and around the North Downs. In this case 11 per cent of opinions were in one or other of the 'objectionable' categories.

Colour Fidelity Including Reverse Compatibility

This feature is highly dependent upon the colour fidelity of film and slides used in the tests. That this has been revealed by the tests shows that photographic colour fidelity leaves something to be desired—at least if very careful selection of material is not undertaken. The replacement of one colour film by another caused the percentage of 'objectionable' opinions to fall from 32 to 3 in a certain high-power field trial. In a final series of tests the 'objectionable' opinions were:

Motion picture film: 9 per cent. (35mm: 2 per cent, 16mm: 13 per cent).

Slides: 4 per cent.

Camera pictures: 14 per cent.

These relatively high percentages indicate that three-tube camera and three-gun receiver colour balances are not easy to achieve and furthermore, receiver stability as regards the primary-colour signal channels is not all that it might be.

With regard to reverse compatibility, that is, the ability of a colour receiver to reproduce achromatic images from a monochrome or black-and-white transmission, those who have had experience of viewing colour television would probably agree that slight patchiness of colour in an otherwise achromatic image is not quite as important as a habitual monochrome viewer might suppose. One receiver gave rise to 17 per cent of opinions in the 'objectionable' categories, but an average of many receivers gave 11 per cent on slides and a lower percentage on film.

Misregistration of the Primary-Colour Images

This can occur in receivers of the three-gun tube variety or in cameras employing three separate pick-up tubes. Trinoscope displays may also suffer since they have an image-registration problem of the same order.

Thus with film and slide transmissions originating from flying-spot scanners misregistration arises only in the receiver. It should therefore be identical on colour and monochrome transmissions provided a colour receiver is being used. In general, this was found to be the case, even when a considerable percentage in the 'objectionable' category was encountered. There were, however, some exceptions. In a final high-power test involving motion picture film and cameras, the 'objectionable' percentages shown in Table 5 were found.

TABLE 5

	FILM (per cent)	CAMERAS (per cent)
Colour transmission	3	17
Monochrome transmission	3	—

Misregistration is the least favourable feature of colour television as at present conceived, but it should be noted that it is not caused by the method of coding and decoding the colour signals. It will occur with any system in which the scene to be televised is analysed or synthesized into separate primary-colour images. Receiver and camera instability are the main causes of misregistration.

Cross-Colour

Unless a 'notch' filter is included in the luminance channel of a colour transmitter in order to eliminate luminance components in the neighbourhood of 2·66Mc/s (where the chrominance signal will be) there will be crosstalk or cross-colour interference due to these luminance components being synchronously detected by the chrominance demodulators in the receiver and their subsequent passage through the colour-difference channels and treatment as colour signals. For example, the 2½Mc/s bars on Test Card C will appear as a spurious colour signal having a frequency of 2·66-2·5 = 0·16Mc/s. The effect of the inclusion in the luminance channel, of a notch filter about 400kc/s wide and 20dB deep is to reduce the percentage of observers who report the phenomenon as being in one way or another 'objectionable' from a maximum of 9 to 0. The price paid for the virtual elimination of cross-colour is a loss of resolution at frequencies above 2·4Mc/s. This loss need not occur when black-and-white transmissions take place since the notch filter may be taken out of circuit for these. If the notch filter is intended to allow a useful contribution to resolution to be made by luminance components having frequencies well above that of the chrominance sub-carrier, that is, close to 3Mc/s, then it is necessary to ensure that a reversal of phase does not take place when a component below sub-carrier frequency is moved to a frequency above sub-carrier. A satisfactory notch filter is the one shown in Fig. 1.

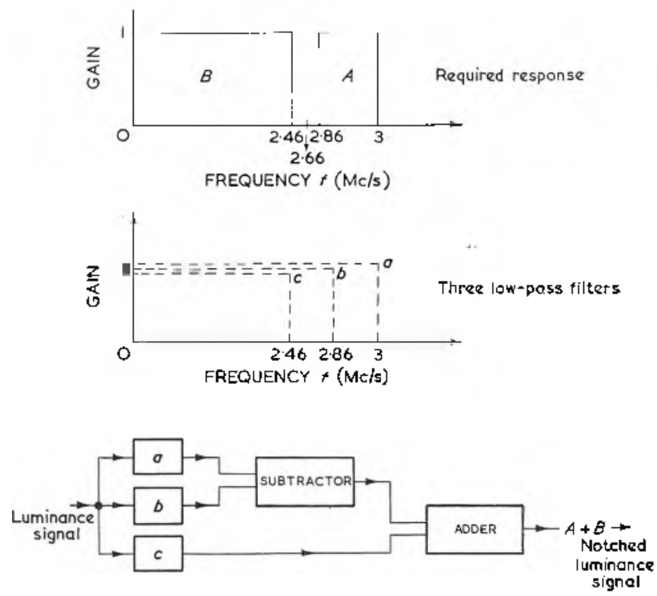


Fig. 1. Transmission luminance notch filter

(Output of a) − (Output of b) = A
Output of c = B
(Output of a) − (Output of b) + (Output of c) = A + B

Noise, Random Fluctuation and Impulsive

The eye is less sensitive to chromaticity flicker than it is to luminance flicker. A very beautiful experiment described in many textbooks on colour television used a random fluctuation noise generator to superimpose a noise signal upon a (red + green) = yellow raster. The green noise could be polarity reversed with respect to the red noise. It was found by a subjective test that viewers could put up with 8dB more noise when the two noise components (red and green) were in phase opposition and therefore contributing only to chromaticity. Thus the addition of colour to a monochrome luminance picture would be expected to add only 1/6·3 (= −8dB) more noise power—a negligible amount. This is further reduced, in theory, by the ratio of chrominance-to-luminance bandwidth, say 0·5Mc/s/3Mc/s = 1/6. Thus an overall increase of only 1/36 in noise power might be expected. Unfortunately, this is not obtained to the full extent because the eye sees low-frequency noise more clearly than high-frequency noise, thus a good deal of the luminance bandwidth is not contributing fully to noise visibility while all the synchronously-demodulated chrominance noise is, since the chrominance bandwidth is low. A further factor which increases the noise contribution from the chrominance channel is the existence of any aperture correction in the primary-colour signal channels at the picture source. Such aperture correction contributes a 'tipped-up' noise spectrum to the luminance signal. The chrominance channel at 2·66Mc/s finds itself situated in that part of the luminance channel where the noise spectrum would contribute little to visibility, but contributes considerably to noise cross-colour by synchronous demodulation into the colour-difference channels, thereby undergoing a frequency conversion from 2·66Mc/s down to very low frequencies. Once again, a transmission luminance notch filter is a great help, if the noise is contributed mainly by the picture source rather than by the signal-frequency circuits of the receiver.

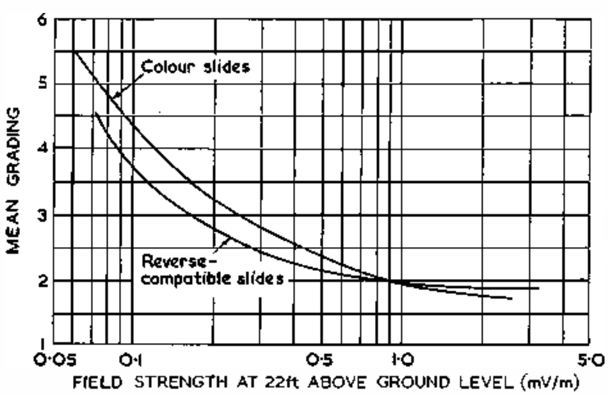


Fig. 2. Grading given to noise as a function of field strength

Subjective tests showed no change in noise visibility under practical conditions of reception on a colour receiver between colour or black-and-white transmissions. It should be remembered that during black-and-white transmissions the receiver chrominance circuits are inactivated by a 'colour killer' and so no cross-colour phenomenon can occur. This shows that the noise-power addition of about 1/36th due to the chrominance may not be far wrong, under normal circumstances. Some subjective tests done in a mobile laboratory showed that the 8dB advantage due to eye insensitivity to chromaticity flicker is reduced when the noise level is high enough. Fig. 2 shows the grading given to noise as a function of field strength for reception by a colour receiver of coloured scenes on the one hand and black-and-white scenes on the other. The 'colour noise' curve rises more steeply at low field strengths corresponding to high noise levels.

There is nothing to report with regard to impulsive interference except that close inspection of the individual spots of light on the colour tube reveals that they have a bright, small black-and-white centre with spread-out coloured leading and trailing edges due to the low bandwidth of the chrominance channel.

Continuous-Wave Interference

The ratio of wanted to unwanted signal strengths which will give a certain level of interference with monochrome reception is shown in Fig. 3.

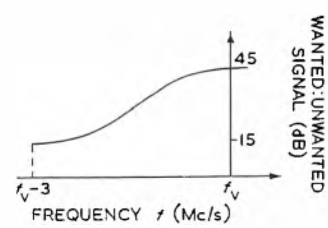
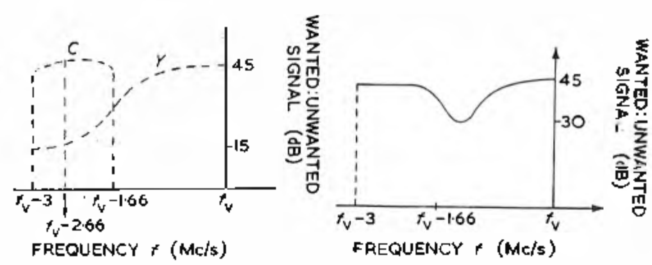


Fig. 3. C.W. interference: monochrome



Figs. 4 and 5. C.W. interference: colour

Now a colour receiver has two frequencies near to which it is highly sensitive to interference. The first is, as with monochrome, the vision (or luminance) carrier frequency; the second is the chrominance carrier frequency. If the chrominance carrier had the same amplitude as the vision carrier and if chromaticity flicker were just as visible as luminance flicker then the interference curve for colour, corresponding to Fig. 3 would be as shown in Fig. 4.

Now, it is known that about 8dB more interference can be accepted if it gives rise only to changes of chromaticity so one would expect the dotted curve marked C in Fig. 4 to have a maximum at about 45 − 8 = 37dB instead of at 45dB as shown. It must be noted, further, however, that the chrominance carrier is restricted to an amplitude which is never greater than about a quarter of that of the peak-white vision (luminance) carrier and in any given area of colour it cannot exceed about a half of the vision carrier. This fact approximately compensates for the advantage obtained from chromaticity flicker, with the result that the curve C and the over-all curve shown in Fig. 5 both have maxima which reach the 45dB figure at f_v and at (f_v − 3)Mc/s.

It is evident from Fig. 5 that co-channel interference with a colour transmission is more likely to occur than with a monochrome transmission.

Conclusions with Regard to Colour Reception

Misregistration, colour fidelity and noise were the most important undesirable features and they were all most prominent when colour cameras were used as the picture source. This is not discouraging, since it is reasonable to forecast improvements in colour cameras. Misregistration is a matter of selecting camera tubes and improving the camera unit particularly with regard to stability. Colour fidelity is largely a matter of lighting and operating technique while noise may be improved in a number of ways including the use of cameras with larger targets than the 3in image orthicons which are at present the only colour camera tubes in use.

The N.T.S.C. method of coding and decoding seems to leave little to be desired, although band sharing with a 50 field-per-second television system may be less effective than with a 60 field-per-second system.

Features of the N.T.S.C. System Likely to be Used in Future Colour Television Systems

Even if compatability will not be required it is most likely that a luminance signal will be used, since it does seem as if the eye appreciates an illuminated scene by taking note of luminance, hue and saturation.

Tests done partly with the aid of colour television equipment seem to show that hue demands a tolerance of the order of $1/36^{th}$ although, of course, this figure varies over the gamut of colours. Saturation is much more tolerant; a tolerance of about $1/5^{th}$ being sufficient. If primary-colour signals were transmitted then, since a change in any one of them of $1/36^{th}$ would give rise to a change of hue which would be of the same order, one would be forced to transmit three wide-band high-stability signals instead of one wide-band signal (luminance), one narrow-band high-stability signal (hue) and one narrow-band low stability signal (saturation). These considerations lead to the concept of a luminance and chromaticity signal in which amplitude carries saturation while phase carries hue. Such a signal package may differ in only two respects from that of the N.T.S.C. Band sharing need not be adopted, with the consequent advantage to the compatible viewer of the absence of dot-pattern interference and the disadvantage of having to put up with a non-panchromatic picture. True chromaticity might be sent instead of chrominance. This would mean so constructing the chromaticity signal that while its phase carried hue information, like the chrominance signal, its amplitude would carry true saturation unlike the chrominance signal wherein amplitude carries saturation multiplied by luminance.

One aspect of the constant-luminance principle may be stated simply as; 'If two signals are transmitted, one of which carries only luminance information, then interference with the other signal will make no contribution to luminance.' Another principle which might be called the constant-chromaticity principle might be stated as; 'If two signals are transmitted, one of which carries only chromaticity, then interference with the other signal will make no contribution to chromaticity.' It is just because the chrominance signal carries some luminance as well as chromacity that interference in the luminance channel contributes to a change of saturation. How much improvement would be obtained by transmitting in conformity with both constant-luminance and constant-chromaticity principles is not possible to say in the absence of subjective tests. It is not difficult to think of what might be a suitable type of chromaticity signal:

$$Ch = 1/Y \sqrt{[(R-Y)^2 + (B-Y)^2]} \sin [\omega_s t + \arctan (R-Y)/(B-Y)]$$

Acknowledgments

The author wishes to acknowledge the help of his colleagues in the BBC, upon whose work the information given in this article depends; in particular his erstwhile chief, Mr. G. G. Gouriet, Mr. A. V. Lord, Mr. C. B. B. Wood, Mr. W. N. Sproson, Mr. S. N. Watson, Mr. T. Worswick and many others who were directly concerned in the colour field trials. For help and encouragement, he is grateful to the Head of the Research Department, Mr. W. Proctor Wilson, C.B.E., and the Director of Engineering for kind permission to publish.

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A Tape-Programmed Drilling Machine

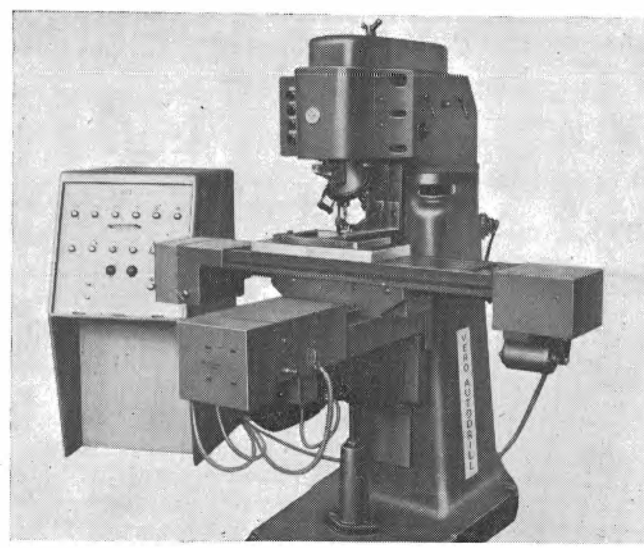
A tape-programmed drilling machine with purely electrical control has recently been developed by Vero Precision Engineering Ltd, Southampton, and will be marketed by the Catmur Machine Tool Corp. Ltd of 103 Lancaster Road, London W.11.

It is capable of drilling and tapping a pattern of up to 600 holes of 6 different sizes in any order and this is achieved by controlling both the work position and the 6 spindle turret head entirely automatically from a punched tape.

The control system, which is manufactured by Airmec Limited of High Wycombe, operates by measuring very accurately the total angular rotation of the leadscrews, a function performed by a positioner unit consisting of a series of commutator-type switches. Manual operation, if required, is performed from a control panel, and automatic drilling is controlled by a plastic tape carrying information for up to 600 complete drilling or tapping operations.

The drilling machine itself is fitted with a 32in by 9in co-ordinate table having a longitudinal traverse of 24in and a cross traverse of 12in. The positioning speed is 60in per minute and an overall drilling accuracy of $\pm 0{\cdot}003$in is guaranteed. The turret head is fitted with 6 chucks for accommodating drills of up to 5/16in diameter, and spindle speeds from 300 to 6 000rev/min are provided. The maximum distance from the spindle to the table is 18in.

The drilling machine and its control equipment



ANALOGUE COMPUTER TECHNIQUES

(Part 1)

The Solution of Simultaneous Equations, Eigenvalue Problems and Polynomial Equations

By F. C. Harbert*, B.A.

The first part of this three-part article on analogue computer techniques describes the computer arrangements required to solve simultaneous algebraic equations, eigenvalue problems and polynomial equations using an iterative method.

The second part will describe a method for obtaining approximate solutions of partial differential equations by finite difference techniques.

The last part will present different computer arrangements required to simulate transportation delay using the Padé approximations, illustrated by an example showing the effect of such a delay on the stability of a closed loop system and a technique for counteracting its destabilizing effect.

(Voir page 126 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 131)

A LARGE number of engineering problems require the solving of a set of simultaneous linear algebraic equations of the form:

$$\begin{array}{l} a_{11}x_1 + a_{12}x_2 + a_{13}x_3 + \ldots\ldots + a_{1n}x_n - c_1 = 0 \\ a_{21}x_1 + a_{22}x_2 + a_{23}x_3 + \ldots\ldots + a_{2n}x_n - c_2 = 0 \\ \quad ,, \qquad ,, \qquad ,, \qquad ,, \quad ,, \quad ,, \\ \quad ,, \qquad ,, \qquad ,, \qquad ,, \quad ,, \quad ,, \\ \quad ,, \qquad ,, \qquad ,, \qquad ,, \quad ,, \quad ,, \\ a_{n1}x_1 + a_{n2}x_2 + a_{n3}x_3 + \ldots\ldots + a_{nn}x_n - c_n = 0 \end{array} \qquad \ldots\ldots (1)$$

where a_{ij} and c_i are real constants and the x_i are real variables.

This system of equations may be prepared for solution on an analogue computer by one of two methods. Each method results in replacing the linear algebraic system by a linear differential system whose steady-state solution exists and is the solution of the linear algebraic system. The first method is the simpler and requires a minimum of equipment but can be relied upon to yield stable and convergent computer solutions only if the matrix of coefficients of the algebraic system is positive definite. The second method uses twice as much computer equipment but can be relied upon to yield stable and convergent computer solutions regardless of the arrangement of the linear algebraic system.

There are classes of problems having the form of equations (1) which yield stable solutions on the analogue computer without requiring conversion to the differential equation form. These may therefore be solved to advantage by programming directly from equations (1). However, there is considerably less assurance that the direct method will prove to be stable and it is therefore more satisfactory to employ the differential equation form.

The system in equations (1) is said to be positive if it is real and if the quadratic form:

$$\begin{array}{l} Q = \tfrac{1}{2}\,(a_{11}x_1^2 + a_{12}x_1x_2 + \ldots\ldots\ldots\ldots + a_{1n}x_1x_n \\ \qquad + a_{21}x_2x_1 + a_{22}x_2^2 + \ldots\ldots\ldots\ldots + a_{2n}x_2x_n \\ \qquad\quad ,, \qquad\qquad ,, \qquad\qquad\qquad ,, \\ \qquad\quad ,, \qquad\qquad ,, \qquad\qquad\qquad ,, \\ \quad a_{n1}x_nx_1 + a_{n2}x_nx_2 + \ldots\ldots\ldots\ldots + a_{nn}x_n^2) \end{array} \qquad \ldots\ldots\ldots\ldots (2)$$

constructed from the coefficients and solutions of equations (1) is non-negative for all possible combinations of the real x_i. The system is said to be positive definite if Q is non-negative and moreover is only zero when every x_i is zero.

* *Louis Newmark Ltd.*

Many important physical systems lead to a system of equations having a matrix of coefficients which is positive definite, these include elastic systems, d.c. networks and many architectural structures. The criterion is not satisfied in general for a.c. networks.

If the system is thought to be positive definite, equations

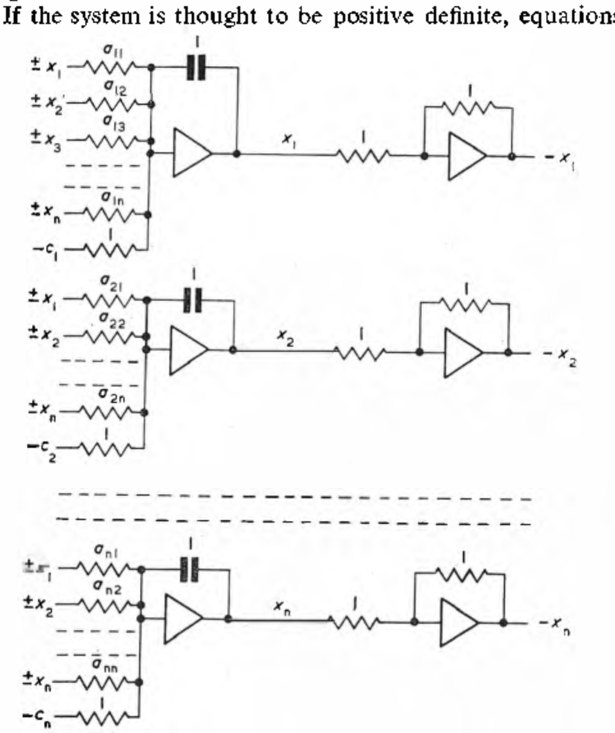


Fig. 1. Computer diagram for solution of positive definite algebraic system

(1) may be prepared for analogue computer solution by rewriting in the following form:

$$\begin{array}{l} a_{11}x_1 + a_{12}x_2 + \ldots\ldots + a_{1n}x_n - c_1 = -dx_1/dt \\ a_{21}x_1 + a_{22}x_2 + \ldots\ldots + a_{2n}x_n - c_2 = -dx_2/dt \\ \quad ,, \qquad ,, \quad \ldots\ldots \quad ,, \quad ,, \qquad ,, \\ \quad ,, \qquad ,, \quad \ldots\ldots \quad ,, \quad ,, \qquad ,, \\ \quad ,, \qquad ,, \quad \ldots\ldots \quad ,, \quad ,, \qquad ,, \\ a_{n1}x_1 + a_{n2}x_2 + \ldots\ldots + a_{nn}x_n - c_n = -dx_n/dt \end{array} \qquad \ldots\ldots\ldots\ldots (3)$$

This system of equations may be set up directly on to the computer as shown in Fig. 1. If the computer solution is stable the steady-state values of the variables will clearly be the desired solutions to the given algebraic system, since then $dx_i/dt = 0$.

The analytical test for positive definiteness of the algebraic system of equations (1) is difficult to apply and requires as much labour as solving the system. The practical test for positive definiteness would in fact be to place the transformed system of equations (3) on to the computer and to find whether steady-state solutions are delivered. Obviously much time and effort may be wasted if it is not known whether the algebraic system of equations (1) is positive definite.

If there is any doubt that the system of equations (1) is positive definite, the second method of computer solution is recommended. It can be shown that the algebraic system

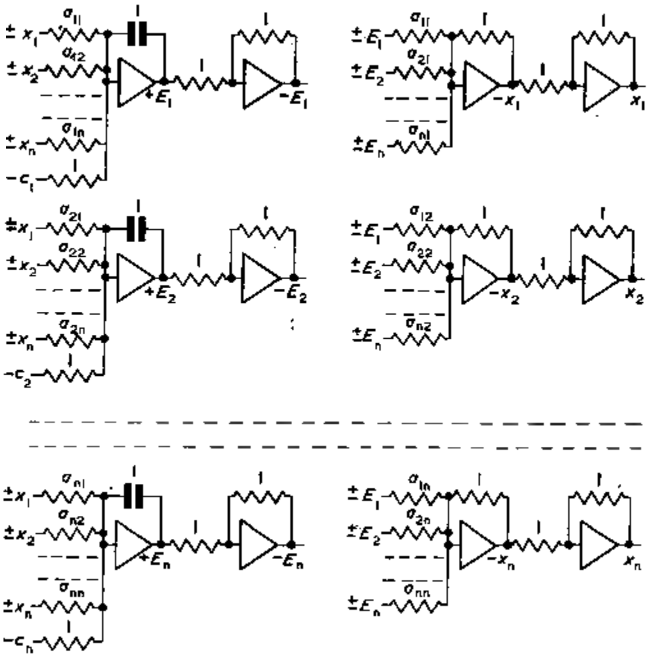


Fig. 2. Computer diagram for solution of general algebraic system

of equations (1) may be converted into the following form which will always yield stable solutions on an analogue computer.

$$\begin{array}{l} a_{11}x_1 + a_{12}x_2 + \ldots\ldots\ldots + a_{1n}x_n - c_1 = -dE_1/dt \\ a_{21}x_1 + a_{22}x_2 + \ldots\ldots\ldots + a_{2n}x_n - c_2 = -dE_2/dt \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ a_{n1}x_1 + a_{n2}x_2 + \ldots\ldots\ldots + a_{nn}x_n - c_n = -dE_n/dt \end{array} \qquad \ldots\ldots\ldots (4)$$

where E_i is defined by the equations:

$$\begin{array}{l} x_1 = a_{11}E_1 + a_{21}E_2 + \ldots\ldots\ldots\ldots + a_{n1}E_n \\ x_2 = a_{12}E_1 + a_{22}E_2 + \ldots\ldots\ldots\ldots + a_{n2}E_n \\ ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \\ x_n = a_{1n}E_1 + a_{2n}E_2 + \ldots\ldots\ldots\ldots + a_{nn}E_n \end{array} \qquad \ldots\ldots\ldots\ldots (5)$$

These equations may be set up on a computer as in Fig. 2.

From Fig. 1 it can be seen that the computer solution for equations (1) requires $2n$ amplifiers. By contrast the second method requires $4n$ amplifiers. However, the assurance of stable solutions gives the second approach strong advantages.

Accuracy

The accuracy of computer solutions is a direct function of the quality of the computing equipment and the accuracy to which the potentiometers may be adjusted. It is difficult to obtain accurate solutions to ill-conditioned systems of equations by any method. The accuracy of computer solutions obtained by either of the methods outlined above may be improved to any desired extent by employing a formal iterative process.

Assume that there is a set of approximate solutions to the system of equations (1) which will be denoted by $x_1', x_2', x_3', \ldots . x_n'$. As approximate solutions, these values differ from the exact solutions by small quantities which will be denoted as $\Delta x_1, \Delta x_2 \ldots\ldots\ldots \Delta x_n$

and therefore:

$$x_i' = x_i + \Delta x_i \ldots\ldots \qquad \ldots\ldots\ldots\ldots (6)$$

and

$$\begin{array}{l} a_{11}x_1' + a_{12}x_2' + \ldots\ldots + a_{1n}x_n' - c_1 = r_1 \\ a_{21}x_1' + a_{22}x_2' + \ldots\ldots + a_{2n}x_n' - c_2 = r_2 \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ a_{n1}x_1' + a_{n2}x_2' + \ldots\ldots + a_{nn}x_n' - c_n = r_n \end{array} \qquad \ldots\ldots\ldots\ldots (7)$$

where r_i are called residues

Combining equations (6) and (7)

$$\begin{array}{l} a_{11}(x_1 + \Delta x_1) + a_{12}(x_2 + \Delta x_2) + \ldots + a_{1n}(x_n + \Delta x_n) - c_1 = r_1 \\ a_{21}(x_1 + \Delta x_1) + a_{22}(x_2 + \Delta x_2) + \ldots + a_{2n}(x_n + \Delta x_n) - c_2 = r_2 \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \quad ,, \\ a_{n1}(x_1 + \Delta x_1) + a_{n2}(x_2 + \Delta x_2) + \ldots + a_{nn}(x_n + \Delta x_n) - c_n = r_n \end{array} \qquad \ldots\ldots\ldots\ldots (8)$$

finally subtracting corresponding members of equations (1) and (8):

$$\begin{array}{l} a_{11}\Delta x_1 + a_{12}\Delta x_2 + \ldots\ldots + a_{1n}\Delta x_n - r_1 = 0 \\ a_{21}\Delta x_1 + a_{22}\Delta x_2 + \ldots\ldots + a_{2n}\Delta x_n - r_2 = 0 \\ ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \\ ,, \quad ,, \quad ,, \quad ,, \\ a_{n1}\Delta x_1 + a_{n2}\Delta x_2 + \ldots\ldots + a_{nn}\Delta x_n - r_n = 0 \end{array} \qquad \ldots\ldots\ldots\ldots (9)$$

Equations (9) have exactly the same form and characteristic determinant as the original system of equations (1). As a result precisely the same computer arrangement employed to find the approximate solutions x_i' may be used to solve for the first order corrections Δx_i provided only that the c_i in equations (1) are replaced by r_i in the solution of equations (9). The values of the residues r_i may be easily determined arithmetically.

If the residues are already small, equations (9) may be multiplied by 10 or 100 for example and the computer solution will yield $10\Delta x_i$ or $100\Delta x_i$. Continuing the iteration in the same manner second order corrections may be found, etc., etc., until solutions are carried out to the required degree of accuracy. Even if the coefficients a_{ij} in equations (1) were not set accurately in the computer arrangement the solutions can be obtained by iteration to any required degree of accuracy provided the correct values of the coefficients a_{ij} are used when determining the residues.

It has been assumed so far that equations (1) contain only real coefficients a_{ij}. The methods are, however, applicable to equations where the coefficients a_{ij} are complex quantities. In such cases equations (1) may be separated into their real and imaginary parts whereupon each group of equations may be solved by the methods already outlined. In the general case where all coefficients have the form $a_{jk} + ib_{jk}$ and none of the a_{jk} or b_{jk} terms are zero, twice as many computing amplifiers are required as for the general case with real coefficients.

Eigenvalue Problems

Most eigenvalue problems may be expressed in terms of a system of linear algebraic equations having the general form:

$$\begin{array}{l}
(a_{11}-b_{11}\lambda)\,x_1+(a_{12}-b_{12}\lambda)x_2+\ldots\ldots+(a_{1n}-b_{1n}\lambda)x_n=0\\
(a_{21}-b_{21}\lambda)\,x_1+(a_{22}-b_{22}\lambda)x_2+\ldots\ldots+(a_{2n}-b_{2n}\lambda)x_n=0\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,,\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,,\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,,\\
(a_{n1}-b_{n1}\lambda)\,x_1+(a_{n2}-b_{n2}\lambda)x_2+\ldots\ldots+(a_{nn}-b_{nn}\lambda)x_n=0
\end{array}\qquad\ldots\ldots\ldots\ldots\ldots\ldots\ (10)$$

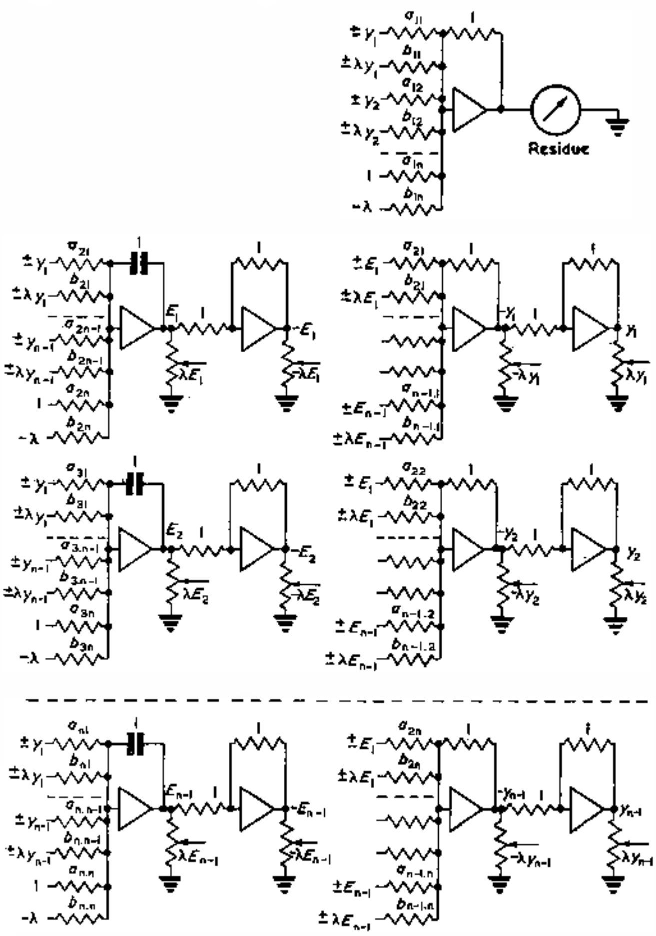


Fig. 3. Computer arrangement for solution of eigenvalue problems

where all the coefficients and variables are real. An eigenvalue is a value of the dimensionless parameter λ for which there are non-zero values of the x_i which satisfy equations (10). These eigenvalues are the n values of λ which make the characteristic determinant of equations (10) vanish. Practical interpretations of the eigenvalues would be natural frequencies, normal modes, buckling loads, critical speeds, critical loads, etc., according to the problem.

Since the equations (10) are homogeneous one may divide each equation by x_n (assuming x_n is not zero) to form n new equations with $(n-1)$ unknowns $y_1(=x_1/x_n)$, $y_2(=x_2/x_n)$, $\ldots y_{n-1}(=x_{n-1}/x_n)$ considered as the new variables. The terms $(a_{1n}-b_{1n}\lambda)$ up to $(a_{nn}-b_{nn}\lambda)$ formerly the coefficients of x_n in equations (10) will now be considered the constant terms in the new group of n equations.

$$\begin{array}{l}
(a_{11}-b_{11}\lambda)y_1+(a_{12}-b_{12}\lambda)y_2+\ldots\ldots+(a_{1,n-1}-b_{1,n-1}\lambda)y_{n-1}+(a_{1n}-b_{1n}\lambda)=0\\
(a_{21}-b_{21}\lambda)y_1+(a_{22}-b_{22}\lambda)y_2+\ldots\ldots+(a_{2,n-1}-b_{2,n-1}\lambda)y_{n-1}+(a_{2n}-b_{2n}\lambda)=0\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,, \qquad\qquad ,,\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,, \qquad\qquad ,,\\
\quad ,, \qquad\qquad ,, \qquad\qquad ,, \qquad\qquad ,,\\
(a_{n1}-b_{n1}\lambda)y_1+(a_{n2}-b_{n2}\lambda)y_2+\ldots\ldots+(a_{n,n-1}-b_{n,n-1}\lambda)y_{n-1}+(a_{nn}-b_{nn}\lambda)=0
\end{array}\qquad\ldots\ldots\ldots\ldots\ (11)$$

From these n equations one takes $(n-1)$ and solves for the $(n-1)$ variables $y_1, y_2, \ldots y_{n-1}$, making use of either general method of analogue computer solution already outlined. In this process the computer delivers solutions $y_1, y_2, \ldots y_{n-1}$ which depend upon the particular value of λ (which must of course be introduced consistently in all

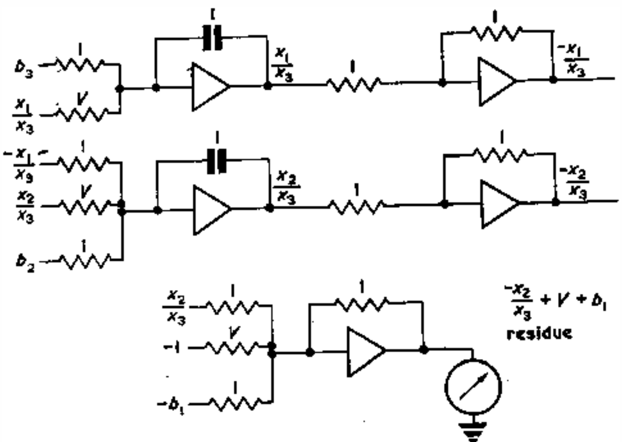


Fig. 4. Solution of cubic $V^3+b_1V^2+b_2V+b_3=0$ by residue method

$n-1$ equations). If λ is altered continuously, the family of solutions delivered by the computer will likewise change continuously.

The single equation from (11) which was not employed in solving for the $n-1$ variables may be used to compute a continuous residue, since values of $y_1, y_2, \ldots y_{n-1}$ and the corresponding value of λ are available from the first process. As λ is changed slowly and continuously the residue from the single equation will be found to vanish and remain at zero for particular values of λ. These are the eigenvalues which are sought, and the corresponding values of $y_1, y_2, \ldots y_{n-1}$ are the ratios of the original variables x_1 which are appropriate to the particular eigenvalue for the system.

The computer set-up is shown in Fig. 3 (based upon the second general method described above for solving linear algebraic equations). It is assumed that the value of λ is introduced consistently by the use of mechanically ganged potentiometers. It is also assumed that the last $(n-1)$ equations from (11) are used to obtain automatically the solutions of the $(n-1)$ variables as functions of λ. The first of equations (11) is then used to generate the residue.

Iterative methods as described above are applicable for improving the accuracy of the computer solutions for the eigenvalues and the corresponding ratios of the variables. Extension of the outline given above also permits analogue computer solution of problems with complex eigenvalues.

The Solution of Polynomial Equations

First Method

An equation of the n^{th} degree of the form.

$$V^n+b_1V^{n-1}+\ldots b_n=0 \qquad \ldots\ldots\ (12)$$

can be written in the form of the characteristic determinant of equations (10) set equal to zero. It is then equivalent in form to the equation of an eigenvalue problem and may be solved using the same techniques.

As an example, take the cubic form

$$V^3 + b_1V^2 + b_2V + b_3 = 0 \quad \ldots\ldots\ldots \quad (13)$$

This equation can be written as

$$\begin{vmatrix} V & 0 & b_3 \\ -1 & V & b_2 \\ 0 & -1 & V+b_1 \end{vmatrix} = 0 \quad \ldots\ldots\ldots\ldots \quad (14)$$

Equation (14) can be considered to represent the linear algebraic system

$$\begin{aligned} Vx_1 + 0 + b_3x_3 &= 0 \\ -x_1 + Vx_2 + b_2x_3 &= 0 \\ -x_2 + (V + b_1)\,x_3 &= 0 \end{aligned} \quad \ldots\ldots\ldots\ldots \quad (15)$$

which is clearly in the same form as the basic eigenvalue problem of equations (10). The same techniques of analogue computer solution can therefore be applied. The eigenvalues which satisfy equations (15) are the roots of the cubic. The computer set-up is shown in Fig. (4). The first two equations are solved for x_1/x_3 and x_2/x_3; the third equation is used to compute the residue.

Second Method

The main disadvantage of the previous method is that it requires ganged potentiometers which are manually operated. In this second method time is the variable V which is integrated to obtain its powers. Fig. 5 illustrates the computer arrangement required to obtain the powers of V. From these the residue is computed. The time elapsed from the start of the compute sequence to the time when the residue is zero gives a positive real solution to the equation. Multiplying the equation by -1 makes it possible to obtain the negative real values of V which satisfy the cubic.

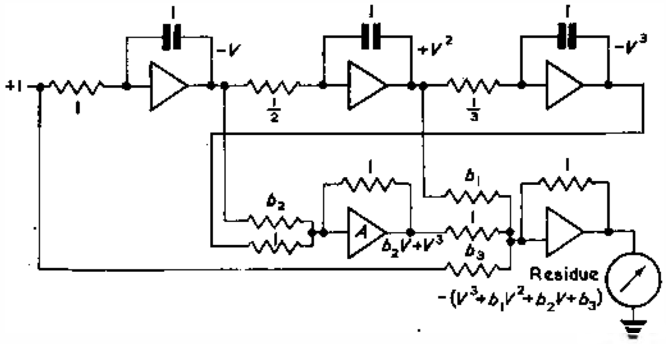


Fig. 5. Solution of cubic by means of the time integral method

(By omitting amplifier 'A', negative roots may be investigated)

(To be continued)

A Transistor Curve Tracer

A comprehensive transistor curve tracer has been developed by De Havilland Propellers Ltd with test facilities providing the following facilities:

(1) *Automatic Test*

Up to 80 semiconductors can be accommodated and these are divided into four banks of 20, each bank being connected to a uniselector. Each semiconductor is, in turn, subjected to the specified test conditions by the appropriate uniselector and the required parameter is displayed on a cathode-ray tube. The numbers of the bank and semiconductor under test are shown by lights on the uniselector control unit. Any one, or all, of the test banks may be selected for test as required, and the banks chosen will run through in succession.

(2) *Test Quantities*

Provision has been made to test either one semiconductor at a time, or up to 80 automatically. A 'single' or 'auto' switch is provided on the supply and control unit.

(3) *pnp and npn Transistors*

The supply and control unit contains a pnp/npn switch for correct voltage polarity.

(4) *Diode Characteristics*

Forward and reverse diode characteristics may be displayed.

(5) *Input and Output Characteristics*

Base currents are supplied by three constant current generators, each of which is set and monitored individually. Base current switching is either manual or electronic. Electronic switching provides a family of seven curves.

When the specified currents and voltages have been set, and the input or output characteristics have been selected by means of the switch on the supply and control unit, the monitor meter switch is put to 'auto' and the appropriate family of curves is displayed on the tube.

(6) *Test Capabilities*

The curve tracer is designed to test signal, medium power and power transistors, in any configuration. Sweep voltage may be applied as forward or reverse half wave, and full wave.

(7) *Power Limiting*

To avoid excess dissipation when displaying the output characteristics, a variable power limiter is provided. The maximum dissipation is then controlled by the limiter switch setting.

(8) *Camera Unit*

In addition to the monitor tube already mentioned, a further display unit to which a camera is attached is incorporated. The camera drive is part of the automatic system, and the camera exposure time is set by means of a switch.

When a trace is displayed a photograph is taken and, as the next semiconductor is selected, the camera drive moves the film along ready to photograph this trace.

The tube has a calibrated graticule to facilitate subsequent analysis of the photographs which contain the following information:

(a) The selected Input or Output family of curves.
(b) A graticule with current and voltage axes.
(c) The full-scale values of current and voltage as selected on the sensitivity controls.

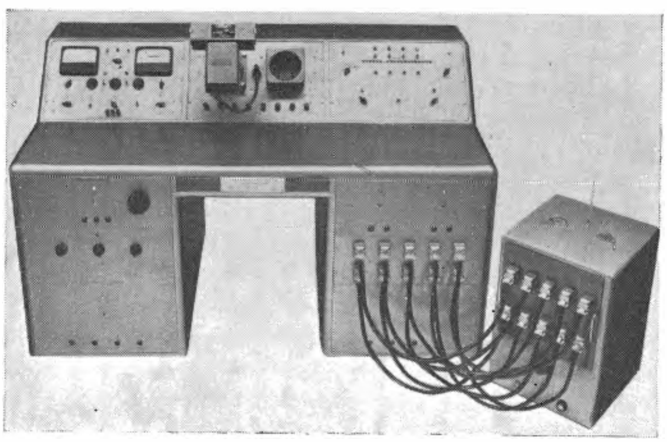
(9) *Use of Unskilled Operator*

Once the curve tracer has been set up to the specified test conditions, the operator merely presses the 'start' button on the uniselector control unit. The tracer will then automatically test and photograph however many banks of semiconductors have been switched in. During this process the operator watches the monitor tube and notes down any faulty traces using the numbers shown by lights on the uniselector control unit. Any suspect traces may then be examined in greater detail from the photographs by a more skilled operator.

(10) *Measurement of I_{co}*

A socket is provided for direct measurement of the transistor leakage current by the use of a suitable meter.

The de Havilland Transistor Curve Tracer Unit. The transistors under test are mounted in the separate container seen on the right of the main console. This container can take the form of an oven, a refrigerator, or a humidity cabinet.



The Characteristics and Applications of Zener (Voltage Reference) Diodes

By J. A. Chandler*, B.Sc. (Eng.)

The detailed characteristics of Zener diodes are presented and curves are given showing the variation of a.c. and d.c. slope resistance with both current and voltage, breakdown voltage variation with temperature, and the forward characteristics of the diodes.

The general circuit of the simple stabilized supply is considered, with the limitations due to the diode and circuit limits. Temperature compensation of the diodes is considered and results shown. The stability of the output voltage with input and load variations is given in detail using the characteristics already mentioned. The two stage stabilizer is considered and a practical example of a high stability reference is given. The results show that a very high stability reference can be constructed using relatively simple circuits.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

THE Zener diode is a silicon junction diode possessing a reproducible breakdown in the inverse direction. 'Breakdown' implies that as the inverse voltage is increased, the current remains small until a certain voltage (known as the breakdown voltage), is reached, at which point the current increases rapidly with very little increase in voltage.

The breakdown is a function of the diode material and construction, and breakdown voltages can be produced from about one volt up to several hundred volts, and with various current and power ratings depending on the junction area and method of cooling. This article is concerned with the behaviour of diodes having breakdown voltages from 2 to 11V, a junction area of 5×10^{-4}in^2 and a power rating of approximately 2W in free air at 25°C ambient. It is proposed, first to describe the complete characteristics of these diodes and then relate these characteristics to application as a simple stabilizer, voltage reference, and finally as a precision voltage reference.

The simple methods available to obtain stabilized supplies of the voltage range considered have been very limited in the past, with various types of primary and secondary cells being the only possibilities[1]. The Weston unsaturated standard cell has been the only practical reference source, and this has to be used purely as a voltage reference with zero current drain. The stability obtainable from the Zener diode enables it to be used with advantage as a stabilizer and reference source capable of supplying constant current loads to a high order of accuracy.

The diode characteristics presented are in the main typical curves, and represent the average values obtained from many units, in some cases up to several hundred units. The general application data is calculated using these typical characteristics, but in cases where typical is not mentioned, measured values on individual units are being quoted.

Characteristics

General

Zener diodes in the 2 to 11V range possess the typical characteristics shown in Fig. 1. The voltage across the diode is plotted against the direct current through the diode, and as the breakdown voltage is reached the current can be seen to increase rapidly. These curves are of limited use in the design of voltage stabilized supplies, since the information they contain is not presented in a form readily suited to mathematical treatment, and they do not show clearly the differences between diodes of different voltages. In order to present further information and to enable calculations to be made, other characteristics are required.

A.C. Slope Resistance

One useful method of describing these curves is by their slope, which is of course a resistance. If this slope resistance is measured by a small a.c. signal it is not always found to be the same as the slope of the curves of Fig. 1 which are taken with d.c. values. This is due to thermal effects, since the diodes possess a voltage variation with junction temperature. If the current is changed, the power is changed and hence the temperature of the junction alters, resulting in a slight change of voltage. With a small a.c. signal the temperature variation is negligible and the true diode resistance is measured. These two effects are illustrated in Fig. 2 where a small part of the characteristics of a diode with a positive temperature coefficient is shown. The full lines indicate the d.c. characteristics

Symbols Used

I	=	Steady current
$I_{D(max)}$	=	Maximum rated diode current
$I_{D(min)}$	=	Minimum rated diode current
i	=	Small change in current
k	=	Per unit change of input voltage
R_{s1}	=	A.C. slope resistance of diode
R_{s2}	=	Temperature resistance of diode
R_D	=	D.C. slope resistance of diode
S	=	Stability ratio
ΔT_A	=	Change in ambient temperature
V	=	Steady voltage
V_T	=	Voltage due to temperature change
v	=	Small change in voltage
α	=	Temperature coefficient
θ	=	Diode thermal resistance

Subscripts

D = diode (current, voltage etc.)
1 = Input, 1st stage
2 = 2nd stage
L = Load
O = Output

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for three different ambient temperatures, and the dotted line indicates the a.c. characteristic starting from 20mA at 20°C. If the current is increased to 21mA, the point B is reached, and if this current is sustained the voltage rises to point C, which is the d.c. condition.

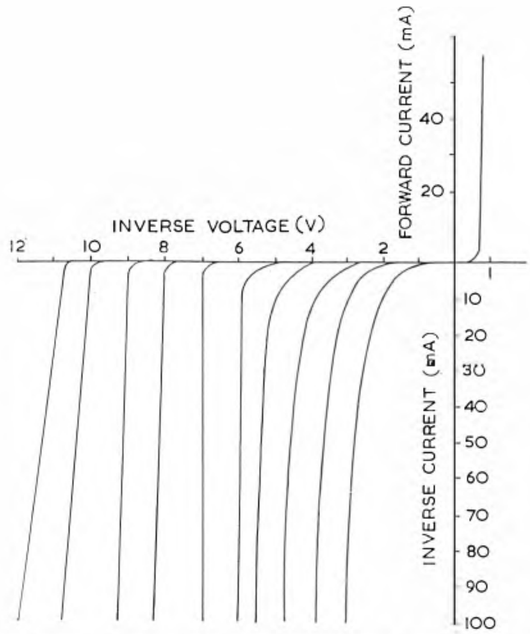

Fig. 1. Typical characteristics of silicon voltage reference diodes (d.c.)

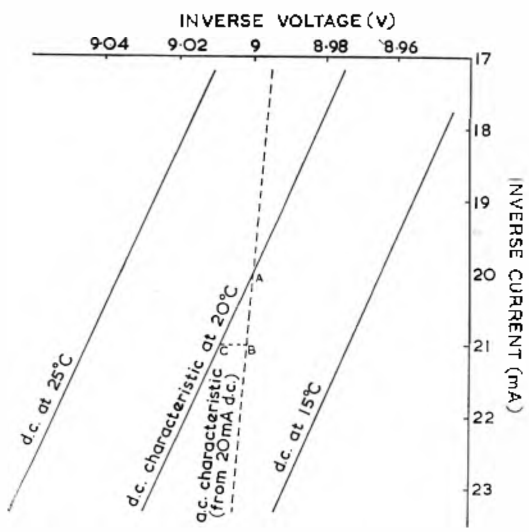

Fig. 2. A.C./D.C. characteristics

The a.c. slope resistance is defined as 'The resistance of the diode, operating at a given value of direct current, to an alternating current which is small in comparison with the direct current'. (This avoids first order non-linearities).

The importance of distinguishing between the a.c. and d.c. slope resistance is considered further in a section on the diode equivalent circuit.

By measuring large numbers of diodes, the a.c. slope resistance, at a given inverse current, may be plotted against breakdown voltage. For this purpose the breakdown voltage is defined as the voltage at an inverse direct current of 20mA, this value of current being arrived at by consideration of the slope resistance, consistency between units, and the reduction of noise pulses in higher voltage units.

Fig. 3 shows the typical a.c. slope resistance plotted against the breakdown voltage for inverse currents of 10, 20 and 50mA. It can be seen that these curves are of an unusual nature, having a minimum value at about 7V, rising to a maximum at about 3V, and also rising above 7V. From experiments performed on units with breakdown voltages up to 170V it has been confirmed that the

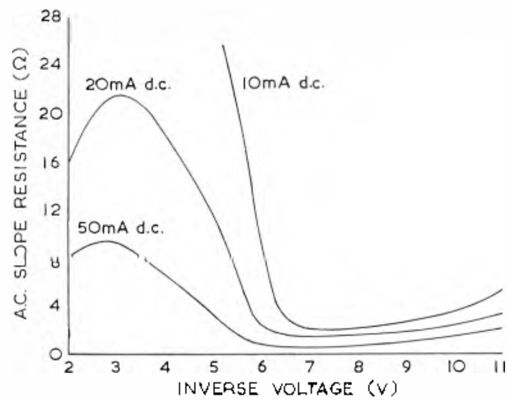

Fig. 3. Typical curves of a.c. slope resistance against inverse voltage

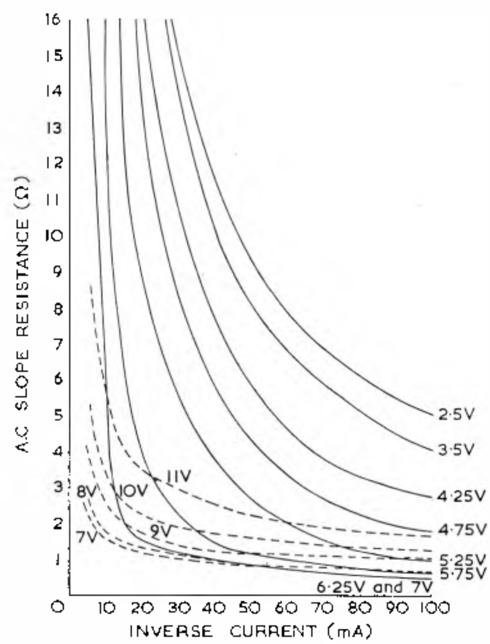

Fig. 4. Typical curves of a.c. slope resistance against inverse current

rise is continued to high values, e.g. 300Ω for a 100V diode (at 20mA d.c.).

Variation of A.C. Slope Resistance

From Fig. 3 it can be seen that the a.c. slope resistance decreases as the direct current increases. The typical variation with current for the various voltage diodes is shown in Fig. 4.

For the lower voltage diodes up to about 6V, the slope resistance is almost inversely proportional to the direct current, as shown by the curves being of hyperbolic form. Above 6V, the slope resistance decreases rapidly with current, reaching an almost constant value above 20mA.

Temperature Coefficient

The other inverse characteristic of great importance is the variation in breakdown voltage with temperature. This is difficult to depict on a single curve, since it varies with diode voltage, diode current, temperature and a.c. slope resistance.

Fig. 5 shows the change of voltage with temperature at particular currents, of 4, 5, 6 and 7V diodes. For the 7V diode the temperature change is linear with temperature and almost independent of current. For the 4, 5 and 6V diodes the change of voltage with temperature depends upon both the temperature and the operating current.

The non-linearity with temperature change occurs in the range[2] $4\frac{1}{2}$ to 6V.

The change in temperature coefficient with a change in operating current is due to the fact that the temperature coefficient depends upon the voltage across the diode, so that as the voltage changes with current so does the temperature coefficient. For the low slope resistance diodes in the 7V region the effect is hardly noticeable, but is prominent in the region below 6V. The nearest approach to a general curve is shown in Fig. 6, which shows the temperature coefficient for diodes from 2 to 10V operating at 25°C in free air, for 10, 20, 50 and 100mA. In the non-linear region of $4\frac{1}{2}$ to 6V it can only be taken as applying to about a 50°C rise, but above and below this it applies up to 200°C. This curve shows that there is a maximum negative value in the same region of voltage as the maximum which occurs in the slope resistance curves, and that the curves cross the zero axis in the middle of the non-linear region, so that in practice a zero temperature coefficient cannot be achieved over any reasonable range of current and temperature. (Garside and Harvey[2] discuss the temperature behaviour in detail).

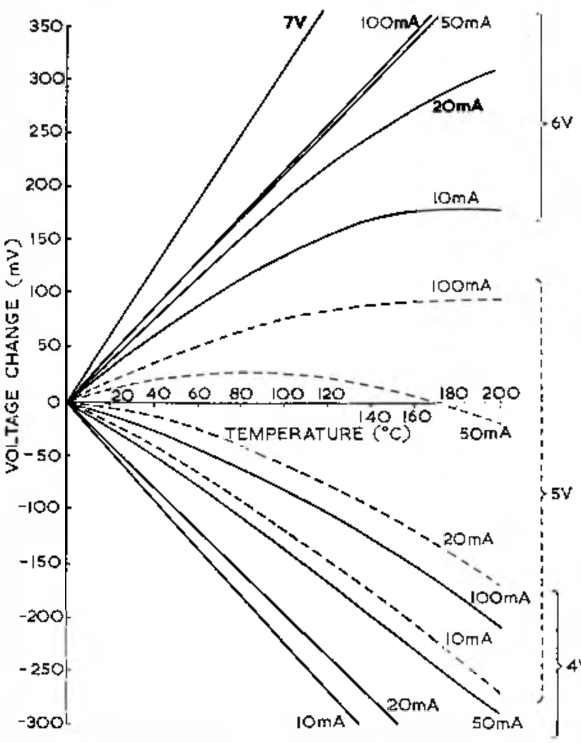

Fig. 5. Inverse voltage variation with temperature

Diode Equivalent Circuit

The diode equivalent circuit can now be represented as a battery, in series with the a.c. slope resistance and a voltage generator representing the temperature effect.

The temperature effect itself is in two parts, the first due to ambient temperature changes, and the second due to power changes in the diode.

If α_D is the temperature coefficient of voltage, and ΔT_A is the change in ambient temperature, the first effect is then given by $V_T = \alpha_D \cdot \Delta T_A$.

In order to evaluate the effect of power changes, the thermal resistance, θ, for the diode must be known. This is the rise in temperature per unit power dissipation (°C/W). If the voltage across the diode is V_D, and the change in current is i_D, the change in temperature is $\theta \cdot V_D \cdot i_D$.

The total voltage change is then:

$$\alpha_D (\Delta T_A + \theta \cdot V_D \cdot i_D) \qquad (1)$$

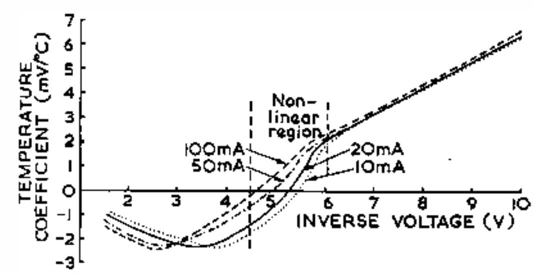

Fig. 6. Temperature coefficient against inverse voltage

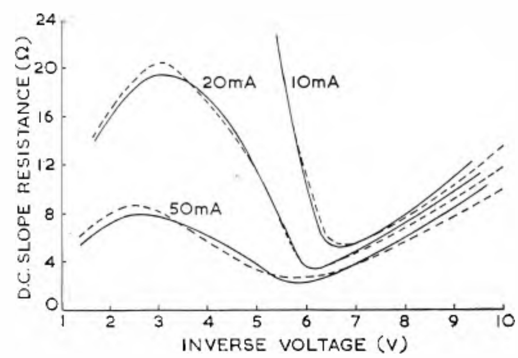

Fig. 7. Typical curves of d.c. slope resistance against inverse voltage
—— *experimental* - - - *calculated*

The second of these terms, $\alpha_D \cdot \theta \cdot V_D \cdot i_D$ can be regarded as a resistance term (even though it is, of course, negative for diodes below 5V) but it depends on the value of θ, the thermal resistance. The thermal resistance can be altered considerably by changing the mounting arrangements of the diode. For example, a normally wire ended unit with $\theta = 120$°C/W can have this changed to $\theta =$ 40°C/W by soldering on to a small copper fin. Provided that θ is to be kept constant, then, this term can be added to the a.c. slope resistance.

D.C. Slope Resistance

It follows from the previous section that the d.c. slope resistance R_D, is the sum of the a.c. slope resistance R_{S1}, and the temperature resistance R_{S2}.

Thus:

$$R_D = R_{S1} + R_{S2} \qquad (2)$$

This requires a new graph to show the slope resistance R_D against breakdown voltage, and is shown in Fig. 7. The dotted line represents the theoretical slope resistance obtained by calculations based on measurements of

R_{S1}, θ, and α_D. The full line shows the mean of experimental measurements made with a precision potentiometer. It should be noted that the experimental figures show a fair degree of scatter about the typical curve, and the agreement can be regarded as very close.

Variation of D.C. Slope Resistance

The change of d.c. slope resistance with current is shown in Fig. 8, and the difference between the a.c. and d.c. resistance is made obvious by comparing with Fig. 4. A noticeable feature is that at higher currents the low voltage diodes have lower slope resistances than the higher voltage units.

Forward Characteristics

The forward characteristics of a Zener diode are similar in general terms to those of high voltage rectifiers, but

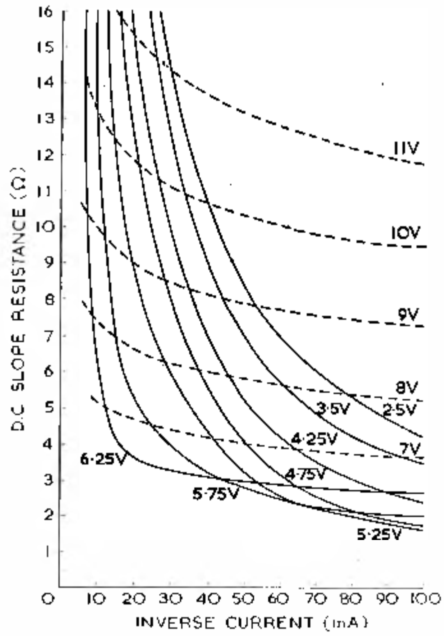

Fig. 8. Typical curves of d.c. slope resistance against inverse current

they are generally found to be sharper, that is having a lower slope resistance. This is shown as a full-line curve in Fig. 6. The dotted line curve is a mean curve taken for high voltage diodes having the same nominal area, showing their higher slope resistance.

The forward temperature coefficient is a function of voltage and only secondarily current, but since the majority of diodes have close characteristics the temperature characteristic can be plotted as a function of current, in a similar way to the inverse characteristics.

The forward characteristics are of small importance for reference purposes apart from their uses as temperature compensation elements, which are considered later.

The Stabilized Supply

General

A stabilized supply possesses the characteristics of delivering a constant output (voltage or current) unaffected by temperature, output load or input voltage, within given limits. The simplest example of a voltage stabilized supply is provided by a Zener diode and a series resistance, as shown in Fig. 10 together with the equivalent circuit. It can be seen that if the diode series resistance and the temperature coefficient were zero the stabilization would be perfect, since the output voltage would equal the diode voltage, provided that the input voltage is greater than the diode voltage plus the series resistance drop. The problem of output voltage variation is considered later, but in order to design such supplies the circuit must be analysed to determine the limits due to diode current ratings and input voltage variation. Such an analysis (for a neon tube) is given by Benson[1], and assumes a constant voltage across the diode. This is included here in order to display the effect of circuit conditions and requirements and enable a design procedure to be adopted.

Determination of Circuit Values

The determination of circuit values for a particular case

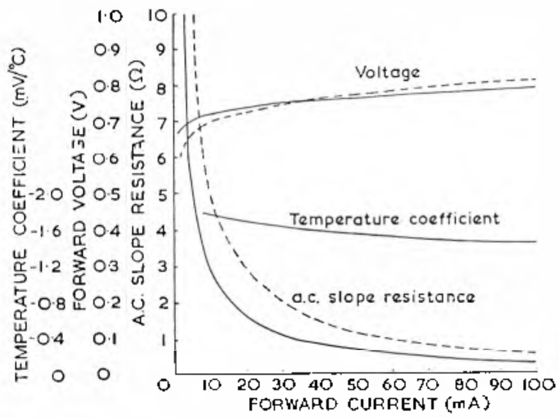

Fig. 9. Typical forward characteristics with current
—— *silicon Zener diodes* · · · *silicon rectifiers*

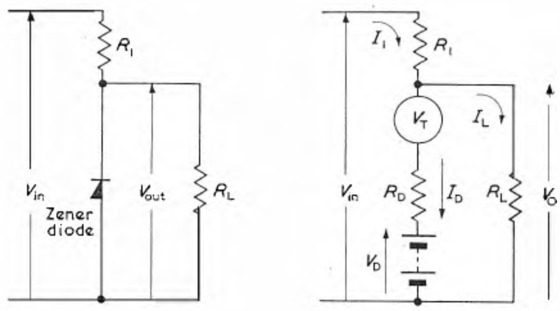

Fig. 10. The simple stabilized supply

requires a knowledge of the following factors:

Diode voltage V_D

Input voltage V_1

Per unit change of input voltage k

Maximum rated diode current $I_{D(max)}$

Minimum rated diode current $I_{D(min)}$

Then it is shown in Appendix 1 that if:

$I_{L(max)}$ = maximum load current

$I_{L(min)}$ = minimum load current

and R_D, the diode resistance is neglected,

then:

$$I_{D(max)}(1-k) - I_{D(min)}(1+k) \geqslant I_{L(max)}(1+k) - I_{L(min)}(1-k) \quad \ldots\ldots\ldots\ldots (3)$$

This shows that the effect of input variation is to reduce the permissible load current swing.

For example, consider a diode with current ratings of $I_{D(min)} = 20$mA and $I_{D(max)} = 250$mA. (That is a 7V diode at 50°C). Then with a minimum load current of zero, the maximum permissible load current swing is plotted against k in Fig. 11(a).

Increasing the minimum load current decreases the permissible swing of load current. In Fig. 11(b) the load current change is plotted against the minimum load current using $k = 0{\cdot}1$, and other values as before.

Also:

$$V_D \leqslant V_1\left\{1 - k\cdot\left(\frac{I_{D(max)} + I_{L(min)} + I_{D(min)} + I_{L(max)}}{I_{D(max)} - I_{D(min)} + I_{L(min)} - I_{L(max)}}\right)\right\} \qquad \ldots\ldots\ldots\ldots (4)$$

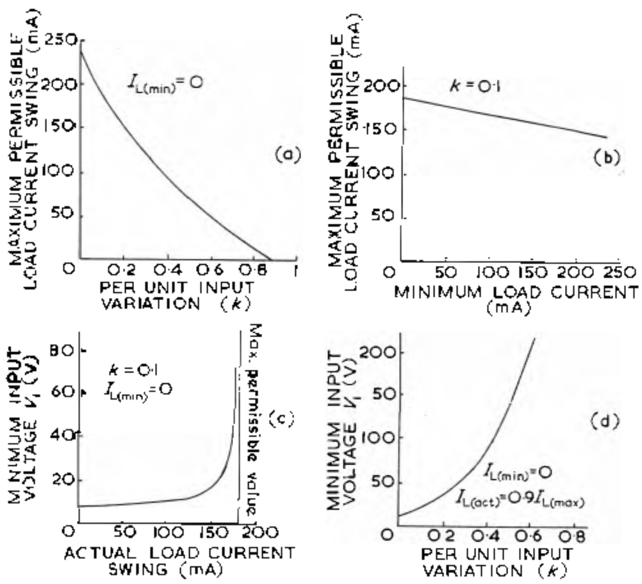

Fig. 11. Circuit limits (7V diode, minimum current 20mA, maximum 250mA)

The previous curves do in fact assume that V_1 can be infinite since equation (3) has been used as an equality. To find V_1 for a given value of $I_{L(max)}$ equation (4) can be used. Taking $I_{L(min)} = 0$, $k = 0{\cdot}1$ the curve of Fig. 11(c) is produced.

It can be seen that as the load current approaches the maximum permissible value, the input voltage rises sharply. In order to achieve a reasonable efficiency, the load current would be limited to about 0·9 of the maximum theoretical value. Taking 0·9 $I_{L(max)}$ as a reasonable load current, the variation of input voltage required for various values of k can be determined. This is shown in Fig. 11(d).

Having determined the values of current and voltage to be used the series resistance R_1 is then given by:

$$\frac{V_1(1 + k) - V_D}{I_{D(max)} + I_{L(min)}} \leqslant R_1 \leqslant \frac{V_1(1 - k) - V_D}{I_{D(min)} + I_{L(max)}} \quad \ldots (5)$$

In the majority of cases, having ensured that the current rating of the diode is suitable for the load current swing and input voltage variation, the resistance R_1 is chosen to ensure a given minimum diode current, so the right-hand side of equation (3) is used, giving:

$$R_1 = \frac{V_1(1 - k) - V_D}{I_{D(min)} + I_{L(max)}} \quad \ldots\ldots\ldots\ldots (6)$$

Temperature Compensation

General

It has been shown in a previous section (Figs 5 and 6) and more fully by Garside and Harvey[2] that a diode with a zero temperature coefficient over a practical range of current and temperature cannot be obtained, and so some form of compensation has to be used. The forms which suggest themselves all use devices with positive or negative temperature coefficients and are shown diagrammatically for a Zener diode with a positive temperature coefficient in Fig. 12. For a diode with a negative temperature coefficient the sign of the compensating element has to be reversed.

In this diagram the series and load resistances are assumed to have zero temperature coefficient.

In the first case, shown in Fig. 12(a), an element in series with the diode is used.

The effective diode slope resistance is now the resistance of the combination, which means a decrease in stabilization against input changes and an increase in the output resistance.

One advantage of this method is that it does not depend on circuit conditions (within its ratings) and so can be a

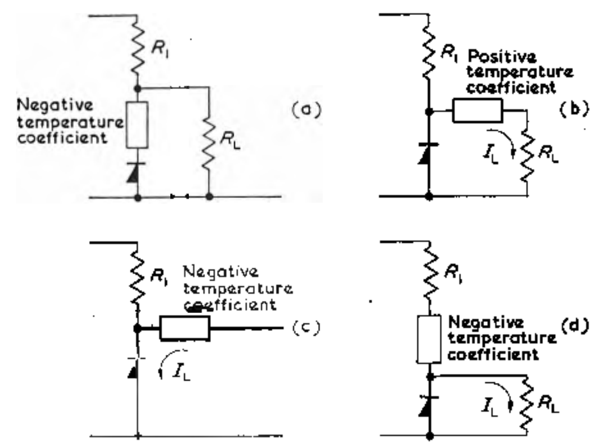

Fig. 12. Temperature compensation (for diode with positive temperature coefficient)

generally applicable device made up by the diode manufacturers.

The second method, Fig. 12(b), uses an element in series with the load resistance, which gives an increase in output resistance without increasing the diode slope resistance. This is mainly used for constant current loads. Where the load current flows in the reverse direction (from a higher potential than the diode) the temperature coefficient of the compensating element has to be reversed, Fig. 12(c).

The fourth case, Fig. 12(d) alters the diode current, but since a good stabilizer is inherently insensitive to diode current, this is not a practical method.

Since the d.c. slope resistance is due to a self-heating effect, if the compensating element can be arranged to be in thermal contact with the diode, the d.c. slope resistance of the combination can be lower than that of the Zener diode alone, thus improving the stabilization.

Devices which are applicable for compensation are firstly low voltage Zener diodes (<5V), forward diodes and thermistors, all with negative temperature coefficients and then higher voltage (>5V) Zener diodes, and metallic resistors with positive temperature coefficients.

Compensation by other Zener Diodes

Since both positive and negative temperature coefficient Zener diodes are available, the use of appropriate pairs would seem an obvious choice. However, the full details of temperature coefficients given previously show that the non-linearity of the low voltage units with current and temperature renders this unsuitable, (except for diodes below $4\frac{1}{2}$V) and in addition the high slope resistance of these diodes gives a high overall slope resistance or output impedance.

Compensation by Forward Diodes

The temperature coefficient here is linear with temperature, negative, and varies only slightly with current (Fig. 9). Used as in Fig. 12(a), the slope resistance of the combination remains reasonably low since the forward slope resistance of a Zener diode is low (Fig. 9). It can be seen that the magnitude of the temperature coefficient would balance

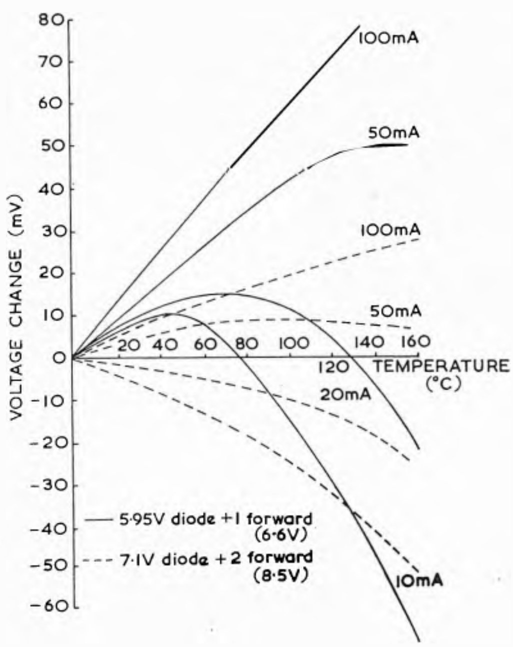

Fig. 13. Temperature compensation by forward diodes
——5·95V diode + 1 forward (6·6V)
- - -7·1V diode + 2 forward (8·5V)

a 5·95V Zener diode, which is slightly non-linear with current and voltage. The experimental figures for a combination of this type are shown in Fig. 13, where the effect of the non-linearity can be seen. In order to overcome this, a 7·1V diode and two forward diodes presents a better choice. Fig. 13 also shows experimental results for a combination of this type.

The value of this method of compensation is shown quite clearly by these curves, and is a practical proposition, since initial selection involves only voltage measurement with the temperature measurement only performed on a selected number of diodes; the 7·1V diode used in Fig. 13 was selected by voltage measurement.

Used as in Fig. 12(b) the forward diode carries the load current, but still provides good compensation.

Fig. 12(c) is mainly included to come under this heading since it represents the case of a Zener diode in the emitter of a transistor (used in a feedback stabilizer) where the base-emitter junction is acting as the forward diode.

Compensation by Thermistor

The long-term stability of currently available thermistors is not known to the author, but in the past they have been regarded as subject to random variations in resistance. However, they may prove suitable for compensating non-linear coefficient diodes.

Compensation by Metallic Resistors

The voltage change across a resistance is directly proportional to the current flowing in it, which renders it unsuitable for Fig. 12(a) but suitable for use as in Fig. 12(b) with constant current loads.

This has no disadvantages at constant current output and is a very suitable method.

Output Voltage Stability

In order to obtain a useful expression for stability, the overall performance of the circuit is required. However, the variation of output voltage can be separated into two components, the first due to input voltage changes, and the second due to load current changes. An overall expression makes the separation of these components, and an evaluation of their effects, difficult. Thus the stability will be considered in two distinct sections, the first due to input changes and the second due to load current changes.

Stability Ratio (Single Zener Stabilizer)

Consider the circuit of Fig. 10 with the various voltages and currents as shown, and a fixed load resistance R_L, and assuming that input voltage changes are small,

then:

$$V_1 = I_1 R_1 + V_o \quad \ldots\ldots\ldots\ldots\ldots \quad (7)$$

Consider a small change, v_1, in V_1,

then:

$$V_1 + v_1 = R_1(I_1 + i_1) + V_o + v_o \quad \ldots\ldots\ldots \quad (9)$$

Now $i_1 = i_D + i_L$

and $i_D = (v_o/R_D)$, and $i_L = (v_o/R_L)$ (assuming R_D to be constant).

Substituting in equation (9)

$$v_1 = R_1\left((v_D/R_D) + (v_o/R_L)\right) + v_o$$

$$v_1 = v_o\left\{1 + R_1((1/R_D) + (1/R_L))\right\} \quad \ldots\ldots \quad (10)$$

In order to obtain a measure of the stabilizing action, a useful criterion is given by:

$$\frac{\text{per unit change of input}}{\text{per unit change of output}} \quad \ldots\ldots\ldots\ldots \quad (11)$$

Let equation (11) be known as S, the stability ratio. Then S indicates the degree of improvement in per unit voltage changes, so if $S = 100$ and the input changes by 0·1 (10 per cent) the output changes by 0·001 (0·1 per cent).

Thus:

$$S = (v_1/V_1)\,.\,(V_o/v_o) \quad \ldots\ldots\ldots\ldots\ldots \quad (12)$$

Substituting equation (10) in equation (12):

$$= (V_o/V_1)\ \{1 + R_1\left(1/R_D\right) + (1/R_L))\} \quad \ldots\ldots \quad (13)$$

Now R_1 is a function of V_1, V_o and R_L

$$\text{and } R_1 = \frac{V_1 - V_o}{I_D + I_L} \text{ and also } I_L = (V_o/R_L)$$

so

$$S = (V_o/V_1) + (1 - (V_o/V_1))\left(\frac{1}{I_D + I_L}\right)((V_o/R_D) + I_L) \quad (14)$$

This expression can now be examined to observe the effect of the various parameters.

Stability Ratio with Output/Input Voltage Ratio

Using equation (14), and assuming values of $V_o = 7$V,

I_D = 20mA, $R_D = 4{\cdot}6\Omega$ and taking I_L = 20mA, the effect of changing the input voltage can be determined. This is shown in Fig. 14(a) and shows that a high value of S, V_o/V_1 must be small, that is V_1 must be high. In addition it was shown previously that the input voltage must be high to permit large current swings and input variations.

STABILITY RATIO WITH LOAD CURRENT CHANGE

The change referred to here is a change in the mean load current, the actual voltage change due to the swing in load current is given in a later section. From the previous section, taking V_o/V_1 as 0·1 and the other values as before the curve of Fig. 14(b) is obtained. From this it can be seen that the larger the load current the lower the stability ratio.

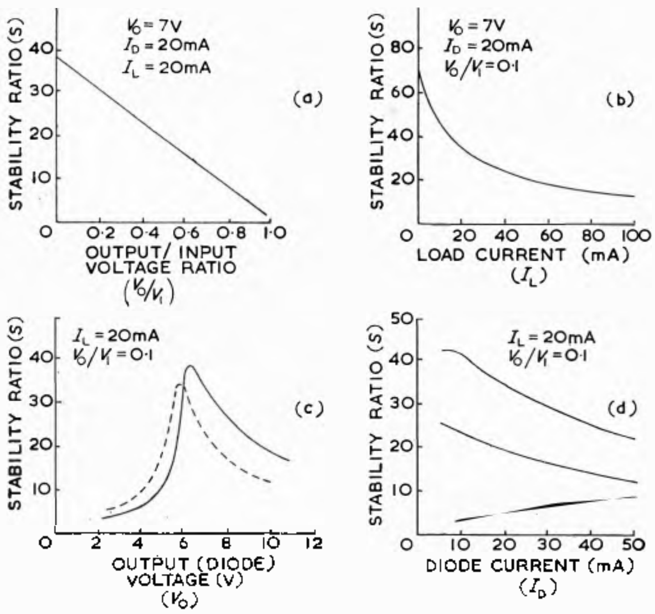

Fig. 14. Variation of stability ratio

STABILITY RATIO WITH OUTPUT VOLTAGE

In order to evaluate this, the values of diode slope resistance must be taken from Fig. 7, and again using I_L = 20mA, $V_o/V_1 = 0{\cdot}1$ the curves of Fig. 14(c) are produced, for diode currents of 20 and 50mA. A maximum is observed at 6·2V, (for 20mA) due to the minimum slope resistance (at 20mA) at this point, and a maximum at 5·8V for the 50mA curve. If temperature compensation thermally connected to the diode is used the higher values of voltage give higher values of S.

STABILITY RATIO WITH DIODE CURRENT

Again considering $V_o/V = 0{\cdot}1$, I_L = 20mA, and taking values of I_D, the corresponding values of R_D must be determined. The manner in which R_D varies has been shown to depend on the voltage, so 4, 7 and 10V diodes are shown in Fig. 14(d).

It can be seen that a peak is observed at about 10mA for the 7V diode, but the 4V diode improves with current. These curves are all calculated from the typical data and any particular diode may vary, especially below 20mA.

STABILITY WITH OUTPUT CURRENT CHANGE

If the slope resistance is considered as constant, then the output voltage change is the current change multiplied by the slope resistance. From Fig. 8, this assumption is seen to be reasonable for diodes of 6V and above.

Then:

$$V_o = (I_{L(max)} - I_{L(min)})\,R_D \quad \ldots\ldots\ldots\ldots (15)$$

For diodes below 6V, the slope resistance varies inversely as the current, (approximately, for 3 to 4V diodes this is a very close approximation).

So if $R_D = (K/I_D)$

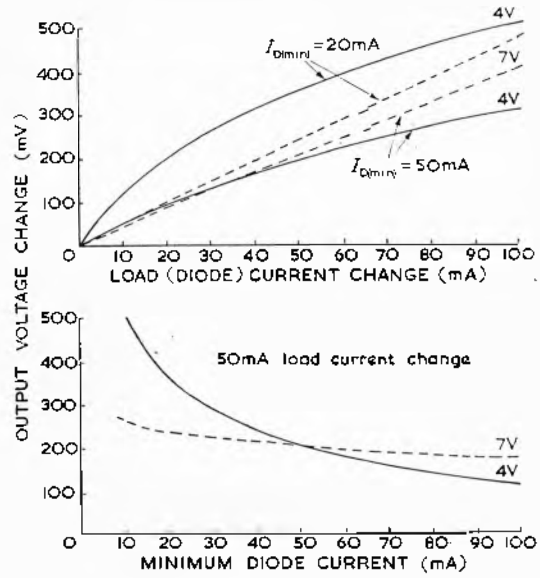

Fig. 15. Output voltage change with load current change

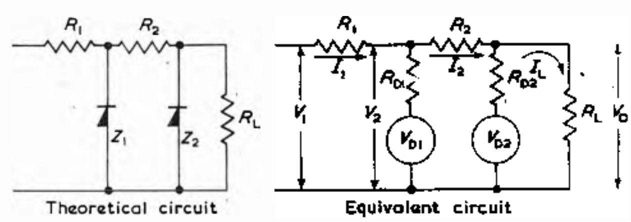

Fig. 16. The double stabilizer

Then:

$$V_o = \int_{I_{D(min)}}^{I_{D(max)}} (K/I_D)\,dI_D = K \log \frac{I_{D(max)}}{I_{D(min)}}$$

$$= K \log \frac{I_{L(max)} - I_{L(min)} + I_{D(min)}}{I_{D(max)}} \quad \ldots\ldots (16)$$

Hence the variation depends on $I_{D(min)}$ as well as the magnitude of the load current swing.

The output voltage change against load current is shown in Fig. 15(a) for 4 and 7V diodes at 20 and 50mA, and Fig. 15(b) shows the effect of the minimum diode current for a 50mA load current swing.

The Double Stabilizer

STABILITY RATIO (DOUBLE ZENER STABILIZER)

The double stabilizer consists of one stabilizer cascaded with a second stabilizer. The arrangement is shown in Fig. 16, and by a method similar to that given for the single Zener stabilizer the stability ratio:

$$S = V_o/V_1 \div \frac{1 - V_2/V_1}{I_1}\left\{(V_o/R_{D1}) + \left(\frac{V_2 - V_o}{I_2 R_{D1}} + 1\right)((V_o/R_{D2}) + I_L)\right\} + \frac{V_2 - V_o}{V_1 I_2}((V_o/R_{D2}) + I_L) \quad \ldots\ldots (17)$$

If the load on the first stage is considered as a current drain I_2, and the two stability factors are multiplied together,

then:

$$S_1 . S_2 = V_o/V_1 + \frac{1-V_2/V_1}{I_1}\left\{V_o/R_{D1} + \left(\frac{V_2-V_o}{I_2R_{D1}} + 1-(V_o/V_2)\right)\left((V_o/R_{D2})+I_L\right) + \frac{V_oI_2}{V_2}\right\} + \frac{V_2-V_o}{V_1I_2}\left((V_o/R_{D2})+I_L\right) \quad \ldots\ldots\ldots\ldots (18)$$

When the first expression, equation (17), is examined

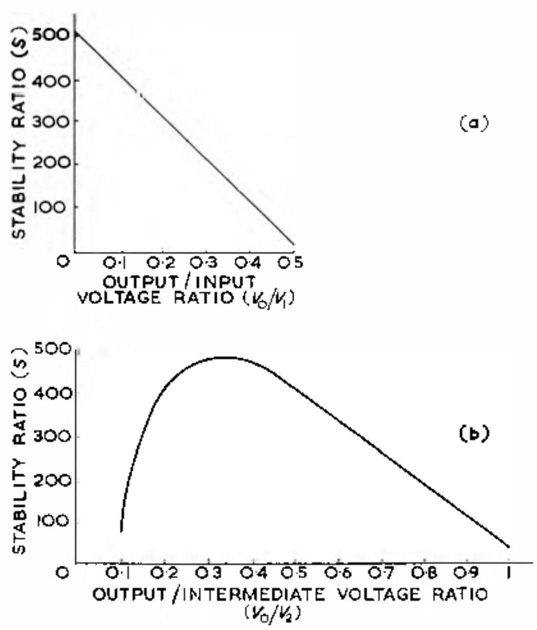

Fig. 17. Stability ratio for double stabilizer

for the effect of V_o/V_1 and V_2/V_1 it becomes evident that expressions (17) and (18) are nearly equal, since V_o/V_2 is small compared to $\frac{V_2-V_o}{I_2R_{D1}} + 1$, and V_oI_2/V_2 is small compared with the rest of the terms in its brackets.

STABILITY WITH OUTPUT/INPUT/INTERMEDIATE VOLTAGES

As would be expected from the previous section, the variation of the stability ratio with input/output voltage ratio change is similar to the single stage, when a fixed value of intermediate/output voltage ratio is considered. It is of course much larger, and the variation is shown in Fig. 17(a) for the following values:

$$V_o = 7\text{V} \quad V_2 = 14\text{V} \quad R_{D1} = 9{\cdot}4\Omega$$

$$R_{D2} = 4{\cdot}7\Omega \quad I_{D1} = I_{D2} = I_L = 20\text{mA}.$$

The further variation which is possible in this case is a change in intermediate/output voltage ratio. The effect of this is shown in Fig. 17(b) where a fixed ratio of $V_o/V_1 = 0{\cdot}1$ is assumed, and the intermediate voltage is varied in steps of 7V, the diodes each having a slope resistance of 4·7Ω.

It can be seen that the maximum value of S occurs at $V_o/V_2 = 0{\cdot}316$, which is the point at which V_2/V_1 is also equal to 0·316, i.e. where $V_o/V_2 = V_2/V_1$ (this is inherent for a fixed value of V_o/V_1 since S is approximately equal to $S_1 . S_2$).

The Design of a High Stability Reference

A unit was required in the laboratory as an adjunct to a precision potentiometer to enable the simultaneous measurement of two variables. The requirements for this supply are to give fixed reference voltages of 100, 200 and 500mV to an accuracy of ±0·1 per cent, and a suitable impedance was 50Ω at 500mV, that is a current drain of 10mA.

From the previous sections, since the unit is to be run from the mains with a possible variation of ±10 per cent, a double stage stabilizer is used, with a high value of input voltage (Fig. 17(a)) an intermediate/output ratio near 3 (Fig. 17(b)) and a diode voltage near 6V (Fig. 14(c)). Since the output current is constant, temperature compensation by a copper coil in series with the load, and wound on to

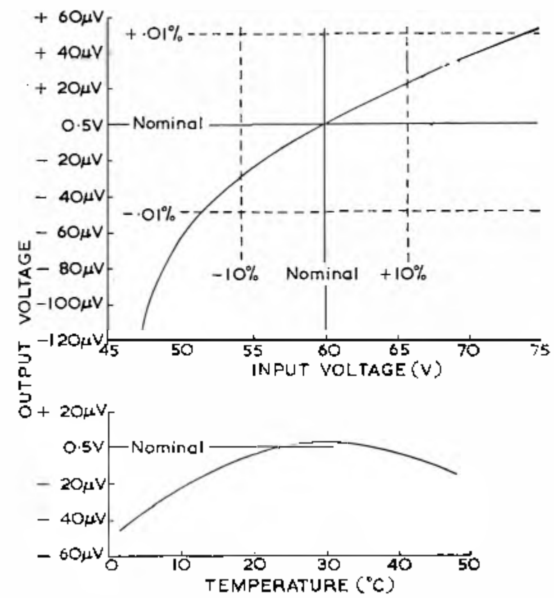

Fig. 18. Performance of practical circuit

the body of the output diode is used. In order to obtain linearity of the diode temperature coefficient, diodes about 6·3V are used.

Taking 6·3V as the intermediate voltage with 60V as the input voltage, and running the diodes at the recommended current of 20mA, it is found from Fig. 7 that the d.c. slope resistance for each diode is 3·6Ω. Then from equation (17), $S = 950$. It can be assumed that S will be higher than this since the temperature compensation is thermally connected to the diode, so that the d.c. slope resistance will approach the a.c. value. If $R_{D2} = 2{\cdot}0\Omega$ (assuming the effective thermal resistance to be equal to 35°C/W then $S = 1690$.

From equation (6), $R_1 = 880\Omega$

$$R_2 = 420\Omega$$

In order to compensate for temperature changes, a copper coil in series with the load is used.

If V_T is the diode voltage change per degree Celsius,

Then:

$$V_T = V_{(copper)} = I_L \times \alpha_{(copper)} \times R_{(copper)}$$

assuming a zero temperature coefficient load resistance.

For a 6·3V diode, $V_T = 2{\cdot}4\text{mV}/°\text{C}$

so at 10mA load current, $R_{(copper)} = 55\Omega$.

The performance of the practical circuit is shown in Fig. 18, with both input voltage and temperature changes. The copper coil was actually increased to 63Ω to give better compensation, and the stability ratio can be seen to be about 1 750 at the nominal input voltage, which is a close approximation to the calculated value. Due to the temperature effects being compensated, rapid changes of temperature cause larger changes of output voltage than shown by Fig. 18. However, by suitable thermal design this can be minimized.

Conclusions

The characteristics given in the section 'Characteristics' are suitable for describing the circuit behaviour of the Zener diode in all types of stabilized supplies, and the performance can be calculated with reasonable accuracy using these characteristics.

The a.c. slope resistance is a very useful parameter, since it is easy to measure, and provides a suitable check on the 'goodness' of the diode. For a given value of diode voltage, provided that reasonable limits are placed on the a.c. slope resistance, the temperature coefficient (and hence d.c. slope resistance) can be predicted.

The diode ratings give the limiting values of load current and input voltage for a given amount of input variation, and it has been shown that if the input variations are large, high input voltages or low load currents have to be used.

The temperature variation of breakdown voltage can be compensated by several methods, and as shown in the practical example, can be reduced to a value much less than the temperature variation of a standard cell, over a reasonable range of temperature (40°C or more). This points to a difficulty encountered in making the measurements, in that the variation per degree Celsius of a standard cell (40μV) is as large as the variation over 40°C for the Zener diode circuit, when the compensation is adjusted correctly.

The stability ratio enables the effect of input variations to be calculated, and shows the steps required to produce a high stability ratio. The graphs given in Fig. 14 have been verified experimentally with reasonably close agreement.

The output voltage change with load current alters according to the value of the load current change and diode current, and for larger changes the lower voltage diodes become superior, for the simple air cooled diode. The effect of lowering the thermal resistance on the d.c. slope resistance has been considered, and this improves the higher voltage units considerably.

The double stabilizer circuit, even without the addition of refinements such as bridge connexion, enables high stability reference sources to be constructed. The practical example verifies the validity of the method, and since the output current in this case is 10mA, this suggests that equipment such as potentiometers could be supplied directly from a reference source. Long term precision measurements would be required to establish the limit of accuracy obtainable.

Acknowledgments

The author wishes to thank the Director of Research, A.E.I. (Rugby) Research Laboratory, for permission to publish this article.

Appendix

Let V_D = Diode voltage

V_1 = Input ,,

k = per unit change of input voltage

$I_{D(max)}$ = maximum rated diode current

$I_{D(min)}$ = minimum rated diode current

$I_{L(max)}$ = maximum load current

$I_{L(min)}$ = minimum load current

R_1 = series resistance

$I_{1(max)}$ = maximum input current

$I_{1(min)}$ = minimum input current

Then:

$$I_{1(max)} - I_{L(min)} \leqslant I_{D(max)} \quad \ldots\ldots\ldots\ldots (19)$$

$$I_{1(min)} - I_{L(max)} \geqslant I_{D(min)} \quad \ldots\ldots\ldots\ldots (20)$$

Now:

$$I_{1(max)} = \frac{V_1(1+k) - V_D}{R_1}$$

Assuming V_D is constant (i.e. $R_D = 0$).

and:

$$I_{1(min)} = \frac{V_1(1-k) - V_D}{R_1}$$

So:

$$\frac{V_1(1+k) - V_D}{I_{D(max)} + I_{L(min)}} \leqslant R_1 \leqslant \frac{V_1(1-k) - V_D}{I_{D(min)} + I_{L(max)}} \quad \ldots\ldots (21)$$

So:

$$\{V_1(1+k) - V_D\}(I_{D(min)} + I_{L(max)}) \leqslant \{V_1(1-k) - V_D\}(I_{D(max)} + I_{L(min)}).$$

$$V_i(1+k)(I_{D(min)} + I_{L(max)}) - V_D(I_{D(min)} + I_{L(max)}) \leqslant V_i(1-k)(I_{D(max)} + I_{L(min)}) - V_D(I_{D(max)} + I_{L(min)}).$$

$$V_D(I_{D(max)} - I_{D(min)} + I_{L(min)} - I_{L(max)}) \leqslant V_i\{(I_{D(max)} + I_{L(min)} - I_{D(min)} - I_{L(max)}) - k(I_{D(max)} + I_{L(min)} + I_{D(min)} + I_{L(max)})\}$$

$$V_D \leqslant V_i\left\{1 - k\left(\frac{I_{D(max)} + I_{L(min)} + I_{D(min)} + I_{L(max)}}{I_{D(max)} - I_{D(min)} + I_{L(min)} - I_{L(max)}}\right)\right\} \quad \ldots\ldots\ldots\ldots (22)$$

Now since $V_i > V_D$

$$I_{D(max)} - I_{D(min)} > I_{L(max)} - I_{L(min)} \quad \ldots\ldots (23)$$

and since V_i and V_D are of the same sign

$$k\left\{\frac{I_{D(max)} + I_{L(min)} + I_{D(min)} + I_{L(max)}}{I_{D(max)} - I_{D(min)} + I_{L(min)} - I_{L(max)}}\right\} < 1$$

Hence:

$$kI_{L(max)} + I_{L(max)} + kI_{L(min)} - I_{L(min)} < I_{D(max)} - kI_{D(max)} - I_{D(max)} - kI_{D(min)}$$

or:

$$I_{D(max)}(1-k) - I_{D(min)}(1+k) > I_{L(max)}(1+k) - I_{(min)}(1-k) \quad \ldots\ldots\ldots\ldots (24)$$

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2. Garside, Harvey. The Characteristics of Silicon Voltage Reference Diodes. International Convention on Transistors and Associated Semiconductor Devices. (London, May, 1959).

A Transistor-Thermistor Feedback Quadrature Suppressor

By I. C. Hutcheon*, M.A., A.M.I.Mech.E., A.M.I.E.E., and D. N. Harrison*

Quadrature effects in a.c. servo systems are eliminated if the amplified residual signal is demodulated and used to control a negative feedback signal precisely in quadrature phase with the a.c. reference. A design of suppressor is described which uses transistor pre-amplification and two indirectly-heated thermistors to control the feedback.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

Quadrature Effects in A.C. Servo Systems

TWO conditions must be satisfied for an a.c. servo system to arrive at a perfect balance, namely equality in both magnitude and phase of the input and the feedback signals. This contrasts with d.c. servos in which the only requirement is equality in magnitude.

If the phase of the input signal differs from that of the feedback by as little as one or two degrees, quite substantial errors may arise in the magnitude of the feedback signal, making it no longer an accurate representation of the input.

Consider the servo system shown in Fig. 1(a). A is an a.c. amplifier, and M is a servo motor which drives a slide-wire contact, so providing a feedback signal of variable magnitude and reference phase. Suppose that the input signal has two components, namely V_r at reference phase and V_q in quadrature with it, and that the amplifier and motor have their maximum response at some angle ϕ to V_r. ϕ should be zero, of course, but it seldom is.

Assuming the system to have infinite loop gain, it balances at a position where there is zero signal at angle ϕ, i.e. where:

$$(V_r + V_1)\cos\phi = V_q \sin\phi,$$

as can be seen from Fig. 1(b). (More strictly, all the residual signal is in the direction of zero response, but this is assumed to be $\phi \pm 90°$ and so the two statements are the same). The feedback signal at balance differs slightly in magnitude from the reference-phase component of the input signal, the fractional error being:

$$\frac{V_1 + V_r}{V_1} = (V_q/V_1)\tan\phi = (V_q/V_r)\tan\phi$$

very nearly,

and there exists a residual signal $V_q \sec\phi$ at an angle $\phi \pm 90°$.

The physical reason for the error is quite simply that the quadrature component is capable of driving the motor unless ϕ is zero. Since ϕ is the sum of the phase-shift through the amplifier and phase errors in the demodulator it may in practice be quite large. If the demodulator is a two-phase servo motor, for example, production spreads in value of the associated resonating and phase-shifting capacitors, and of the motor inductance itself may easily give rise to errors of 10 or 20°, to which must be added any shift through the amplifier. If $\phi = 20°$, then $\tan\phi = 0{\cdot}36$, and a 2 per cent quadrature component gives an error of 0·7 per cent in the magnitude of the feedback signal.

Causes of Quadrature

Quadrature components may arise in several ways, the main causes being as follows:

(a) The combined effect of input signal source resistance and transmission line self-capacitance.

(b) Pick-up, either magnetically induced or capacitive.

(c) Phase shifts through transformers etc. in the input line.

(d) Phase shifts through transformers etc. in the feedback. These may partially compensate for (c).

(e) Other effects in the source.

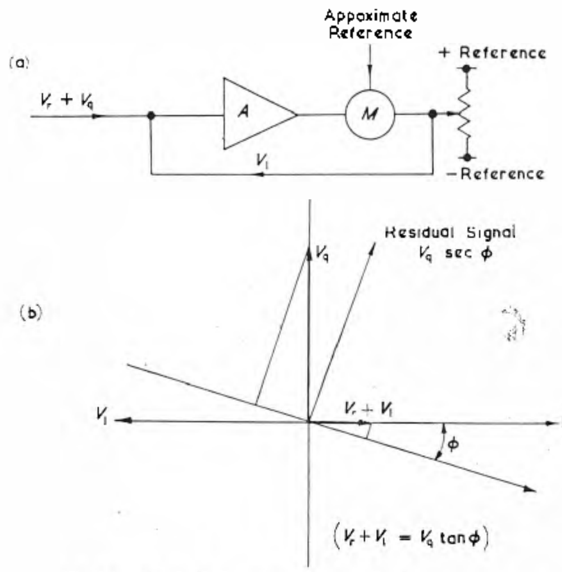

Fig. 1. Effect of input quadrature signal on a.c. servo

Any or all of these causes may be present simultaneously, and it is worth noting that the rotation of a signal by a few degrees has very little effect on its magnitude, but introduces a quadrature component of 1·75 per cent/degree.

To quote just one example[1], a typical small industrial magnetic flowmeter has a source resistance which varies between a few hundred ohms for high-conductivity liquids, and 40kΩ for soft tap water. If the output from the flowmeter is transmitted to an a.c. self-balancing recorder via a 100ft cable whose self-capacitance is 30pF/ft, (giving a shunt reactance of 1MΩ at 50c/s) then a quadrature component is introduced, whose magnitude varies from almost zero to 4 per cent. If the recorder response were a maximum for signals at 20° to the reference (which in this case would be the flowmeter field current) this amount of quadrature signal would cause a recording error varying between zero and 1·4 per cent of the full-scale reading.

Quadrature Suppression

Reducing ϕ to zero appears at first sight to be a good

* *George Kent Ltd.*

solution, but in fact is not for two reasons. Firstly, it is troublesome, since the motor and the amplifier are usually separate units, and each must be phase-corrected individually if interchangeability is required. Secondly, the residual signal $V_q \sec \phi$ is not significantly reduced by making ϕ zero, and may block the amplifier, so rendering it insensitive to inputs of reference phase. It is much better to remove the quadrature signal itself before it reaches the servo amplifier.

The simplest method is to inject a compensating quadrature signal whose amplitude is adjusted manually until the residual balance signal, which can be observed at the amplifier output, is a minimum. This is quite satisfactory if the quadrature effect is a stable one, though it does require the use of an oscilloscope or other indicator. Sometimes the servo motor itself can be used as the indicator, the supply to its power winding being temporarily shifted through 90°, or alternatively a 90° shift can temporarily be introduced within the amplifier.

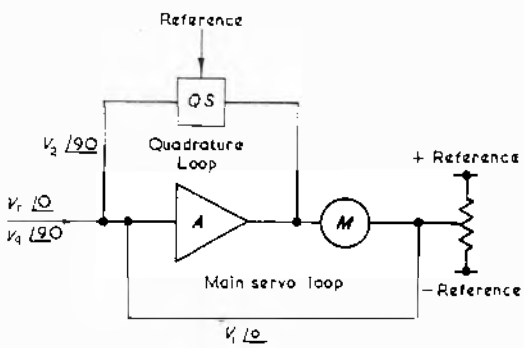

Fig. 2. System with feedback quadrature suppression

Variable quadrature effects are much more serious, and essentially require some form of automatic suppression or elimination.

One method consists in the introduction of a series type of quadrature eliminator[2] between the input signal and the amplifier. This demodulates the signal, smoothes the resultant output, and remodulates it at precisely the correct phase, the quadrature component thereby being removed. However, the device has to work at low signal levels, and the smoothing circuit may introduce a significant lag into the system.

By far the most satisfactory solution is the use of some form of feedback quadrature suppression, and Fig. 2 is a block diagram showing how this is applied. Any residual quadrature signal appears in amplified form at the output of the main servo amplifier and is demodulated, smoothed and remodulated at the correct phase by the quadrature suppressor, then being fed back to the input where it cancels the residual signal.

The phase of the feedback signal must be precisely correct if further errors are to be avoided, but phase errors in the demodulator and phase shift through the main amplifier are relatively unimportant, causing merely a loss in sensitivity.

In one version of this method, the main amplifier drives a second a.c. servo motor whose power winding is supplied in such a way that the motor responds to quadrature signals. This motor drives the contact on a slidewire fed from a precise quadrature supply, and so provides the quadrature-suppressing feedback.

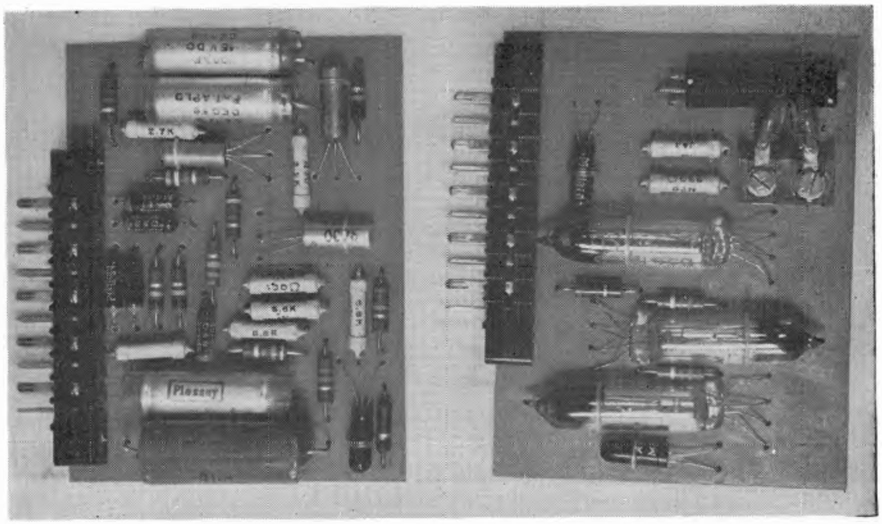
Fig. 3. Thermistor potentiometer
(The two parts bolt together, forming a single plug-in unit)

Precisely analogous to the above is the transistor-thermistor quadrature suppressor described in the following section. Two indirectly-heated thermistors take the place of the motor and slidewire, and there are no moving parts to wear out.

Transistor-Thermistor Quadrature Suppressor

Fig. 3 is a photograph of a 'thermistor potentiometer' which, with the addition of an a.c. reference supply and an output phase-shifting capacitor, becomes a quadrature suppressor[3,4]. Fig. 4 is a block diagram, and the complete circuit, less power supplies, is given in Fig. 5.

As Fig. 4 shows, the quadrature suppressor comprises an optional input phase-shifting circuit, an a.c. amplifier, a phase-sensitive demodulator, a variable potential divider using two thermistors, and a capacitor for shifting the output by 90°. The plug-in unit also carries the components required for providing d.c. power supplies from a centre-tapped transformer winding giving 10-0-10V r.m.s. ±10 per cent. Four germanium transistors are employed, and overall d.c. feedback via a third thermistor gives stabilization against changes in ambient temperature between 0 and 60°C.

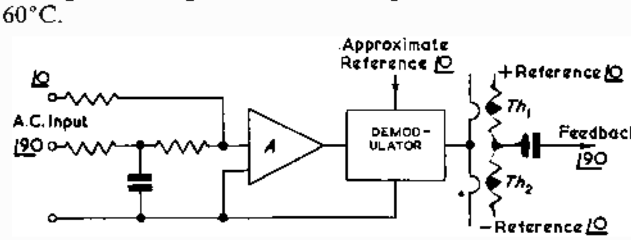

Fig. 4. Arrangement of quadrature suppressor

The device operates as follows. The output of the main servo amplifier is applied to the transistor amplifier via either a resistor or an *RCR* circuit giving approximately 90° lag. Which of these is chosen depends on the phase of the main servo carrier signal, the aim being to arrange that the undesired quadrature effects drive the demodulator which, for convenience, operates directly from the supply.

When driven by the signal, the demodulator increases the power supplied to the heater of one of the output

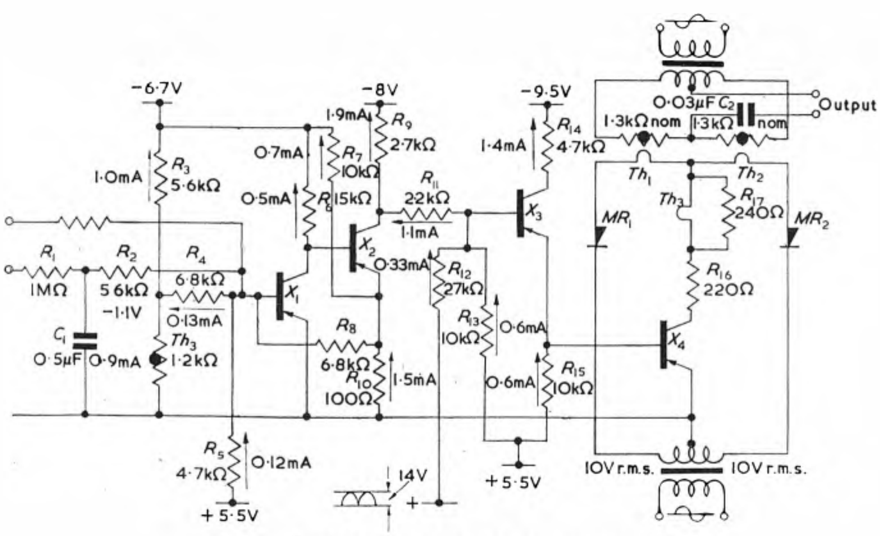

Fig. 5. Circuit of the quadrature suppressor

thermistors Th_1 and Th_2, while decreasing the power supplied to the other. This raises the temperature and lowers the resistance of one of the thermistor beads, while lowering the temperature and raising the resistance of the other. The demodulator is phase-sensitive, and a reversal of signal phase reverses the above effects.

The two thermistor beads thus form a resistive potential divider, their common point being movable either way about the centre of the total resistance. If a centre-earthed reference supply is connected across the divider, a small quadrature feedback current can be taken off through a small capacitor connected to the common point. Larger outputs can be obtained if the phase-shifting is performed before the divider, instead of after it.

An input to the amplifier of $\pm 1\mu A$ r.m.s. swings the output by ± 75 per cent of the theoretical maximum, and the overall characteristic is reasonably linear up to this point. The corresponding input, when the phase-shifting circuit is used, is $\pm 10V$ r.m.s. across $1M\Omega$. About 2V can be connected across the divider without overloading the thermistors, giving a useful output of $\pm 0{\cdot}75V$. The amplifier input resistance is kept low by feedback, being about 150Ω, while the output resistance of the potential divider varies between about 500 and 700Ω according to the drive. Thermal lag in the thermistors causes the response to be fairly slow, a large change in output taking several seconds to complete as can be seen in Fig. 8. This is usually no disadvantage, however, since the quadrature signal itself rarely changes rapidly. An advantage accruing from the thermal lag is that the gain of the quadrature signal loop, comprising the main amplifier and the quadrature suppressor, can be increased to a value of about 500 before self-oscillation sets in. Quadrature effects are reduced by a factor equal to the quadrature loop gain, and a value of 100 or less is generally sufficient to make them quite negligible.

Circuit Details

Transistors X_1 and X_2 constitute the a.c. amplifier, while X_3 and X_4, together with two diodes, form the demodulator. The circuit is direct coupled from end to end, and overall d.c. feedback is applied via a third thermistor Th_3 to stabilize the output characteristics.

Overall D.C. Feedback

Th_3 is connected to the amplifier input in such a way that a ± 5 per cent deviation in its resistance from a nominal $1{\cdot}2k\Omega$ is sufficient to swing the power in all three thermistor heaters from 0 to 30mW simultaneously. This, in turn, would vary the resistance of Th_3 very widely, and so in fact controls it within much less than ± 5 per cent of $1{\cdot}2k\Omega$, transistor drifts being neglected.

When there is no a.c. drive, all three heaters receive the same power, and all the beads have the same predetermined resistance of $1{\cdot}2k\Omega$, whatever the ambient temperature, any rise in temperature being counteracted by a fall in the heater dissipations. This arrangement has the great advantage of stabilizing the output characteristics of the potential divider, which otherwise would be very temperature-dependent.

All transistor drift effects can be compensated by a change of about $\pm 20\mu A$ in X_1 base current, and this too is provided by a small change (± 15 per cent) in Th_3 resistance. Thus if the thermistors are identical, Th_1 and Th_2 resistances are held within ± 15 to 20 per cent of their nominal value for all ambient temperatures between 0 and

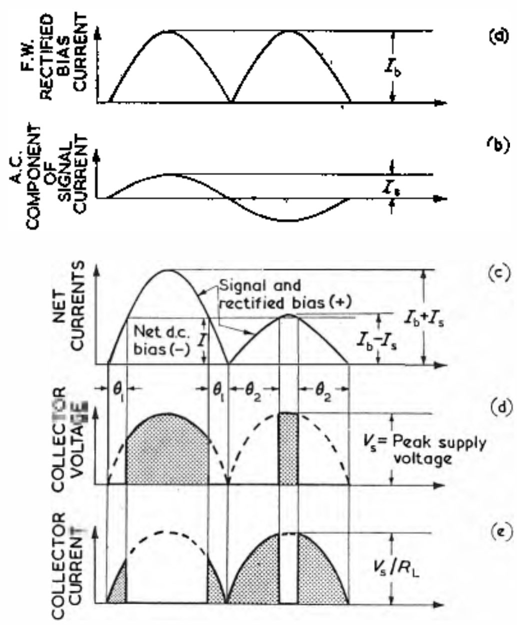

Fig. 6. Demodulator waveforms

60°C. Since there is 1mW additional dissipation in Th_3 bead due to the bias circuit, the nominal resistance of the other two beads is slightly higher in practice, at $1{\cdot}3k\Omega$.

A.C. Operation

The current gain of the amplifier section is determined by feedback from X_2 emitter to X_1 base at a value $R_8/R_{10} = 68$ approximately. The voltage dropped across R_{10} is never less than 70mV, and this provides adequate collector potential for X_1 which handles only small signals.

X_2 output current divides about equally between R_9 and R_{11}, so the current gain to X_3 base is about 34.

X_3 acts as an emitter-follower to drive X_4, and both operate together in switched mode to drive the heaters;

typical waveforms are shown in Fig. 6. The a.c. signal is mixed on X_3 base with a direct positive bias current, a positive full-wave rectified bias, and a negative direct current from X_2 which is varied by the overall feedback. The X_3X_4 combination is thus switched on or off according as the a.c. signal or the d.c. bias predominates. During alternate half cycles, therefore, X_4 collector current comprises two large or two small current pulses as shown in Fig. 6(e). Since Th_1 and Th_2 heaters receive current on alternate half cycles, they are heated differentially by the demodulator. A full analysis of the operation is given in

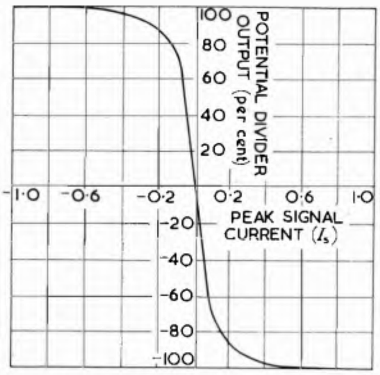

Fig. 7. Gain of demodulator and potential divider

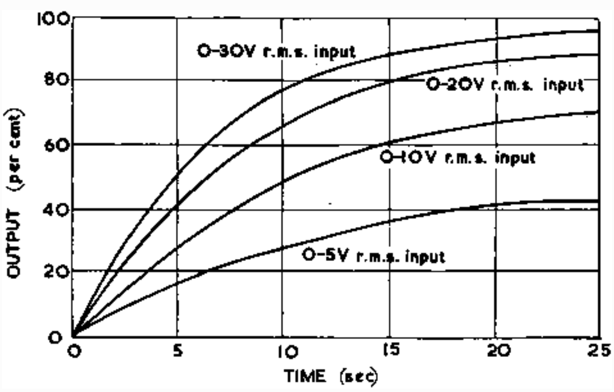

Fig. 8. Response of quadrature suppressor to step input changes

the Appendix and Fig. 7 shows the overall demodulator characteristic in terms of the potential divider percentage output against I_s/I_b, where:

I_s = Peak signal current supplied to X_3 base

I_b = Peak full-wave rectified bias current = 0·5mA

V_s = Peak demodulator supply voltage, = $10\sqrt{2}$V

R_L = Demodulator load resistance on either half cycle, = 390Ω

$Th_1, Th_2\ Th_3$ = S.T.C. Thermistors type B54 (heater resistance 100Ω)

MR_1, MR_2 = Mullard OA5 junction diodes

Interstage bias arrangements are such that the collector leakage of each transistor may reach the maximum value specified for a junction temperature of 70°C (0·4mA for X_2, X_3 and X_4 if these are type OC76 transistors), without limiting the ability of X_3 and X_4 to switch fully on and off. Dissipation in X_3 and X_4 is low on account of the switched mode of operation, while X_1 and X_2 operate with low collector voltages and so also have low dissipations. In no case, therefore, does the junction temperature rise more than a few degrees above the ambient temperature, and this in consequence may safely reach 60°C. In practical tests, a number of units operated satisfactorily at 70°C for several hundred hours. In order to minimize the drift which the d.c. feedback has to correct, X_1 is a low-leakage transistor of the OC45 type.

Appendix

Analysis of Demodulator Operation

The heater currents controlled by X_4 are never greater than 35mA and since the current gain of the X_3X_4 combination exceeds about 1 000, they are switched on and off by a change of 35μA or less in X_3 base current. This is very small in comparison with the signals applied to X_3 base, and switching is therefore sharp within the potential divider operating range of ±75 per cent of full output.

X_3 base is supplied with a positive full-wave-rectified bias current, the a.c. signal, a positive d.c. bias and a variable negative direct current from X_2 which is varied by the overall feedback to control the dissipation in Th_3 heater. The rectified-bias and signal waveforms, shown in Figs. 6(a) and 6(b) combine to give the waveform shown in Fig. 6(c), and this switches X_3 and X_4 off or on according as it is greater or less than the net negative steady d.c. bias. In practice a positive current can never flow into X_3 base which therefore makes a positive excursion, but to make description simpler it can be assumed that a diode is connected between X_3 base and earth, and carries the excess positive current. The circuit performance remains unaffected if this is done.

The collector voltage and current waveforms resulting from this mode of operation are shown in Figs. 6(d) and 6(e). The voltage dropped across X_4 and the diodes is neglected in the analysis, but compensated in practice by reducing the load R_L from a theoretical 410Ω to 390Ω.

The following terminology is necessary in addition to those terms already defined.

I Net steady negative direct current supplied to X_3 base (A)

I_{ms1} Mean square current in Th_1 heater, (A²)

I_{ms2} ,, ,, ,, ,, Th_2 ,, (A²)

I_{ms3} ,, ,, ,, ,, Th_3 ,, (A²)

θ_1, θ_2 Demodulator conducting angles for alternate half cycles (degree)

Switching occurs at conducting angles given by:

$$\sin\theta_1 = \frac{I}{I_b + I_s} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

and

$$\sin\theta_2 = \frac{I}{I_b - I_s} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

whence

$$I_s/I_b = \frac{\sin\theta_2 - \sin\theta_1}{\sin\theta_2 + \sin\theta_1} \quad \ldots\ldots\ldots\ldots (3)$$

while the mean square currents in the heaters of the two output thermistors are:

$$I_{ms1} = \tfrac{1}{4}(V_s/R_L)^2\left(\frac{2\theta_1 - \sin 2\theta_1}{\pi}\right) \ldots\ldots (4)$$

$$I_{ms2} = \tfrac{1}{4}(V_s/R_L)^2\left(\frac{2\theta_2 - \sin 2\theta_2}{\pi}\right) \ldots\ldots (5)$$

Th_1 and Th_2 heater currents pass alternately through Th_3 heater, and this is shunted so that:

$$I_{ms3} = \tfrac{1}{2}(I_{ms1} + I_{ms2}) \quad \ldots\ldots\ldots\ldots (6)$$

whatever the value of the a.c. drive. If the drive is zero, then $I_{ms1} = I_{ms2} = I_{ms3}$.

The overall d.c. feedback is arranged so that, when the ambient temperature is midway between its limits (i.e. 30°C) I_{ms3} is held at a value equal to half the greatest possible value of I_{ms1} or I_{ms2}. These have maxima, when X_1 is on for a complete half cycle, of $(\frac{1}{4})\ (V_s/R_L)^2$, thus:

$$I_{ms3} \text{ at } 30°C = \tfrac{1}{8}\,(V_s/R_L)^2 \quad \ldots\ldots\ldots\ldots \quad (7)$$

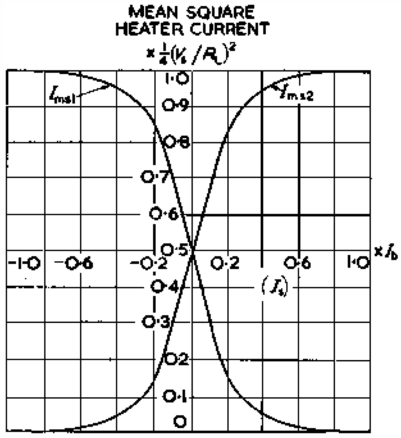

Fig. 9. Mean square heater current versus signal at design-centre ambient temperature

and, from equation (6),

$$I_{ms1} + I_{ms2} = \tfrac{1}{4}\,(V_s/R_L)^2 \text{ at } 30°C \quad \ldots\ldots \quad (8)$$

Thus the a.c. drive increases the dissipation in one heater and decreases that in the other by equal amounts about a design-centre value which, when the ambient temperature is at its design-centre value, is equal to half the maximum dissipation in either.

In practice, the greatest possible dissipation in either Th_1 or Th_2 is made equal to 30mW by suitable choice of V_s and R_L, while Th_3 is controlled at 15mW at 30°C. Thus Th_1 and Th_2 heaters can each be driven ±15mW about a design-centre value of 15mW when the ambient temperature is 30°C.

At higher ambient temperatures the heater dissipations are made to fall (by 2mW every 10°C for B54 thermistors) by the action of the overall d.c. feedback, and the operating range is restricted somewhat by one of them reaching zero before the other reaches 30mW. At lower temperatures the dissipations are made to rise, and the range is similarly restricted since one heater reaches a dissipation of 30mW before the other reaches zero.

These effects, of course, are a necessary result of the chosen method of d.c. stabilization which prevents the output resistance characteristics from altering with ambient temperature. Even in the extreme cases, however, there is no difficulty in achieving ±75 per cent output from the potential divider, and it is quite sufficient to assess the demodulator overall gain at 30°C ambient.

Combining equations (4), (5) and (8) gives:

$$\frac{2\theta_1 - \sin 2\theta_1}{\pi} + \frac{2\theta_2 - \sin 2\theta_2}{\pi} = 1 \text{ at } 30°C \quad \ldots\ldots \quad (9)$$

which can be solved to give the relation between θ_1 and θ_2. Equation (3) now permits I_s/I_b to be related to conducting angle, and hence to mean square heater current as shown in Fig. 9. This gives the heater dissipations in terms of I_s/I_b and hence allows the bead resistances, and the potential divider percentage output, to be calculated for any particular type of thermistor whose characteristics are known. Fig. 7 is the result of such a calculation.

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Remote Control of a Power Station

A remote-controlled power station has recently been brought into operation at Princetown, Devon, by the South Western Electricity Board for operation during peak or emergency conditions.

It consists of a 3MW alternator driven by a 4 250 h.p. land version of the Proteus turboprop engine as used in the Bristol Britannia aircraft and it will be controlled from the Electricity Board's Control Centre at Bristol at a distance of 100 miles, over the normal telephone network.

The control system used is the Mark II 'Datafonic' system designed and manufactured by Sound Diffusion (Auto-Thermatic) Ltd of Portslade, Sussex and represents an extension of a system introduced in the United Kingdom earlier last year.

This equipment was developed to provide a central Control Room with supervisory facilities over remote unattended sub-stations. Such facilities have normally been available in the past only by the use of equipment operating over privately rented G.P.O. lines which, when capitalized over a 10 year basis, represents a considerable investment.

The object of the Mark I was, therefore, to provide equipment that would operate over the national telephone network and, by giving all information concerning breaker positions and other items in a verbal form, eliminate the need for both private lines and decoding equipment in the Control Room. The system was designed to dial or call a Control Room when equipment in the substation changed its state, and conversely, to give information when interrogated by an incoming telephone call.

The equipment installed at Bristol enables the Control Engineer to carry out a total of up to 40 control functions over the remote equipment.

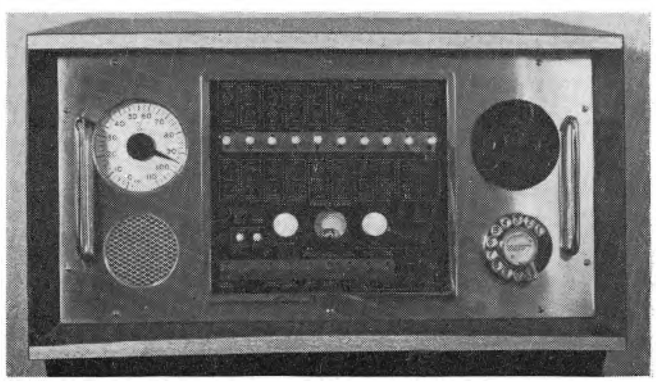
Typical functions are the 'Start', 'Stop' and 'Throttle' controls on the remote Turbo Alternator. These control facilities are supplemented by 22 pairs of indication lamps of which the Turbo Alternator utilizes 20. Telemeter display facilities are also available in the Bristol Control Room.

The control console housing the indication, control and telemetering equipment is a compact unit and is capable of simple extension to allow for the control of up to 24 additional sub-stations.

The equipment operates over the national telephone network and the Control Engineer at Bristol can establish contact with the set by making use of the normal trunk call facilities.

In the event of a change of state in the running condition of the sub-station, the 'Datafonic' equipment in the sub-station automatically engages its own exchange line and verbally requests the exchange operator to connect it with the Bristol Control Room.

The Remote Control Panel



FAST COUNTING CIRCUITS USING E1T TUBES

By V. Radeka*

The possibilities of fast counting with E1T type counter tubes are investigated. Displacement of the electron beam is considered and time required to reach the stable point calculated. The limits of accurate counting against the reliability required are determined theoretically and experimentally. It is found possible to use E1T tubes up to about 10^6 *pulses per second. A circuit diagram of a reliable fast counting decade suitable for general application is presented. The circuit operation is entirely independent of the input waveform and supply voltage variations. Only one d.c. supply voltage is needed.*

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

THE principles of the decade counter tube type E1T and circuits for counting up to 10^5 pulses per second are exhaustively described in the literature[1,2,3]. Increasing of the counting speed was proposed by Jonker *et al*[2]. The idea was further developed by Barneveld[4] and Porat[5], but a reliable circuit suitable for general application so far has not been published. The analysis given in this article is meant to define criteria in considering various applications of E1T. The behaviour of the beam and the theoretical limits of the counting speed are discussed. On the basis of the given analysis the fast counting circuit is developed.

Criteria in Fast Counting with E1T

The decade counter tube type E1T, as described by Jonker *et al*[2]., is a small cathode-ray tube with one dimensional deflexion of the ribbon shaped electron beam, possessing the ability to retain one of the ten stable positions, which the beam once has assumed. Utilizing this property the fast counting principle is based on forced displacement of the beam from one stable position into the adjacent one, by the action of a step-function waveform as a drive pulse. It is apparent that it is not possible to bring the beam exactly into the new stable position by the drive pulse only, so that stabilizing action is required. There is a limited region within which the beam may reach the stable position and time is required to complete it. Thus the main question which arises in investigating the possibilities of fast counting with the E1T is to determine the limits of permissible changes of the drive pulse at a given counting speed to ensure errorless counting.

The basic fast counting circuit for the E1T is shown in Fig. 1. The counting is performed in discontinuous discharging and after a sequence of ten pulses recharging of the capacitance C consisting of circuit stray capacitances referred to the point A. The deflexion plate voltage V_{D2} is lowered by drive pulses V_p and the beam is moved step-by-step through the corresponding stable positions '0' to '9'. The tenth pulse moves the beam towards the auxiliary anode to initiate the flyback circuit which returns the beam to '0'. The anode and the auxiliary anode currents as functions of the deflexion plate voltage V_{D2} measured on a practical tube are given in Fig. 2. The anode characteristics I_{a1} can be represented as a sine waveform with a period equal to the voltage difference V_s between two adjacent stable positions. A part of the characteristics between a stable and a metastable point with respect to the anode resistance line R_{a1} as x-axis is presented in dimensionless form in Fig. 3. Since the position of the beam is corresponding to the deflexion plate voltage V_{D2}, the term 'stable point' instead of 'stable position' will be used in further treatment. The point $v/V_s = 0$ in Fig. 3 is stable and the point $v/V_s = 1/2$ is metastable.

Before discussing the counting process, the behaviour

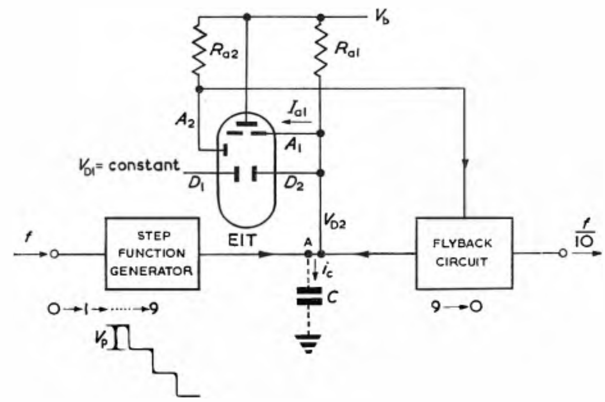


Fig. 1. Principle of a fast counting system using E1T tubes

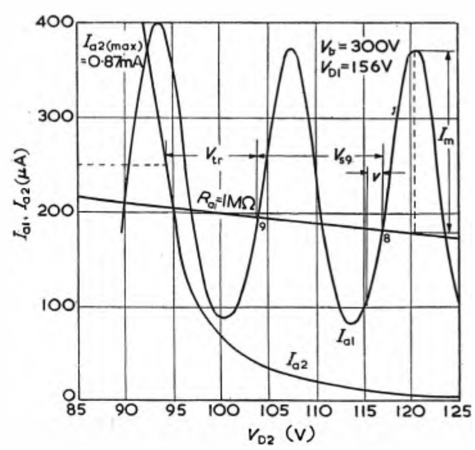


Fig. 2. Anode and auxiliary anode current against V_{D2} for an average tube
Stable points 8 and 9 only can be seen

of the beam when displaced from a stable point will be considered. At stable points the total anode current is supplied through the anode resistance R_{a1}. If the deflexion plate voltage V_{D2} is changed by v, a difference between the anode current and anode resistance current arises which is given by

$$i_{a1} = I_m \sin 2\pi\, v/V_s.$$

The current difference flows through the capacitance C and decreases v returning the beam to the stable point according to the equation

$$-\, C\, dv/dt = I_m \sin 2\pi\, v/V_s. \qquad (1)$$

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It is of interest to determine the time needed for the beam to move from an unstable point 1 to the point 2 in the vicinity of the stable point, Fig. 3. Having separated variables, integration from point 1 to 2 gives

$$t/\tau = \ln \frac{\tan \pi\, v_1/V_s}{\tan \pi\, v_2/V_s} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

where

$$\tau = C/I_m \,.\, V_s/2\pi \quad \ldots\ldots\ldots\ldots\ldots (3)$$

Equation (2) shows that in all cases where $v_1/V_s < 1/2$, i.e. if the beam has not been displaced beyond the metastable point $v/V_s = 1/2$, the beam will return to the stable point $v/V_s = 0$. For v_1/V_s between 1/2 and 3/2 the beam will move toward the next stable point where $v/V_s = 1$. The curve t/τ in Fig. 3 is presented for $v_2/V_s = 1/20$. As it can be seen from Fig. 3 and (2) infinite time is needed theoretically to reach the stable point, but practically this point is reached in a comparatively short time.

Starting the counting process from the initial stable point marked '0' on the tube screen the beam will be brought by an inaccurate drive pulse V_p into the vicinity

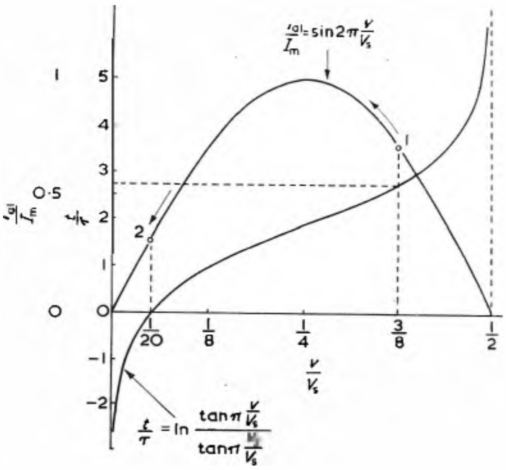


Fig. 3. Sine function representation of the E1T characteristics and the time required to reach a stable point

of the stable point '1'. The deviation from a stable point can be expressed as an angle. The angular deviation from the stable point '1' immediately after the first pulse is

$$\phi = 2\pi\,\frac{V_p - V_s}{V_s} = 2\pi\,(k-1),\; k = V_p/V_s \quad \ldots\ldots\ldots (4)$$

ϕ can be used to express the drive pulse height V_p with respect to the voltage difference V_s. According to equation (2) the deviation of ψ_1 at a time T after the first pulse is

$$\psi_1 = 2\arctan\,(e^{-T/\tau}\tan\phi/2) \quad \ldots\ldots\ldots\ldots (5)$$

The second pulse will bring the beam into the vicinity of the stable point '2' with an angular deviation $\psi_1 + \phi$. Assuming a periodical train of pulses relation $f = 1/(T+t_p)$ exists (f = repetition rate, T = time interval between two successive pulses, t_p = drive pulse duration). The angular deviation just before the arrival of the third pulse, by analogy with equation (5) will be

$$\psi_2 = 2\arctan\,[e^{-T/\tau}\tan 1/2\,(\psi_1 + \phi)]; \quad \ldots\ldots\ldots\ldots (6)$$

so that the angular deviation from the stable point '9' just before arrival of the tenth pulse can be expressed by

$$\psi_9 = 2\arctan\,[e^{-T/\tau}\tan 1/2\,(\psi_8 + \phi)]. \quad \ldots\ldots\ldots\ldots (7)$$

ψ_9 may be expressed as an explicit function of ϕ by successively substituting relations for ψ_8 to ψ_1 in equation (7). Function $\psi_9 + \phi$ representing the accumulated angular deviation at various counting speeds in terms of ϕ, i.e. percentage inaccuracy p of the drive pulse, has been plotted in Fig. 4. The parameter is T/τ, and numerical values of repetition rates f have been calculated for an average E1T tube with $I_m = 80\mu A$ (the smallest peak) and $V_s = 14{\cdot}5V$. The capacitor C was taken as 60pF. In a circuit like that shown in Fig. 6, stray capacitances could hardly be reduced below this value. So from equation (3) it follows that $\tau = 1{\cdot}7\mu sec$. The duration of the drive pulse is $t_p = 0{\cdot}2\mu sec$, which is a convenient value.

The conditions which must be fulfilled to ensure correct counting are

(a) The flyback must be always, and only, started by the tenth pulse.

(b) During the counting the beam must not be brought into any point outside the region between the two metastable points adjacent to the stable one corresponding to the last digit of the pulse number.

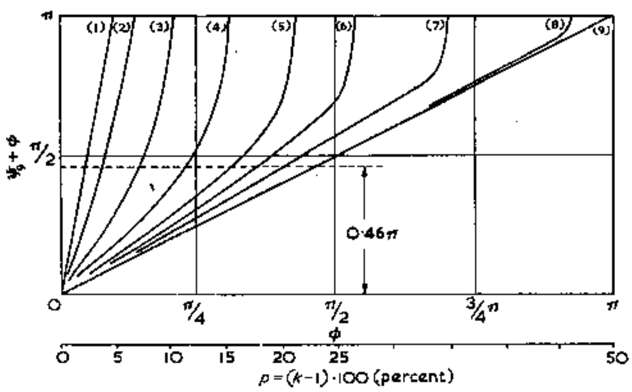


Fig. 4. Accumulation of the deviation in the counting process

Curve	f(Mc/s)	T/τ	Curve	f(Mc/s)	T/τ	Curve	f(Mc/s)	T/τ
(1)	$\to\infty$	$\to 0$	(4)	0·6	0·848	(7)	0·2	2·77
(2)	2	0·173	(5)	0·4	1·33	(8)	0·1	5·66
(3)	1	0·462	(6)	0·3	1·81	(9)	$\to 0$	$\to\infty$

Let V_{tr} be the voltage difference between the stable point '9' and the threshold where the flyback will be initiated (see Fig. 2). Then the boundaries the drive pulse has to be kept within are determined by the following relations.

for $V_{tr}/V_s > 1/2$:

$$\phi > 0 \qquad \psi_8 + \phi < \pi \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

$$\phi < 0 \qquad -(\psi_9 + \phi) < \pi - (V_{tr}/V_s - 1/2)2\pi \quad \ldots\ldots\ldots\ldots (9)$$

for $V_{tr}/V_s < 1/2$:

$$\phi > 0 \qquad \psi_8 + \phi < \pi - (1/2 - V_{tr}/V_s)\,2\pi \quad \ldots\ldots\ldots\ldots (10)$$

$$\phi < 0 \qquad -(\psi_9 + \phi) < \pi$$

It follows from equations (8), (9) and (10) that the widest limits are when

$$V_{tr}/V_s = 1/2 \quad \ldots\ldots\ldots\ldots (11)$$

Since it is impossible to solve equation (7) with respect to ϕ analytically, the solution has been obtained graphically by applying the conditions (8) and (9) on Fig. 4. The difference between $\psi_8 + \phi$ and $\psi_9 + \phi$ is negligible. The permissible inaccuracy, i.e. tolerances of the drive pulse as a function of the repetition rate have been plotted in Fig. 5. Curves 1 — 1′ are for an idealized tube satisfying the condition (11). The curve 2 is an example for $V_{tr}/V_s = 0{\cdot}77$, for which $\psi_9 + \phi \leqslant 0{\cdot}46\,\pi$ as shown in Fig. 4. The experi-

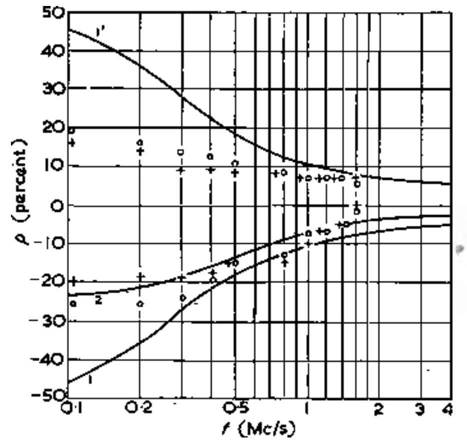


Fig. 5. Drive pulse tolerance as a function of the repetition rate

Curves 1 and 1' are calculated for an idealized tube with $V_{tr}/V_s = 1/2$, while curve 2 is for $V_{tr}/V_s = 0{\cdot}77$
+ and 0 are experimental results for the two tubes with $V_{t1}/V_{s9} = 0{\cdot}9$ and $V_{tr}/V_{s9} = 0{\cdot}77$ respectively

mental results for the two E1T tubes used in the circuit given in Fig. 6 are shown in Fig. 5.

There are certain differences between the theoretical and experimental results given in Fig. 5 which have to be considered. The foregoing analysis was carried out under the assumption of equally spaced stable points which is in fact not true. The voltage difference V_s decreases, irregularly with the ordinal number of the stable point. It was found on a number of tubes that the ratio of the first and the ninth difference is usually situated in the range

$$1{\cdot}14 < V_{s1}/V_{s9} < 1{\cdot}6 \quad \ldots\ldots\ldots\ (12)$$

The suggestion to decrease the drive pulse according to the decrease of the distance between stable points (neglecting irregularities) would lead to wider tolerances if the condition (11) applied to a real tube could be satisfied

$$V_{tr}/V_{s9} = 1/2. \quad \ldots\ldots\ldots\ (13)$$

However, reliable and jitterless flyback initiation can be obtained only on the steep part of the I_{a2} characteristics in Fig. 2, and on the tubes examined it was found that

$$0{\cdot}65 < V_{tr}/V_{s9} < 0{\cdot}95.$$

Considering the large difference between $2V_{tr}$ and V_{s9} in more detail it follows that the above mentioned suggestion would not lead to a remarkable improvement. As it can be expected tubes with a ratio V_{tr}/V_{s9} nearer to one half allow wider tolerances (Fig. 5).

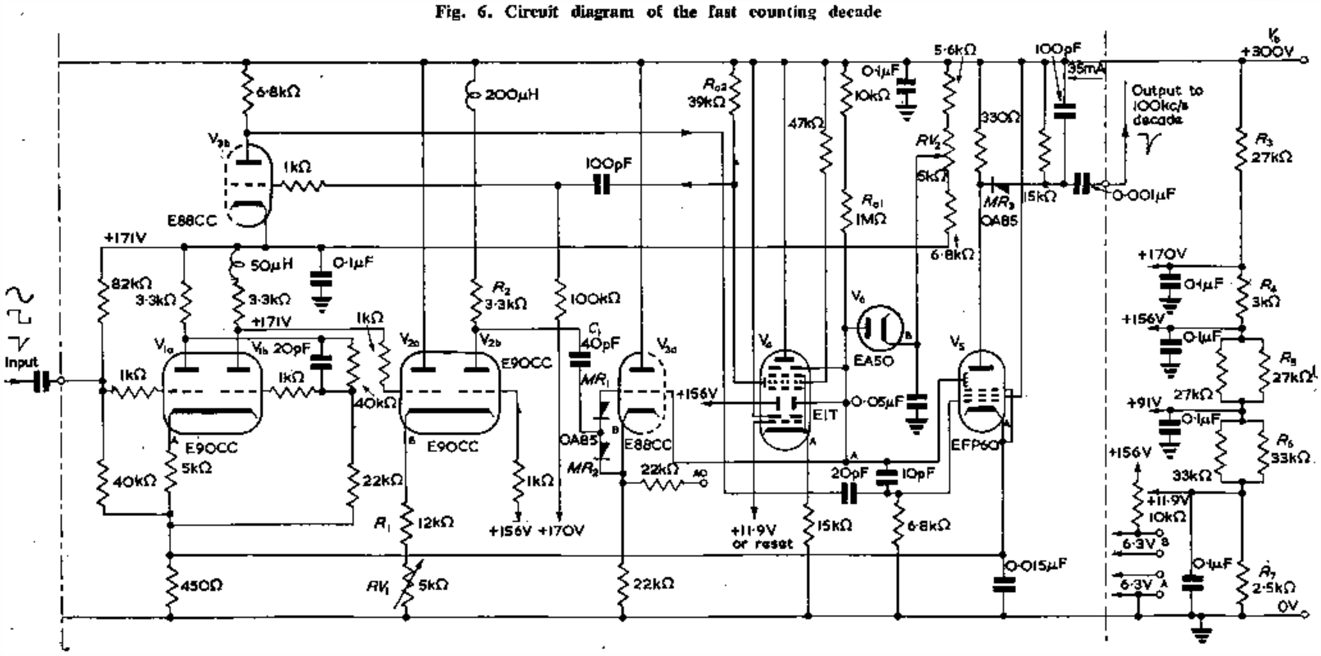
From the preceding discussion and Fig. 5 it can be seen that it is possible to use E1T tubes in fast counting circuits up to about one million periodical pulses per second under the condition to keep the drive pulse within $p \simeq \pm 8$ per cent. At higher counting speeds difficulties arise because of narrower tolerances and shorter transients. The more complicated circuits involve greater stray capacitances which, in turn, because of condition (3) cause a further decrease of tolerances.

A Fast Counting Decade

In designing the fast counting decade it was aimed at staying within tolerances determined in the preceding section making minimum requirements on the stability of components and of the power supply. The circuit diagram of the decade according to the basic principle in Fig. 1 is shown in Fig. 6.

The step function generator is designed in such a way to make the drive pulse entirely independent on the input pulse shape and amplitude. The input pulse is formed in a Schmitt-trigger, V_1. The proper amplitude is obtained by a cathode coupled clipper V_2. The step function waveform is produced by the linear diode pump consisting of diodes MR_1 and MR_2, a cathode-follower V_{3a}, the capacitor C_1 and the stray capacitances C referred to the point A. The step function height V_p is determined only by the resistors RV_1, R_1, R_2 and capacitances C_1 and C. Time and temperature instabilities of the components RV_1, R_1, R_2 and C_1 must not exceed 1 per cent. The stray capacitance C, consisting of many elements, is not influenced largely by small changes of one of them. The height of V_p is set to the required value with RV_1. The auxiliary potentials which are standardized in conventional equipment using E1T tubes are obtained with the divider R_3 to R_7. This divider is unloaded except for the deflexion plate D_1(+ 156V) which draws a current up to 15μA due

Fig. 6. Circuit diagram of the fast counting decade



to non-ideal focusing of the electron beam. The constant level + 171V is obtained by the triode section V_{3b}, functioning as a d.c. cathode-follower. The same section is also used as a trigger amplifier for the monostable circuit with an EFP60 (V_5) which performs the flyback.

The secondary-emission valve EFP60 is connected in a commonly used circuit with regenerative feedback from the dynode to the first grid. Without the diode V_6 the peak of the dynode voltage pulse would be determined approximately by the anode potential because the secondary emission current becomes nearly zero when the dynode potential reaches the anode. The dynode zero current potential is not exactly defined by the anode potential, it differs from tube to tube and is subject to ageing. Considering this, it is not recommended to use the anode potential alone in setting the upper limit of V_{D2} and the diode V_6 was inserted to define precisely the end of the regenerative process. The diode V_6 opens at the end of the process when the dynode current is decreasing, so that the peak diode current does not exceed 15mA. Exact returning of the beam to '0' is set by means of RV_2. A negative 30V output pulse is derived from the anode of the EFP60. Pulse lengthening with diode MR_3 is used to match the pulse to the input of a decade stage[3] for 10^5 pulses per second. The time-constant of the trailing edge is about 1·5μsec.

A marginal test has been made to prove the reliability of the decade. The use of only one d.c. supply voltage makes the circuit entirely independent of d.c. supply voltage variations. No error has been observed at a counting speed of 1Mc/s due to simultaneous variations of the anode and filament voltage in the range.

$$120\text{V} < V_b < 400\text{V};\ 4{\cdot}5\text{V} < V_f < 7{\cdot}5\text{V}.$$

The valves and all components are operated within manufacturers' ratings. Special care has been taken in the operating conditions of the EFP60. The peak currents and dissipations do not exceed one fifth of rated values. The flyback precision is not affected by changes in the characteristics of the EFP60. There is no valve in the circuit which has to be selected, and only simple adjustments of RV_1 and RV_2 have to be made when an E1T is replaced.

The input pulse may be of any convenient shape with a negative part exceeding 20V. Random pulses can be counted with a resolution < 0·5μsec. The top counting speed 1·6 to 1·8Mc/s for periodical pulses has been measured as is indicated in Fig. 5. The staircase output, i.e. the counting result as an analogous quantity can, if desired, be taken from the point AO.

Acknowledgments

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Direct Reading for Data Processing

De La Rue Bull Machines Ltd have announced a system of direct reading for data processing.

The basic principle of the De La Rue Bull direct figure reading system is that the figures are produced in a magnetizable ink which can be printed by any of the accepted printing processes. The figures can also be imprinted by movable type bars—such as those used in adding machines and typewriters—and from embossed plastic or metal printing plates.

Throughout the development of this new direct reading system it was regarded as essential that the figures should be normal in form and instantly readable by the human eye. This feature has been obtained by printing standard figures in the form of a series of vertical lines.

These lines are spaced either 0·2mm or 0·4mm apart and therefore the printing of the figure appears virtually continuous to the human eye. The basis of the reading system is that the magnetizable ink appearing in the vertical lines is simply an alerting signal to the interpreting equipment and the actual interpretation of the printed figures depends on the relative dispositions of the wide (0·4mm) and the narrow (0·2mm) spaces.

Examination of an enlarged printed figure shows that the vertical lines are themselves of different thicknesses. This is purely for aesthetic reasons and has no influence on the reading mechanism which is concerned only with spacings between the lines.

The normal numerical code is obtained by using seven vertical lines giving six spaces, of which two are wide and four are narrow. The spaces between different figures are sufficiently great for the reading apparatus to turn itself off and a quick-acting delay mechanism allows six spaces to have been read before the interpreted figure code is transmitted for subsequent data processing or for direct recording.

The same principle can readily be applied to the reading of alphabetical information.

The advantage of this system as compared to the interpretation of thin-line outlines is that the printed figures are entirely normal. The advantages as compared to the reading of magnetic profiles are that far less accuracy of printing is required—because there is an enormous tolerance on the amount of magnetic material appearing in each vertical line, since this is acting simply as an alerting signal—and that for profile reading a fairly uniform response must be obtained over the full width of the profile. Furthermore, figures encoded in vertical lines can still be read if the documents are significantly crumpled, damaged or perforated—which factors seriously upset profile reading.

The advantages as compared to systems depending on comparing a printed figure or letter with standard figures or letters already programmed into a memory are, first of all speed,

The visual impression created by the typeface (considerably enlarged)

secondly much greater tolerance to damage and misalignment, and thirdly much greater simplicity in electronic circuits since the signal obtained from the vertical line coding can be transformed directly into a binary signal.

Features of the De La Rue Bull direct reading system were demonstrated to the Conference of European Bankers (held in Paris in October last year). Pre-printed cheques were sorted according to various sorting codes and also according to account numbers subsequently printed on the documents. An imprinter developed from a standard adding machine was also demonstrated, by which the amounts for which the cheques were drawn were imprinted in magnetizable ink enabling accounting and clearing house reconciliation to be achieved directly. The imprinting machine has security devices which prevent cheques passing forward to sorting and accounting until the amounts are confirmed as correctly imprinted on them. Through a vast number of tests, no single misread of sorting code, account number or amount was recorded, although a number of cheques were deliberately defaced and damaged.

Economy in the Series Stabilizer

By D. J. Collins*, B.Sc., A.M.I.E.E., and J. R. Pearce*, B.Sc., A.M.I.E.E.

In this article it is pointed out that in order to deal with adverse conditions the series stabilizer is uneconomic in its demands on the series elements. If the load current is essentially constant the series element can be shunted by a resistor. The basic circuit is presented of a practical solution to the problem of a variable load current using a resistance shunted buffer section.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

IN recent years, designers of regulated d.c. power supplies have tended to favour the series stabilizer. This circuit, when enclosed in a d.c. feedback loop, generally gives a performance which is superior to that of the shunt stabilizer, and is applicable to a wider range of duties[1].

A possible disadvantage of the series configuration, is lack of economy in the basic control element. Under certain conditions, this criticism may be justified, and it can be rightly pointed out that, not only is the series stabilizer uneconomic in input voltage requirement, but it may also have to be capable of handling a wider range of power and will, therefore, be the more expensive of the two systems.

The various arguments apply to either valves or transistors, and both have a particularly difficult task when there is a wide variation in the unstabilized supply voltage. This article describes a system which has been devised in order to provide an economical series stabilizer for transistor power supplies.

Shunting of Series Control Section

Fig. 1 shows the basic arrangement of a transistor series stabilizer. Only one transistor X_1 is shown, but, in practice, a more complicated arrangement of several transistors may be necessary. The unstabilized supply can consist of a conventional transformer, rectifier and smoothing system. To give 30V at the load, the minimum unstabilized voltage is usually not less than about 32V, this allowing for low mains voltage and transformer errors, etc. When the mains voltage is high, and again allowing for other errors, the input to the stabilizer might rise to 40V. Assuming, for simplicity, that the load current is a constant 10A, this gives a maximum power dissipation in the control transistor of 100W.

The maximum power which can be dissipated in a transistor varies with many factors, such as junction size, method of mounting, the form of heat sink, ambient temperature, etc. A typical maximum for transistors available at the moment might be 10W each, thus needing ten transistors in parallel.

Under the lightest conditions, however, the total dissipation is 20W—a mere 2W per transistor. In order to reduce the number of transistors employed, the designer may resort to large or complex heat sinks and blowers; thus, replacing a high transistor cost with an equally expensive heat sink.

If the load current is essentially constant, the transistor may be shunted by the resistor R in Fig. 2; the value being selected so that, under conditions of maximum voltage, most of the load current flows through R. In the example chosen, with a maximum of 10V across the transistor, a resistor of 1·25Ω will pass 8A, leaving 2A for the transistor. This condition does not, however, give the maximum power in the transistor, since, as the unstabilized voltage is reduced, more current must pass through X_1. It can be shown that the maximum transistor power will be $I^2R/4$ where I is the load current.

Hence, in this example, the series section transistors would have to deal with only 31·2W under the worst conditions. Thus, the transistor complement can be reduced to four from the original ten.

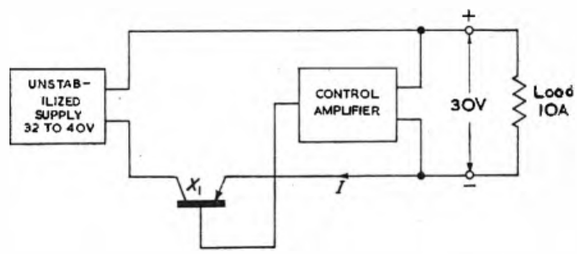

Fig. 1. Basic arrangement of transistor series stabilizer

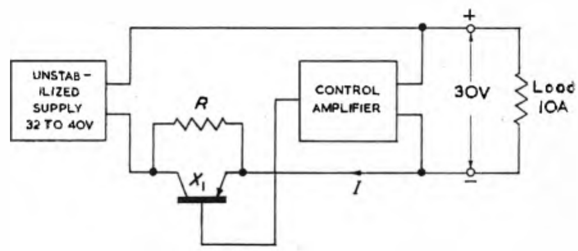

Fig. 2. Transistor shunted by resistor

With this arrangement, the load current cannot be reduced much below 10A, for in the limit—i.e. when the load is 8A—the series element has become cut off and the stabilizer ceases to function. If it is necessary to reduce the load current to zero, a shunt resistor is not the solution to the power dissipation problem.

Dual Series Section

When the load current of a transistor power supply is to be reduced to zero, or even varied over a wide range, a dual series section may be used to effect economy in the number of transistors. As with the arrangement described above, the excess power is dissipated in a resistor.

The basic circuit[2] is shown in Fig. 3, where it will be seen that the control section X_1 now has in series with it a buffer section X_2 shunted by R. The base of the buffer transistor X_2 is taken to a stable potential E with respect to the negative output, and the sum of the voltage across X_1 (V_1) and the bias on X_2 (V_{eb}) must always be equal to E. Since E is constant, if V_1 rises, V_{eb} must fall, thus tending to cut off X_2. This is the action upon which the operation of the circuit depends.

Consider first the mains input to be low. The unstabilized supply will be 32V, which gives the minimum required across X_1. In this case, the voltage drop across the buffer section would ideally be zero, but in practice, an allow-

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ance is made for a small working voltage across X_2. Since V_1 is small, V_{eb} will ensure that X_2 is conducting heavily, and this transistor may be visualized as shunting the resistor. This condition is maintained over the full range of load current, and the power in the whole section has a maximum of 20W, nearly all being dissipated in X_1. Hence, as long as the unstabilized supply is low, the buffer section is not fully utilized.

Consider now that the mains input voltage is high (unstabilized voltage 40V) and the load current small. The increased voltage across X_1 now ensures that X_2 is cut off. All the current flows through R, and the voltage across X_1

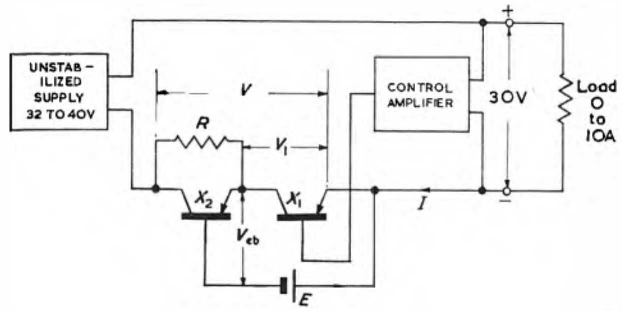


Fig. 3. Basic arrangement of dual series section stabilizer

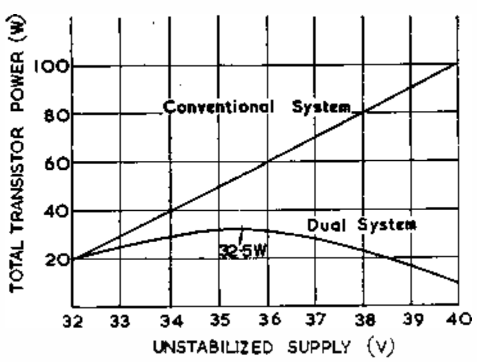


Fig. 4. Comparison of dissipation between conventional stabilizer and dual series section system

falls as the load current is increased. Thus, the power dissipated in X_1 is not so great as in the original example, a substantial amount of power being transferred to the resistor. It is found that the power in X_1 rises to a maximum at very roughly half full load.

The appendix details the basic calculations, and Fig. 4 shows a comparison of dissipation between a conventional stabilizer and a dual section system. The enormous advantage is readily apparent, and it can be seen that the number of transistors for a given application can be considerably reduced.

Practical Considerations

It will be appreciated that an ideal case has been dealt with. Some of the points which must be considered in a practical design are:

(1) Series elements dealing with large currents and large powers are invariably composed of transistors connected in parallel, and in order to obtain good current gain the elements are usually compound emitter-followers. Allowance must be made for compounding and for any emitter resistance used for ensuring load sharing.

(2) Variations (such as ripple) in a nominal fixed supply voltage must enter into design considerations, and allowance must be made for the regulation of the unstabilized supply.

(3) Although the transistors of both sections may be mounted on a common heat sink, it should be noted that the collectors of the two sections are not common. It is interesting to note that each has a condition of maximum dissipation, although they do not occur at the same operational point.

(4) A possible refinement is that the reference potential E may be chosen in order to allow X_2 to dissipate some power at full load. This reduces the maximum power in X_1, and may enable full utilization to be made of an integral number of transistors, instead of a possible duplication of calculated fractional transistors in each section.

Power supplies using a dual series section are commercially available, a typical unit supplying an output of up to 10A at a maximum of 30V.

Acknowledgments

The authors wish to thank the Directors of Solartron Research and Development Ltd. for permission to publish this article.

Appendix

Consider the circuit of Fig. 3. At any operating condition the current in X_2 is:

$$I = \frac{V - V_1}{R}$$

The power in X_2, $P_2 = I(V - V_1) - \frac{(V - V_1)^2}{R}$

and in X_1, $P_1 = V_1 I$

Total transistor dissipation, $P = P_1 + P_2$

$$\therefore\ P = V_1 I + I(V - V_1) - \frac{(V - V_1)^2}{R}$$

$$= VI - \frac{(V - V_1)^2}{R} \quad \ldots\ldots\ldots\ldots (1)$$

The total dissipation, for a fixed value of V, is obviously proportional to I, and will be maximum at full load current. Consider the value of V, which, at a set value of I, gives maximum power:

$$dP/dV = I - (1/R)(2V - 2V_1)$$

and P is maximum when:

$$I = (2/R)(V - V_1)$$

or when:

$$V = (IR/2) + V_1 \quad \ldots\ldots\ldots\ldots (2)$$

Substituting equation (2) in equation (1):

$$P_{max} = [(IR/2) + V_1]\, I - (1/R)(IR/2)^2$$
$$= (I^2R/2) + V_1 I - (I^2R/4)$$
$$= (I^2R/4) + V_1 I$$

In Fig. 4, a comparison is made between a conventional series stabilizer and a dual system. The following values have been taken:

Load current (I)	10A
Buffer resistor (R)	0·9Ω
V_1	1V
Unstabilized supply	32 to 40V.

REFERENCES

1. Collins, D. J., Smith, J. E. Regulated Power Supplies. *Electronic Engng* 31, 222 (1959).
2. Patent Provisional Specification No. 33106 : 58.

Some Performance Parameters of Silicon Junction Power Rectifiers

By D. R. Coleman*, B.Sc.(Eng.), A.M.I.E.E.

The electrical performance of silicon power rectifiers—seen in terms of the power output of a conversion connexion—is related to their characteristics, and shown to be dependent upon the junction temperature. Temperature control is discussed from the thermal resistance standpoint, and a method is given for the calculation of power dissipation and connexion ratings for silicon rectifiers. Mention is made of some considerations necessary for satisfactory transient performance.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

THE electrical characteristics of silicon junction rectifiers in the forward and reverse directions are well know[1,2,3,4]. The performance of such rectifiers in power conversion connexions is directly related to those characteristics, and in this sense the variations of characteristic with junction temperature are of critical importance.

In general terms, the forward characteristics suggest an improved performance with increase of junction temperature, since a given current in the working region produces a decreasing voltage drop with increasing temperatures. On the other hand the reverse characteristics indicate a degradation in performance, in that at a fixed working voltage the current increases with rise of junction temperature.

A typical temperature coefficient of forward voltage drop would be $-0{\cdot}1$ per cent/°C, showing the effect on performance to be slight. The reverse current, however, may vary considerably. Although the saturation current can increase by as much as[2,5] 15 per cent/°C the total cell current variation with temperature is usually less, by virtue of the less temperature sensitive leakage currents present. The overall effect on performance is not necessarily so severe as these variations might suggest, since the effective rectification ratio of a silicon rectifier is usually greater than 10^5

JUNCTION LOSSES

It is obvious that operation of a junction rectifier in a conversion connexion will produce electrical losses, which may be assessed by the methods described later. The consequent heat generation from those losses means that the rectifier junction will always operate at a temperature greater than its surroundings.

JUNCTION TEMPERATURE

Apart from its influence on the electrical characteristics of the rectifier, operation at an extremely high junction temperature may produce undesirable effects in the constituents of the rectifier assembly. Improvements in processing and encapsulation have been shown to give rectifiers capable of operation at 150°C junction temperature for very long periods; there is reason to believe, however, that a shorter life results from continuous operation at, for example, 200°C junction temperature.

For some applications a short life is acceptable, and in these cases an uprating of operating temperature can give useful increases in the electrical ratings obtained. For a general purpose rectifier a continuous operating temperature consistent with long life has the additional advantage of permitting a greater momentary temperature to be sustained if necessary, under transient conditions of forward voltage or reverse current surges.

Thermal Considerations

JUNCTION LOSSES AND TEMPERATURE DIFFERENCE

Each watt of junction dissipation causes a heat flow rate of approximately 0·24g cal/sec from the junction, that heat flow producing a temperature difference across any thermal resistance in its path, such that under steady state conditions,

$$T = W.\,\theta \qquad {}^\circ\mathrm{C},$$

where W is the heat flow rate (W) and θ is a thermal resistance (°C/W). The operating junction temperature is therefore greater than the cooling medium temperature, since a thermal resistance cannot be avoided in the extraction of the heat produced.

TYPES OF RECTIFIER

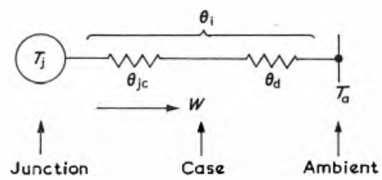
For thermal considerations of rectifier performance, there are two separate types of rectifier:

(a) Those intended for 'convection cooling'.

(b) Those intended for 'conduction cooling'.

The first class includes wire-ended and integral-fin rectifiers, where the rectifier dissipating properties in a given medium are unalterable. The second class describes the stud-ended and similar rectifiers, whose dissipating properties may be varied by the attachment of different fins or heat sinks. The differences between the classes may be readily understood by using electrical analogues of their thermal circuits.

Fig. 1. Heat generation and thermal resistance path of a wire-ended rectifier
θ_{jc} — *internal thermal resistance*
θ_d — *device effective thermal resistance*
θ_i — *'intrinsic' thermal resistance*



WIRE ENDED RECTIFIERS

With a wire-ended or integral-fin rectifier, continuous heat flow is from the junction to the case, and thence by combined convection and radiation, and conduction through the terminations, to the cooling medium (Fig. 1).

The internal temperature gradient is then

$$T_{jc} = W.\,\theta_{ja}$$

and will not be affected greatly by changes in the ambient environment.

The external temperature gradient is

$$T_{ca} = W.\,\theta_d$$

and will be significantly altered by the effectiveness of the cooling medium. The value of θ_d for an air-cooled rectifier of 1W rated dissipation is reduced typically to one sixth by immersion in oil, to give one illustration.

The overall thermal resistance of the rectifier is the sum of the internal and external components—

$$\theta_{ja} = \theta_{jc} + \theta_d \qquad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

and is constant for a given type of cooling, for example convection air cooling, or a forced air cooling rate. It may therefore be considered as an intrinsic property of the rectifier in a stated cooling medium, in which event the

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junction temperature rise will be directly related to the junction losses:

$$T_{ja} = W.\ \theta_i \quad \ldots\ldots\ldots\ldots\ldots \ (2)$$

where θ_i is the 'intrinsic' thermal resistance, equal to θ_{ja} (equation (1)).

The intrinsic thermal resistance is directly measurable, by continuous dissipation of a known junction loss and the detection of the steady junction temperature by reference to its temperature dependent forward characteristic[6]. Temperature sensing by use of the reverse characteristic is also possible[7].

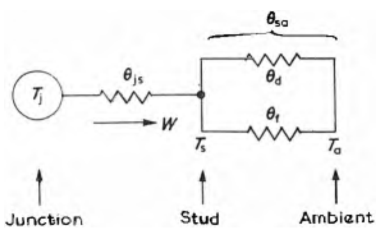


Fig. 2. Heat generation and thermal resistance paths of a stud-mounted rectifier

θ_{js} — *internal thermal resistance*

θ_d — *device effective thermal resistance*

θ_f — *heat sink thermal resistance*

Stud Mounted Rectifiers

Under steady state operating conditions, the heat generated at the junction in a stud-ended rectifier may be considered as flowing through an internal thermal resistance, and then an external thermal resistance to the cooling medium.

The internal thermal resistance from the junction to the stud may be assumed constant. The external thermal resistance is conveniently thought of as a parallel circuit of a constant thermal resistance of the rectifier alone, shunted by a thermal resistance due to the heat sink. The value of the heat sink thermal resistance will be variable (Fig. 2).

Without an added heat sink the considerations of the previous section apply, and the overall thermal resistance (the rectifier 'intrinsic' thermal resistance) is

$$\theta_i = \theta_{js} + \theta_d \quad \ldots\ldots\ldots\ldots\ldots \ (3)$$

in which case θ_d varies according to the cooling medium (or ambient) conditions, but θ_{js} remains virtually constant For a given maximum junction temperature $T_{j(max)}$, the greatest dissipation possible without a heat sink will be

$$W_{i(max)} = \frac{T_{j(max)} - T_a}{\theta_i} \ldots\ldots\ldots\ldots\ldots \ (4)$$

and is called the maximum 'intrinsic' dissipation at T_a ambient temperature.

If the rectifier is attached to a heat sink or fin of thermal resistance θ_f, the thermal resistance from the stud to the cooling medium will be

$$\theta_{sa} = \frac{\theta_f\,.\,\theta_d}{\theta_f + \theta_d} \ldots\ldots\ldots\ldots\ldots\ldots \ (5)$$

Thus the total thermal resistance to heat flow from the junction becomes

$$\theta_{ja} = \theta_{js} + \theta_{sa} \ldots\ldots\ldots\ldots\ldots\ldots \ (6)$$

A given fin of thermal resistance θ_f may have a different effect with different rectifiers (due to variations of θ_d between types of rectifier) and will have a considerably different effect in different cooling media since both θ_f and θ_d are dependent upon the cooling system used.

In the case of a zero thermal resistance (infinite heat sink) being attached to the stud—i.e. $\theta_f = 0$—the external thermal resistance becomes zero (from equation (5)), but the temperature drop from the junction to the stud is unaltered at

$$T_{js} = W.\ \theta_{js} \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (7)$$

and the stud temperature is $T_s = T_j - W.\ \theta_{js}$.

If the rectifier is now operated at its maximum dissipation W_{max}, the maximum junction temperature is produced at a stud temperature of

$$T_{s(max)} = T_{j(max)} - W_{max}\,.\,\theta_{js} \ \ldots\ldots \ (8)$$

W_{max} is thus the 'ultimate' dissipation obtainable at stud temperatures up to $T_{s(max)}$.

The performance of a stud ended rectifier can therefore be determined in terms of a given stud temperature, or of a given heat sink thermal resistance and cooling medium.

Heat Sinks and Cooling Methods

Separately attached heat sinks for silicon power rectifiers intended for natural air convection, forced air draught, and liquid coolant immersion may be of the single or multi-fin types; much ingenuity has been shown in the development of special forms and sections to permit ready interlocking and formation of bus connexions.

Their thermal resistance is measurable, as indicated in the previous section, or may be calculated by a variety of methods[3,8,9]. The θ_f values can vary from, for example, 14°C/W for a simple 3in × 3in fin of 1/16in copper to 0·4°C/W for a 7in length of a multi-finned 6in × 4in extruded aluminium section, suitable for a 1W and a 150W dissipation rectifier respectively, under natural air cooling conditions.

Higher dissipations, or the need to reduce the temperature rise, demand more effective cooling methods such as heat extraction by a circulating liquid coolant system. A heat exchanger of this form, for example, can dissipate 1·2kW with an oil temperature rise of 5°C, by an oil flow of 2gal/min through a section of area 1·75in², over a 5ft length of hollow aluminium bar.

Steady State Conditions

Determination of Forward Losses

The forward characteristic of a silicon rectifier may reasonably be idealized, as shown in Fig. 3, to give the relationship

$$V = a + bI$$

where a and b are constants.

Constant a is termed the 'threshold voltage' (volts) and the constant b is the 'slope resistance' (ohms). Practical values are found by extrapolating the line joining the voltage values at $T_{j(max)}$ corresponding to I_{mn} and I_{pk}, where

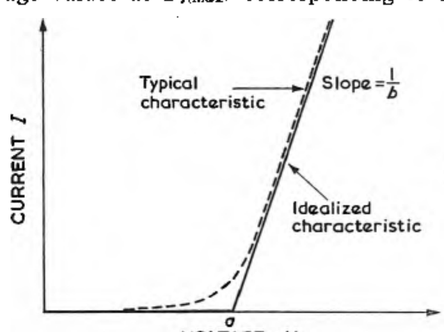


Fig. 3. Electrical characteristics in the forward direction

I_{mn} is the mean current and I_{pk} is the peak current in a rectifier connexion. The slope of that line is equal to $1/b$, and its intercept with the V axis gives the threshold voltage.

When conducting in the forward direction, the instantaneous voltage drop will be

$$v = a + bi$$

where i is the instantaneous current.

and the instantaneous power dissipation is given by

$$vi = ai + bi^2 \text{ watts} \quad \ldots\ldots\ldots \ (9)$$

With a sinusoidal voltage supply and a resistive load, the forward current through the rectifier may be assumed to be a part-sinusoid symmetrical about a peak value I_{pk}, and flowing for a conduction period 2θ radians, where the value of 2θ depends upon the rectifier connexion. For example, $2\theta = \pi$ radians in a single phase connexion, or $2\theta = \pi/3$ radians in a three-phase push-pull connexion.

Equation (9) may then be rewritten

$$vi = a\, I_{pk} \cos\theta + b I_{pk}^2 \cos^2\theta$$

and the mean power dissipation over the conduction period derived as

$$(vi)_{mn} = \frac{aI_{pk}}{2\theta}\int_{-\theta}^{+\theta} \cos\theta\,.\,d\theta + \frac{bI_{pk}^2}{2\theta}\int_{-\theta}^{+\theta} \cos^2\theta\,.\,d\theta \quad \ldots\ldots (10)$$

Since there is a fixed relationship (with a given rectifier connexion) between the peak current I_{pk} and the mean current I_{mn} per rectifier, and because the conduction period is repeated at supply frequency, the mean power dissipation per cycle $(VI)_{mn}$ may be expressed in terms of a and b. $(VI)_{mn}$ is in fact the forward loss per rectifier, or W_f as shown in Table 1.

The same principle of solving the integral $vi\,.\,dt$ to determine the total forward loss per cycle may be employed for other current waveforms. For example, square wave conduction in the bridge connexions may be compared with the part-sinusoid condition, as shown in Table 2.

Determination of Reverse Losses

The total loss in the reverse direction is found from $\int vi\,.\,dt$, integrated over the complete cycle. The voltage waveform is largely dependent on the rectifier connexion being used, and the reverse current varies according to the particular rectifier characteristic. It is therefore difficult to generalize, but since the reverse losses in silicon rectifiers form a relatively small proportion of the total dissipation, it is usually sufficient to approximate their value and consider this as a constant in normal use.

If the reverse voltage is taken to be at its crest value for a duration appropriate to the connexion (e.g. for 0·67 of the cycle in a three-phase bridge) and the reverse current is of the maximum value at the maximum rated junction temperature, the reverse loss thus found,

i.e. $W_r = 2/3\ V_{cw}\ I_r$ watts

where V_{cw} = rated crest working voltage

is the highest value likely in practice, and may be taken as constant without detracting greatly from the calculated performance of the rectifier under other circumstances.

Thermal Instability

It must be remembered, however, that the rectifier characteristics are temperature dependent. The forward characteristic of a silicon pn junction has a negative temperature coefficient; at a given forward current the forward losses decrease with increase of junction temperature. This tends to a temperature stabilizing situation, but the coefficient is relatively small.

The reverse current, however, increases with junction temperature at voltages in the working region. The reverse losses due to these voltages and currents may be expressed in the form[10]:

$$W_r = W_a \exp(cT) \quad \ldots\ldots\ldots\ldots\ldots (11)$$

where W_r = loss at junction temperature of T_j°C
W_a = loss at junction temperature of T_a°C
c = constant of temperature variation of I_r
$T = (T_j - T_a)$ (°C)

Now the total power loss

$$W = W_r + W_f \quad \ldots\ldots\ldots\ldots\ldots (12)$$

and the heat flow rate

$$W_d = (T/\theta) \quad \ldots\ldots\ldots\ldots\ldots (13)$$

where W_d = power dissipation to ambient
θ = thermal resistance from junction to ambient (°C/W).

Substituting equation (11) in equation (12)

$$W = W_a \exp(cT) + W_f \quad \ldots\ldots\ldots\ldots (14)$$

Equation (14) represents the power causing heat generation, in which W_f may be assumed constant.

Equation (13) represents the power dissipation properties of the rectifier.

Thermal instability will occur if the rate of generation of heat exceeds the rate of dissipating of heat; conversely stability is assured if the dissipation rate exceeds the generation rate. If the two rates are equal, the maximum possible heat flow for thermal stability is present.

This situation is found by equating the derivatives of equations (13) and (14),

$$\text{i.e. } 1/\theta = cW_a \exp(cT)$$

or

$$\exp(cT) = (1/cW_a\theta) \quad \ldots\ldots\ldots\ldots\ldots (15)$$

Inserting equation (15) in equation (14),

$$W_{max} = W_a\,.\,(1/cW_a\theta) + W_f$$

or since W_f is assumed constant,

$$W_{r(max)} = 1/(c\,.\,\theta) \quad \ldots\ldots\ldots\ldots (16)$$

The factor c is deduced from the rectifier reverse characteristics; its maximum theoretical value for the true saturation current is approximately 0·115, any surface leakage currents effectively reducing this value. The factor θ can be reduced in the limiting case to θ_{js} for a stud mounted rectifier; the maximum on the other hand will be $\theta_i (= \theta_{js} + \theta_d)$ with this type of rectifier, or $\theta_i (= \theta_{ja})$ for a wire ended design.

The theoretically maximum reverse loss for thermal stability will thus be (in the worst operating condition)

$$W_{r(max)} = 1/(c\,.\,\theta_i)$$
$$= (8{\cdot}7/\theta_i) \text{ watts} \quad \ldots\ldots\ldots (17)$$

This clearly may be exceeded for more efficiently cooled rectifiers, and may also be increased in the practical case by the existence of surface leakage currents.

Division of Losses

The relation between junction temperature and junction dissipation in a rectifier under given cooling conditions, and the method by which the losses caused by forward conduction and reverse blocking may be evaluated have already been discussed.

In order to assess the possible performance of a rectifier

TABLE 1

RECTIFIER CONNEXION	FORWARD LOSS PER RECTIFIER W_f (watts): PEAK CURRENT I_{pk}	RECTIFIER MEAN CURRENT I_{mn}
Single-Phase		
Half wave, Push-pull, Bridge	$0{\cdot}32aI_{pk}+0{\cdot}25bI_{pk}^2$	$aI_{mn}+2{\cdot}46bI_{mn}^2$
Three-Phase		
Half wave	$0{\cdot}28aI_{pk}+0.24bI_{pk}^2$	$aI_{mn}+3{\cdot}10bI_{mn}^2$
Push-pull	$0{\cdot}16aI_{pk}+0{\cdot}15bI_{pk}^2$	$aI_{mn}+6{\cdot}01bI_{mn}^2$
Bridge	$0{\cdot}32aI_{pk}+0{\cdot}30bI_{pk}^2$	$aI_{mn}+3{\cdot}00bI_{mn}^2$

TABLE 2

	FORWARD LOSS PER RECTIFIER W_f: SQUARE WAVE	PART-SINUSOID
Single-phase bridge		
I_{pk}	$0{\cdot}5aI_{pk}+0{\cdot}5bI_{pk}^2$	$0{\cdot}32aI_{pk}+0{\cdot}25bI_{pk}^2$
I_{mn}	$aI_{mn}+2bI_{mn}^2$	$aI_{mn}+2{\cdot}46bI_{mn}^2$
Three-phase bridge		
I_{pk}	$0{\cdot}33aI_{pk}+0{\cdot}33bI_{pk}^2$	$0{\cdot}32aI_{pk}+0{\cdot}30bI_{pk}^2$
I_{mn}	$aI_{mn}+3bI_{mn}^2$	$aI_{mn}+3bI_{mn}^2$

when used in a conversion connexion, a division of the dissipation into forward and reverse losses must be made. The rectifier could perform equally well as a blocking rectifier (100 per cent reverse losses) or as a device conducting continuously in the forward direction (zero reverse losses). Intermediately, a high forward current and low reverse voltage could result in the same total dissipation as the converse, in a rectifier connexion.

Assignment of Ratings

For general purpose power rectifier applications, especially with a range of rectifiers of one case design, the Crest Working Voltage (c.w.v.) rating assigned to a given rectifier will be controlled either by the maximum reverse dissipation allowable for thermal stability, or by the 'breakdown' (turnover) characteristic of the junction in which region the rectifier should not normally be operated. If neither of these considerations is overriding it is probable that the application itself will impose a desirable limit on the reverse current, and consequently the reverse dissipation. In general the reverse loss should not normally exceed 10 per cent of the total dissipation.

The remainder can in principle be produced by the forward losses, equivalent therefore to a certain mean forward current rating in the particular connexion desired.

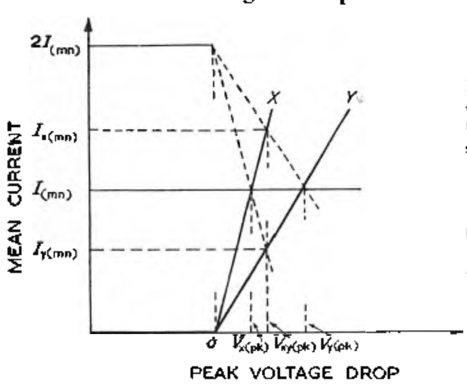
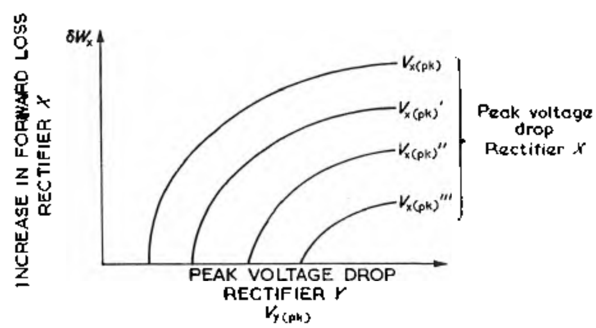


Fig. 4 (left). Parallel operation of rectifiers X and Y causes the total mean current $2I_{mn}$ to be shared as $I_{x(mn)}$ and $I_{y(mn)}$

Fig. 5 (right). Paralleling rectifier X with rectifier Y at double the rated current of each causes an increase δW_x in the dissipation of rectifier X



As Table 1 shows, the same forward dissipation may be caused by different rated currents in different connexions. The relationship between the input voltage equivalent to a required c.w.v. and the resulting output voltage is well known for the more common connexions; thus these ratings may be calculated for the connexion desired.

Assuming the reverse loss and the rated c.w.v. to remain constant, current ratings may be deduced in this way for all variants of the cooling method applied to the rectifier, the ambient conditions in which the rectifier operates, or the type of heat sink used.

Parallel Operation

When rectifiers are operated in parallel in a conversion connexion, the voltage across each rectifier will be the same at any instant. During the period of reverse stress, the reverse losses in one rectifier will not be affected by the presence of another, since the applied reverse voltage is a circuit property only.

During the period of forward conduction, however, the common voltage across paralleled rectifiers may cause quite different currents to flow in each rectifier, due to their dissimilar forward characteristics. As a result, the maximum rated dissipation of one of the rectifiers may be exceeded.

In the absence of any current balancing influence (from series equalizing resistors[2], or magnetically coupled series reactors) excess dissipation can be avoided by limiting the dissimilarity of characteristics of paralleled rectifiers.

Consider one rectifier X carrying a mean forward current I_{mn} in a three-phase bridge connexion. The forward losses will be

$$W_f = I_{mn} \,.\, V_{pk} \text{ watts} \quad \ldots\ldots\ldots \quad (18)$$

V_{pk} being the forward voltage at $I_{pk} = 3I_{mn}$.

If I_{mn} is the rated mean current for that rectifier, one may call its peak voltage drop $V_{x(pk)}$ (i.e. at $3I_{mn}$).

The parallel connexion of a second rectifier Y, of peak voltage drop $V_{y(pk)}$, will cause a common peak voltage drop $V_{xy(pk)}$ when a total mean current of $2I_{mn}$ is taken from the two. Fig. 4 shows that the total current $2I_{mn}$ is composed of $I_{x(mn)}$ from rectifier X and $I_{y(mn)}$ from rectifier Y.

The increase in mean current through the first rectifier from I_{mn} to $I_{x(mn)}$ produces an increase in forward loss of

$$\delta W = I_{x(mn)} \,.\, V_{xy(pk)} - I_{mn} \,.\, V_{x(pk)} \text{ watts} \quad \ldots\ldots \quad (19)$$

which is controlled primarily by the relationship between the peak voltage drop of the two rectifiers, $V_{x(pk)}$ and $V_{y(pk)}$ (see Fig. 5).

The increase in dissipation must not allow rectifier X to exceed its rated maximum dissipation if safe operation is to be assured. The greatest value of δW possible is dependent upon the nominal loss $I_{mn} \,.\, V_{x(pk)}$ of rectifier X; the higher the peak voltage drop $V_{x(pk)}$ the less will be the allowable increase δW. Thus to Fig. 5 may be added a curve showing the maximum allowable increase in forward loss, from zero at the voltage corresponding to the rated maximum forward dissipation, to that dissipation at zero voltage.

The intersection of this maximum forward loss increase curve with the $V_{x(pk)}$ curve gives the limit peak voltage drop $V_{y(pk)}$ of a rectifier that may be safely operated in parallel. Continuation of this construction evaluates the voltage range of rectifiers that may be paralleled in this way (see Fig. 6).

Series Operation

Rectifiers operated in series carry the same current at all instants. Forward conduction losses in one rectifier are therefore unaffected by the presence of others in the series if thermally separate, and each rectifier will perform in the same way as it would if operated separately. In the case of thermal interconnexion (see next section) a temperature equalizing effect will naturally ensue.

The reverse voltage distribution, on the other hand, will be determined by the reverse characteristics of the individual rectifiers in the series and will not necessarily be equal.

Consider two rectifiers X and Y operated in series, whose rated reverse voltages are V_x and V_y. Operation at an applied voltage equal to $(V_x + V_y)$ will cause a voltage distribution shown by the points P, Q, R, S in Fig. 7, the actual operating point being controlled by the relative junction temperatures of the two rectifiers.

If both junctions are at temperature T_1 the operating point is P; it is seen that rectifier X is stressed at greater than its rated voltage V_x. Should this cause an increase in junction temperature, the operating point will tend to move towards s, thus reducing the stress across X. This effect, together with the simultaneous understressing initially of rectifier Y, tends to stability[4]. The diagram also demonstrates the desirability of defining the rated c.w.v. of the rectifiers at their maximum operating junction temperatures.

The use of shunt resistance to impose an equalized voltage distribution in the series is not normally desirable,

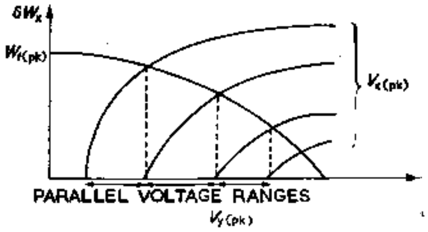


Fig. 6. The maximum forward loss increase curve determines the forward voltage range of rectifiers that may be paralleled

since it reduces the sensitivity of the self-stabilizing effect and could be dangerous in the event of a resistor open-circuit. Shunt resistance and capacitance is sometimes necessary, however, to equalize the distribution of high frequency transient voltages by swamping the rectifier capacitance and the unequally distributed capacitances to earth of the rectifiers.

Thermal Interconnexion

It will be readily appreciated that stud-ended rectifiers on identical but separate heat sinks may operate at different temperatures when carrying the same forward current, due to differences in the total dissipation of particular examples of the same rectifier type. Especially with the higher power rectifiers, small differences of forward characteristic can thus cause considerable variations in junction operating temperatures. Mounting on a common heat sink can reduce these temperature differences, and assist towards reverse voltage equalization in series connexion.

Less apparent, probably, is the solution to the rectifier designer's problem of ensuring acceptable reverse voltage distribution in an overall encapsulation of series connected junctions. The encapsulation could, for example, be of the wire ended form where heat extraction is possible only from the lead terminations, and the separate junctions may have differences in forward characteristic that would cause significant inequalities in operating temperature if the junctions were separately encapsulated and cooled. It can be shown that by manipulation of the internal thermal resistances, a set of asymmetric parallel paths can be presented to heat flow from each junction, and the junction temperature rises made less susceptible to differences in dissipation of the individual junctions.

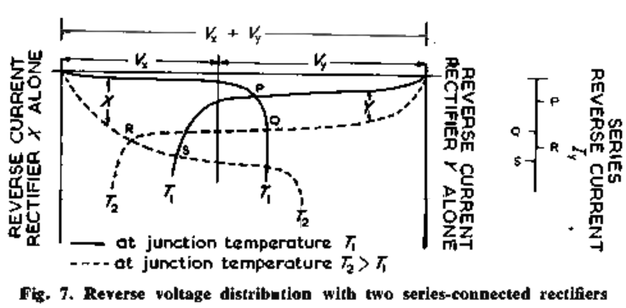


Fig. 7. Reverse voltage distribution with two series-connected rectifiers

Transient Conditions

The previous sections refer to the steady state performance of silicon power rectifiers, where the continuous operating temperature of the junction is the main concern.

Their performance under transient conditions is usually dictated by the two more important criteria of the junction thermal time-constant, and the minority carrier storage effects.

Junction Thermal Time-constant

In the same way that a heat sink possesses a thermal capacity that influences the rate of rise of rectifier stud temperature, there is a thermal capacity associated with the internal thermal resistance of the rectifier, that will control the junction temperature rise under transient conditions. The junction thermal time-constant may be evaluated from its cooling properties after pulsed heating[6,7,11].

In practice, the transient heating effects on the junction of a forward current overload determine on the one hand the type of duty that can be performed by the rectifier from a given steady state condition, and on the other the form and speed of operation of protective devices that are necessary to preserve the rectifier characteristics.

The junction cooling characteristic will to some extent be influenced by the rectifier external thermal resistance, whose properties will thus have a bearing on the intermittent duty possible from the rectifier.

Minority Carrier Storage

The effect of minority carriers present in the base region after commutation is to generate a reverse voltage transient on being swept out, due to the presence of inductance in the circuit. In polyphase connexions these transients appear in the appropriate phase relationship across each rectifier[4,12].

In most cases it is possible by suitable circuit design to reduce the transient voltages to a value consistent with the reverse voltage rating of the rectifiers, whose connexion performance need not then be restricted by their minority carrier storage properties.

Conclusions

The parameters introduced in this article are pertinent to the assessment of performance of individual rectifiers; the considerations for series and parallel arrangements concern the performance of rectifiers in one arm only of a conversion connection.

It should be pointed out, in conclusion, that the determination of rectifier dissipation is useful in the evaluation of overall efficiency of a connexion, and may be used to estimate the voltage drop of a connexion, since the total dissipation is approximately equal to the product of the output current and the connexion voltage drop.

Acknowledgments

The author is indebted to Mr. K. A. Matthews and members of his staff for their advice and assistance, and wishes to acknowledge the permission of Standard Telephones and Cables Limited to publish this article.

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A TACHOMETER WITH A HIGH SHORT-TERM ACCURACY

By C. Reed*

An alternator speed measuring instrument is described in which speed, within a range of approximately ±10 per cent of a nominal, is indicated. The method used results in a linear scale and can be adapted for other nominal speeds and ranges. Accuracy is maintained by a built-in crystal controlled oscillator and calibration is easily checked and adjusted. Some refinements are suggested.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

SINCE the early days of steam engines many methods have been proposed and used for obtaining information on speed of rotation. All of the earlier methods involved mechanical attachments of one kind or another, later came stroboscopic and photocell systems. Apart from differences in accuracy in the relatively modern methods of rotational speed measurement, presentation of the information takes different forms. Systems employing neon indicators[1], Lissajou figures[2], and cold-cathode counter tubes [3,4], have all been described before.

No method, however, is ideal for all applications. A transducer, in conjunction with cold-cathode counter tubes and a one-second count, can give a very accurate measurement of the rotational speed of a body rotating at constant speed. If, however, the speed is changing rapidly and a number of readings are required at short intervals, the use of a suitably damped meter for presentation of the results is practically essential. Such a requirement arose recently during the on-load testing of alternators, the driving force of which was subject to variation, and it is the intention of this article to describe the form of tachometer used.

The Tachometer

The tachometer to be described has been specifically designed for measuring the speed at which alternators are driven. In this particular case, the frequency generated by the alternator during testing was used as the source of speed information, and was nominally 1 100c/s. The allowed limits were 1 075 to 1 125c/s over the range of driving force expected during the tests. However, a frequency range of 1 025 to 1 175c/s was considered desirable to assess the extent of departure from limits. To obtain an open scale over this range of frequency it was decided to subtract 1 000c/s electrically from the input frequency and to indicate the difference only, this, when added to 1 000c/s, giving the result. Fig. 1 shows the method used.

Circuit Operation

The high-pass filter at the input is necessary for reasons which will become plain shortly. The input signal, at a frequency of approximately 1·1kc/s, is mixed with a locally derived signal of 1kc/s and, in the anode circuit of the mixer, there are the usual modulation products including a frequency representing the difference between the local oscillator and the incoming signal frequencies. Taking the nominal case, this will be 100c/s.

The output of the mixer is detected and passed through a low-pass filter, which removes all modulation components, except those in the range 0 to 200c/s. Again, taking the nominal case, the resulting 100c/s sine wave is amplified and then passed to a Schmitt trigger circuit. The output from this circuit is differentiated and passed, via a cathode-follower, to trigger a twin-triode constant area pulse generator. A 1mA meter, in the anode circuit of the first half of the twin-triode, integrates the chain of pulses and indicates speed directly. The meter has a full scale reading of 200 which now corresponds to 1 000+200=1 200c/s. In the nominal case, the reading would be 1 000+100=1 100c/s.

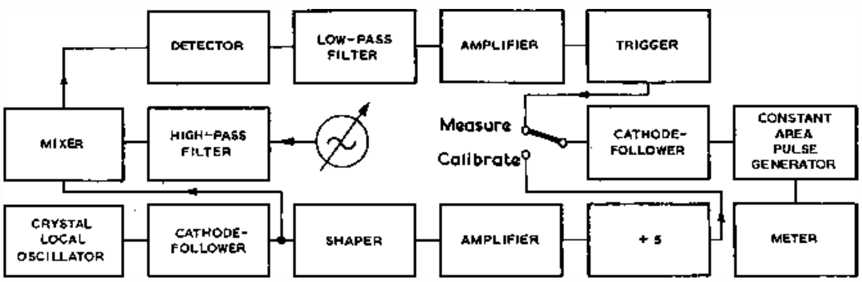

Fig. 1. General arrangement of the tachometer

To set up the constant area pulse generator, a known frequency, in the range 0 to 200c/s is required, and this is conveniently derived from the 1kc/s local oscillator. Its frequency is divided by five and arranged to trigger the last stage at $\frac{1\,000}{5}=200$c/s and potentiometer RV_4 is then adjusted until the meter reads exactly full scale. To take full advantage of the scale dimension, a meter of the 270° deflexion type is used.

A 'press to calibrate' button in conjunction with a pre-set potentiometer permits the calibration to be checked and, if necessary, adjusted immediately prior to taking readings. Thus, the only appreciable inaccuracy is due to reading errors.

Circuit Details

Since an input frequency in the range 800 to 1 000c/s could also produce a difference frequency of 200 to 0c/s and operate the Schmitt trigger, a high-pass filter with a cut-off frequency of 1kc/s is inserted between the input socket and the mixer. The filter used takes the form of a prototype section with two M-derived half-sections having $M=0·6$, as shown in Fig. 2. With an input of one volt, peak-to-peak, no trouble is experienced from 'lower sideband' signals.

The mixer is conventional, but the impedance of the anode load is made to fall with frequency to reduce the amplitude of unwanted modulation products. The low-pass filter, follow-

* *E.M.I. Electronics Limited.*

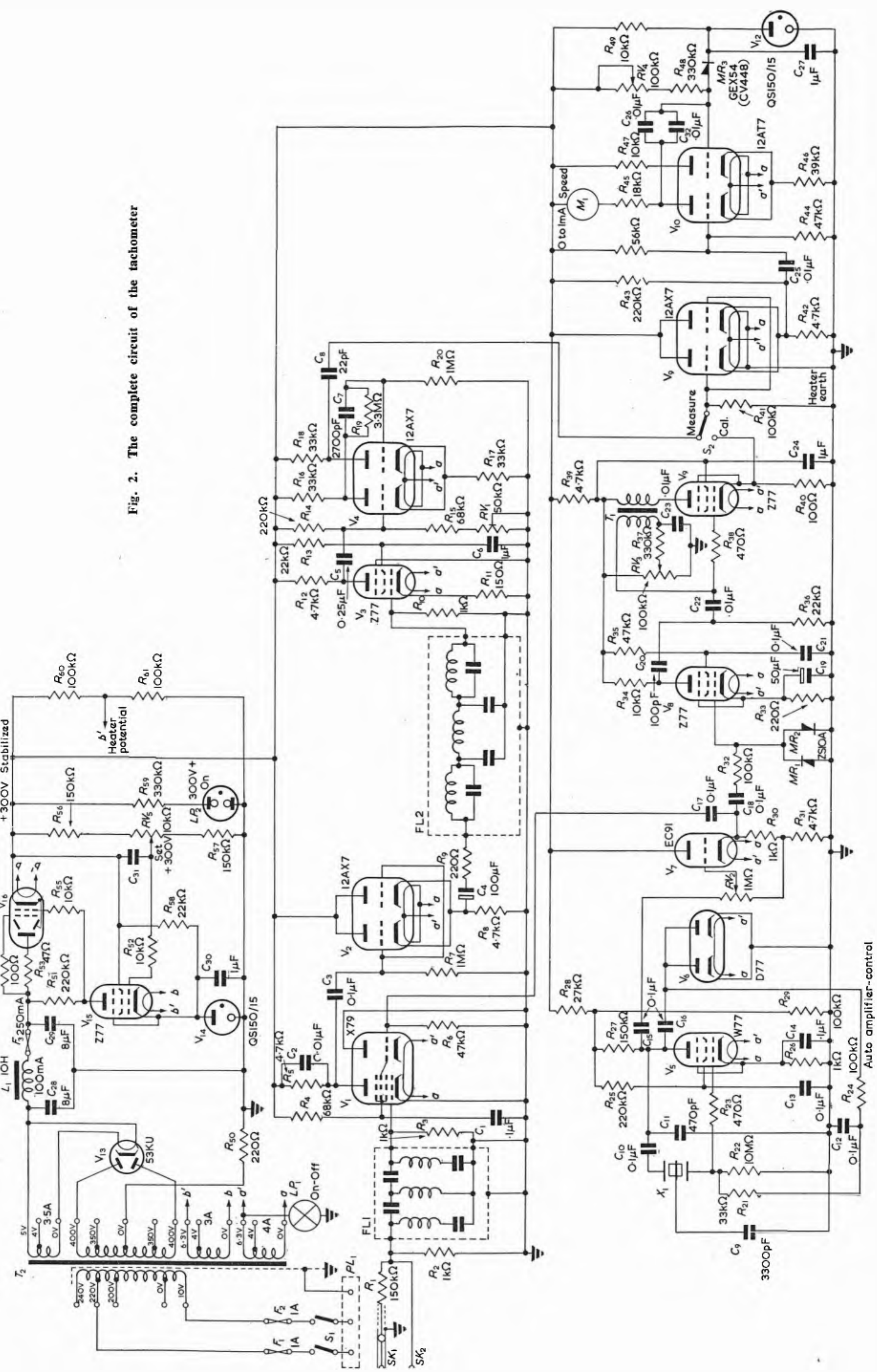

Fig. 2. The complete circuit of the tachometer

ing the detector, is of the same type having $M=0{\cdot}6$ and $f_c=$ 200c/s.

The differentiated output from the Schmitt trigger is fed, via a cathode-follower normally biased to cut-off, to the constant area pulse generator. This ensures that the negative spikes do not interfere with its operation.

The local oscillator is crystal controlled and its output is accurate to within $\pm0{\cdot}015$ per cent (or 0·15c/s) over a temperature range of $-20°$ to $+70°$C. The output of the local oscillator, also passes through a cathode-follower and is squared by using silicon diodes which have negligible forward conductivity until the applied potential exceeds approximately 0·5V. The resulting square wave is amplified and then differentiated and used to lock the 'divide by five' blocking oscillator, the output of which is fed to the grid of the cathode-follower preceding the constant area pulse generator. In this way, the impedance 'seen' by the first grid of the pulse generator is made to appear sensibly the same, no matter whether the input is a tachometer 'signal' or for 'calibration'.

Conclusions

Although the tachometer described satisfies the original requirement, there are some refinements that could be incorporated. To minimize the scale error, the division ratio for calibration could be raised to 10:1 thus ensuring higher accuracy at the nominal frequency. This could be achieved by using a cold-cathode counting tube, since the long term stability of a 'divide by ten' blocking oscillator divider is rather questionable. In addition, by making use of two of the Dekatron cathodes and, hence, dividing by five, a linearity checking means can be provided.

In the event of the speed of the machines under test falling outside the range, 'speed high' and 'speed slow' indications would be desirable. This can easily be achieved by including two relays, one operated by the output of each filter, so as to indicate the following states:

Output from the high-pass, but not from the low-pass filter .. .. Speed High

No output from the high-pass filter .. .. Speed Low

Signal lamps on the front panel could then indicate the prevailing state.

With suitable filter modifications and an appropriate choice of local oscillator frequency and division ratio, this method of speed measurement can be used at practically any other nominal speed and, within limits, for any given range.

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Electronic Control for an Oil Refinery

An electronic control system manufactured by Evershed and Vignoles Ltd, Chiswick, London, is now being installed at the Dinslaken Oil Refinery of the B.P. Benzin and Petroleum Aktiengesellschaft, Germany, which has a daily throughput of 100 000 barrels.

The instrumentation and control equipment applied to two crude distillation plants, a hydrofiner, a catalytic reformer and a redistillation unit. The whole of the indication and control facilities are terminated at the control desks and cubicle. 120 separate control loops are provided, some of them being inter-related by re-setting and cascading arrangements. Temperature measurements from 340 different points throughout the refinery, effected by thermocouples and used for indication or control purposes, form part of the installation. There are also two analysis recorders for percentage H_2 and O_2.

One of the features of the control desks and cubicle, is the use of four quadruplex recorders arranged to permit selective recording from 64 different points, distributed in all parts of the plant.

It is thus possible to provide comprehensive facilities for—say 'start up' with a minimum of panel space and with an exceptional degree of simplicity. Each of the 2in charts of the recorders may be connected by means of a standard jack-plug system to any one of four measuring circuits, so that each recorder covers 16 points. The input signal to the potentiometric recorders is derived across a 1Ω precision resistor which is inserted into the circuit when required.

Selection of temperature measurements to accommodate the magnitude of information required at the central control room is provided by a switching system based on telephone type key-switch boards mounted on each control desk, so that despite the relatively small number of recorders, comprehensive coverage is provided.

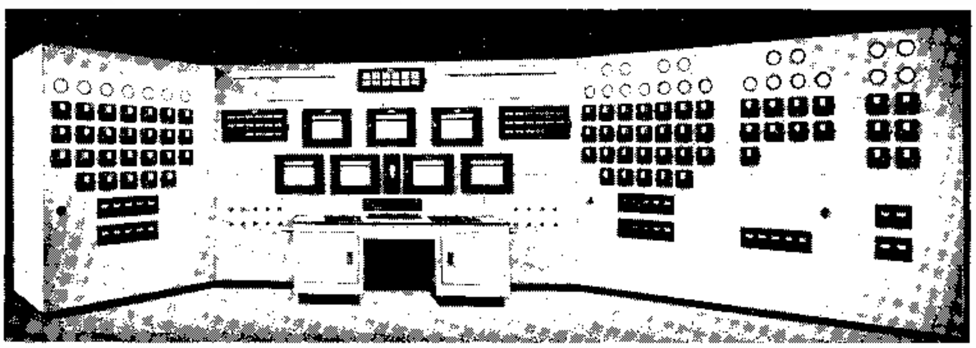
In general, the temperature measuring circuits consist of a thermocouple feeding into a potentiometric recorder via the switchboard. There are, however, certain loops where d.c. amplifiers are employed to convert the thermocouple output voltage into a signal suitable as measured value input for a process controller.

During normal operating conditions in individual sections of the refinery, a 'normal operation' window is illuminated on the control panel. Alarm conditions automatically extinguish this light and provide flashing illumination of an 'alarm condition' window. Acknowledgment of the alarm, by depressing a push button on the panel, will change the flashing signal to a steady alarm signal. The interlocking circuits are so arranged that the steady, 'acknowledged alarm' condition cannot be terminated until the cause of the alarm has been removed.

Rapid 'one glance' supervision of the more significant plant parameters is afforded by the wide use of 'In-Line Scanners'. These instruments are essentially moving coil indicators permanently inserted in the control loops. They present an unbroken shadow line when conditions in each loop are at the desired value. Any break in the shadow line will immediately pin-point the location of an error condition, together with the direction of the error and some indication of its magnitude. By varying the sensitivity of the individual 'In-Line Scanner', small deviations in the more critical parameters can be made to result in a larger deflexion than resulting from changes in those parameters which can be tolerated to fluctuate to some degree.

This system makes it possible for a single operator to supervise a multitude of plant functions, the visual scanning of which would be very difficult if continuous reading of similar numbers of conventional instrument dials were required.

The Control Console



A Multi-range Electrometer Amplifier Using Variable Feedback

By J. H. Leck*, M.Eng., Ph.D., A.M.I.E.E., A.Inst.P. and W. E. Austin*

The simple electrometer amplifier which is described in this article has been found both accurate and reliable over a period of twelve months for the measurement of positive ion currents down to 10^{-15}A. By using a modern miniature electrometer valve in conjunction with a high gain transistor d.c. amplifier having a large overall negative feedback, the advantage of simplicity is combined with that of high accuracy and adequate sensitivity. The inherent disadvantage of the transistor d.c. amplifier, its temperature instability, has been overcome in this application by using a silicon transistor in the first stage and operating with an input from a constant current source.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

THE basic principles of this circuit are well established. The unknown input current develops a voltage across a large resistor ($10^{11}\Omega$ for example) which is the grid leak of an electrometer valve. The resulting changes in anode current of this valve—due to the induced grid voltage—are measured by means of a high gain d.c. amplifier. A known fraction of the output voltage of the amplifier is fed back to the cathode of the electrometer valve to give a large negative feedback. Thus the overall sensitivity is determined by the feedback ratio and is virtually independent of the characteristics of the amplifier. In this design the electrometer anode current is fed into a transistor rather than a conventional valve amplifier. It is well known that a transistor d.c. amplifier can be used only if the instability introduced by temperature changes at the junction are negligibly small. This is achieved by using a silicon pnp transistor (OC201) in the input stage fed from a high impedance source. When operated in this manner the silicon transistor has the advantage over the germanium by at least an order of magnitude.

The electrometer pentode ME1403 provides virtually a constant current input to the transistor because of its inherently high dynamic anode resistance of much greater than $10^{9}\Omega$. A temperature change of 1°C results in an apparent change of the order of 10^{-9}A in the input signal of the OC201 which is equivalent to a change of signal at the electrometer grid of 0·1mV. (The g_m of the ME1403 is assumed to be 10^{-5}A/V.)

Because of the current amplification of at least 20 in the first stage of the amplifier germanium transistors can be used in subsequent stages. With an OC45 the intrinsic current increases by approximately 0·05μA/°C at 20°C. Thus each 1°C temperature rise of this junction is equivalent to an input signal of $-$0·2mV at the input grid. The temperature changes at subsequent stages are, of course, correspondingly less important.

Description of Circuit

The complete circuit diagram is shown in Fig. 1 and is almost self-explanatory. The small input current develops a voltage across the resistor R_g which is applied to the control grid of the electrometer valve. Any guard rings required in the input lead should be connected to the zero voltage supply line. This valve is operated under the normal recommended conditions, with a grid bias of $-$3V and an anode

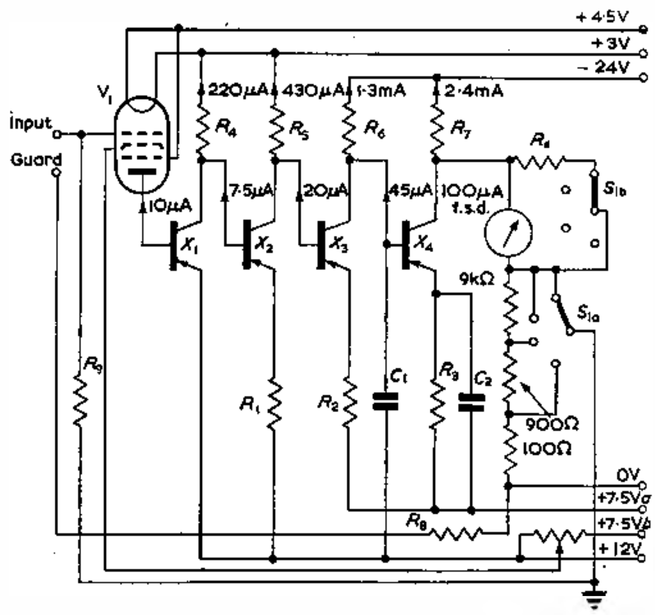


Fig. 1. The complete circuit

$R_1 = 1k\Omega$, $R_2 = 82\Omega$, $R_3 = 1k\Omega$, $R_4 = 39k\Omega$
$R_5 = 10k\Omega$, $R_6 = 22k\Omega$, $R_7 = 10k\Omega$, $R_8 = 1M\Omega$
$C_1 = 8\mu F$, $C_2 = 0{\cdot}02\mu F$
V_1 = ME1403
X_1 = OC201, X_2 = OC45, X_3, X_4 = OC71

current of 10μA, the anode voltage being 9V positive with respect to the cathode and the screen grid variable through the range 4·5 to 9V. Variation of the screen potential (for which a separate coarse and fine adjustment is recommended) allows zero setting of the amplifier. The anode current provides the input to a four-stage d.c. coupled transistor amplifier. In this amplifier the resistors R_1, R_2 and R_3 provide internal feedback at the intermediate stages. The values for the collector load resistors R_4, R_5, R_6 and R_7 are

* *The University of Liverpool.*

determined by the desired input current to their succeeding stages. (The current levels through the amplifier are shown in the figure). The output current flows through a resistor chain of total value 10kΩ.

Overall negative voltage feedback is applied by returning the electrometer grid resistor R_g to a tapping on this output resistor chain. With the particular values shown in Fig. 1 100 per cent, 10 per cent or 1 per cent can be selected by means of the switch S_1. Capacitors C_1 and C_2 are included in the circuit as shown, C_1 to improve the stability and C_2 to improve the frequency response. (C_1 is essential in order to prevent self-oscillation when operating with 100 per cent feedback.) Four ranges of sensitivity can be selected by the switch S_1. In position (1) the shunt resistance reduces the sensitivity of the output meter by a factor of 10 which clearly makes a full scale deflexion for an input signal of 10V (developed across R_g). In position (2) where the meter shunt is removed but full 100 per cent feedback retained the overall sensitivity is obviously increased by a factor of 10. In switch positions (3) and (4) the sensitivity is increased to 0·1 and 0·01V for full scale deflexion by reducing the feedback to 10 per cent and 1 per cent respectively.

Operating Characteristics of the Amplifier

Although all the results quoted in this section refer to one particular amplifier, checks have been made and satisfactory correlation obtained for four other units built to the same circuit diagram but to different circuit layouts.

The overall internal voltage gain was 4 000 and the current gain of the four-stage transistor amplifier 20 000. The sensitivities of the four ranges were as calculated to within ± 1 per cent, i.e., 10, 1·0, 0·1 and 0·01V for full scale deflexion, except for the most sensitive range which was 2 per cent low. (This is to be expected because with minimum feedback ($\beta = 0{\cdot}01$) the internal gain of the amplifier cannot be neglected compared with $1/\beta$). On all ranges the linearity was better than ± 1 per cent of full-scale deflexion (the limit of accuracy of the output meter). The output reached saturation with a positive input signal of 14V. The above performance could be achieved for negative input voltages simply by reversing the output meter connexion. However, for negative inputs saturation was reached with an input of 4V. (This could be overcome by slight redesign of the output stages at the expense of reducing the range for positive signals). These measurements of sensitivity and linearity were made by applying an input voltage across the grid leak resistor R_g. With 100 per cent feedback applied the bandwidth was found to be 25kc/s. The importance of the stability of the various power supplies is summarized briefly in Table 1, which gives both the measured current drain in each line together with the effect on the amplifier output of a 1 per cent change in supply voltage. It is clear that the stability of the 3, 7·5 and 12V lines and the heater supply to the electrometer are the only ones of any critical importance.

TABLE 1

Current drain from each supply line together with the change in the zero setting of the output meter (expressed in terms of an equivalent input signal) for a 1 per cent fluctuation in supply voltage.

SUPPLY (V)	SUPPLY CURRENT (mA)	ZERO SHIFT (mV)
+3	0·66	30
+7·5 (a)	3·7	0·2
+7·5 (b)	0·01	75
+12	0·67	100
−24	3·7	0·5
Electrometer heater	8·2	15

The following measurements of amplifier stability were made with the resistor R_g short-circuited and the supplies, including the electrometer heater, obtained from batteries. The background fluctuations in the output meter were such that an input signal of 0·1mV could just be detected. Under laboratory conditions with reasonably constant temperature (less than 2°C change during the day) drift rates of less than 0·25mV/h were consistently observed. The temperature coefficient of the zero was found to be approximately 1mV/°C. With R_g in circuit there was an additional component to the background fluctuations due to the voltage pick-up at the electrometer grid. This naturally depended upon the magnitude of R_g, the efficiency of screening the electrometer unit and the presence of disturbing electrical fields. Up to the present time experience has been limited to R_g not greater than $10^{11}\Omega$. With this grid resistor and good screening of the electrometer box and grid input lead, the total background level varied considerably, at best being equivalent to an input signal of 0·1mV ($I_o = 1 \times 10^{-15}$A) and at worst 0·5mV ($I_o = 5 \times 10^{-15}$A). These fluctuations were quite random in nature, having time-constants up to 10sec. (The fluctuations in the output meter could no doubt be reduced by increasing its time-constant up to 10sec.)

Conclusions

Operating experience with these amplifiers up to the present time has been confined to the measurement of small positive ion currents in mass spectrometer work using a $10^{11}\Omega$ input resistor. Completely satisfactory operation has been obtained over a period of more than a year. Because of their simplicity and reliability batteries have been used for all the supplies. This has proved most satisfactory. There have been no maintenance problems other than the replacement of the batteries at infrequent intervals. The battery replacement problem can be reduced if a −24 and +7·5(a) voltage supplies are obtained from a mains rectifier. As only a minimum of stabilization is required for these supplies a small constant voltage transformer in the a.c. mains is more than adequate. Thus, because the drain from the +12 and +7·5(b) volt lines is small, battery replacement occurs only at intervals of many months.

The biggest operating disadvantage, and the one which to a large extent controls the long term zero drift is heater current supply to the electrometer valve. This, however, is a problem common to all electrometer amplifiers. To the present, again because of considerations of simplicity, no attempt to use an electronic stabilized supply has been made. All the work has been carried out with either a dry cell or a well charged 2V accumulator. The limiting drift introduced by the transistor amplifier is almost certainly not greater than 0·2mV/h and a total drift level of this order could be obtained given a sufficiently stable supply for the electrometer (see Table 1). Again the short term background fluctuations are set by the input resistor and not by the amplifier. In the present application a current of 10^{-15}A can be detected only under ideal conditions. In normal work the background level is equivalent to a signal of between 2 and 5×10^{-15}A.

Acknowledgments

The authors would like to thank Professor Craggs for the continuous interest which he has shown in this work.

A Voltage Tuned Resistance-Capacitance Oscillator

By W. D. Ryan*, M.Sc. Ph.D. and F. E. Hetherington†, M.Sc. (Eng.).

The variable capacitance exhibited by selenium dry-disk rectifiers with reverse bias is utilized in a bridge type resistance-capacitance oscillator circuit to produce a variable audio frequency dependent on an applied d.c. signal. The characteristics of suitable rectifiers, the circuit diagram of the oscillator and its operating characteristics are described.

(Voir page 127 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 132)

IN frequency-modulated sub-carrier telemetering systems difficulty is experienced in obtaining the relatively large deviation of the audio frequency employed. Methods proposed in the past include the use of beat-frequency oscillators with reactance modulation and resistance-capacitance oscillators in which the frequency is controlled by non-linear resistance in the phase-shift network. This may be the anode resistance of a valve[1] or a voltage-dependent resistor such as silicon carbide[2]. The present system, which was to be used to relay scintillation data from a radio astronomy field station to a centrally located receiver and pen recorder, utilizes the capacitance of a dry-disk rectifier to control the frequency of the modulating oscillator.

Selenium Rectifiers as Voltage-Dependent Capacitors

Certain selenium rectifiers exhibit a relatively large capacitance dependent upon the applied d.c. voltage. With selected elements the capacitance may be reduced to about one tenth of its zero bias value with the application of a suitable reverse bias. Associated with this capacitance is a shunting resistance of such a value that the unit has a Q which may be as low as ten in the audio frequency range. This tends to limit applications to simple resistance-capacitance networks where elements of low Q are acceptable.

In investigating the properties of different samples, a wide variation in characteristics is found although identical units did not vary greatly among themselves. Undesirable properties found included low values of d.c. back resistance and consequent variation in capacitance due to heating, long period time lag in capacitance change, possibly due to trapping centre migration, which severely limited the frequency of the applied bias variation, and a hysteresis effect associated with rising or falling bias voltage. The available ranges of percentage capacitance variation differed widely among the types of rectifiers tested. Differences in materials and manufacturing techniques undoubtedly account for most of these anomalies. The most suitable elements for the present application were found to be those in which the selenium had been deposited electrolytically.

The performance of the phase-shift network may most simply be described in terms of the capacitance C and the dissipation factor D of the capacitive elements. Fig. 1 shows the variation of rectifier capacitance with frequency for selected values of reverse bias. It can be seen that the capacitance falls slightly with increasing frequency. This effect is small in comparison with the change produced by the bias and is assumed to be constant in the analysis. The equivalent parallel conductance G of the rectifier is a function of both frequency and bias as can be seen from Fig. 2 which shows the variation in conductance for a unit tested by a.c. small signal bridge methods. Since G varies almost linearly with frequency, a sufficient approximation would be to write:

$$G = G_0 + K\omega \qquad \text{.......... (1)}$$

where G_0 (the intercept at zero frequency) and K (the

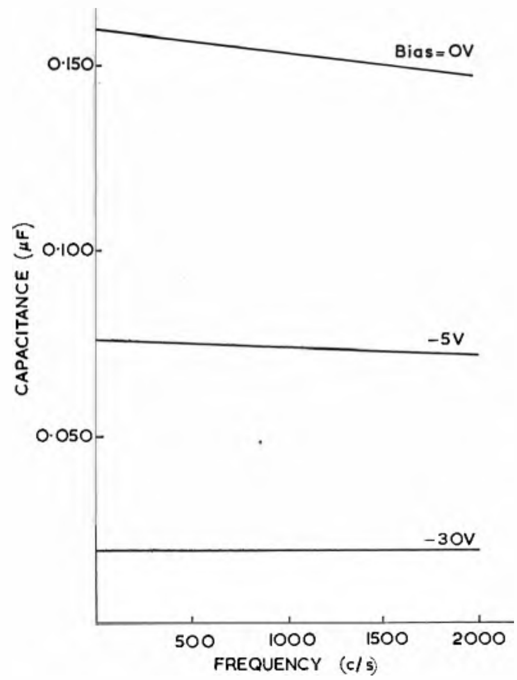


Fig. 1. Variation of rectifier capacitance with frequency for fixed values of reverse bias

slope) are functions of bias. If B is the capacitive susceptance of the element, then:

$$D = G/B$$
$$= G_0/\omega C + K/C \qquad \text{....... (2)}$$

Thus the characteristics of the rectifier may be described in terms of G_0, K/C and C. The variation of these parameters with applied bias is shown in Fig. 3. Since, in the present oscillator, ωC is approximately constant, it is apparent that D will increase at low and high values of bias reaching a minimum in the mid-range. It will be shown that a corresponding loss occurs in the phase-shift network. A frequency range of about eight to one may be expected as a result of the variation in capacitance from $0{\cdot}16\mu F$ to $0{\cdot}02\mu F$.

The Phase-Shift Network

The most economical method of incorporating these units in an oscillator is through the use of a modified Wien

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† *Royal Military College of Canada*

bridge network, shown in Fig. 4. The modification enables bias voltage to be applied to the rectifier units in parallel as shown in the circuit diagram (Fig. 5). With an ideal amplifier, oscillation will occur when the phase-shift

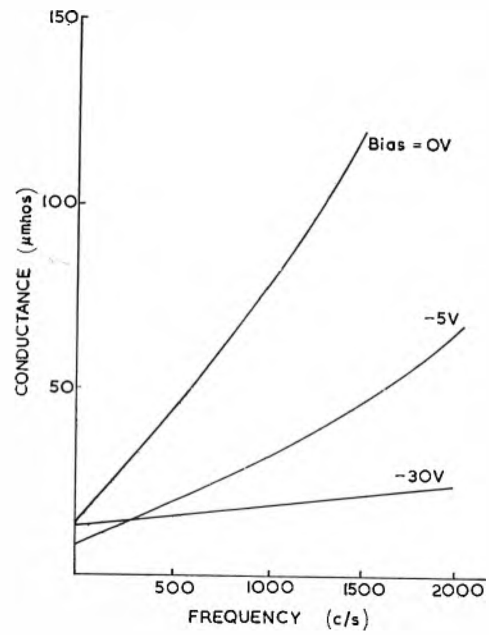


Fig. 2. Variation of rectifier a.c. conductance with frequency for fixed values of reverse bias

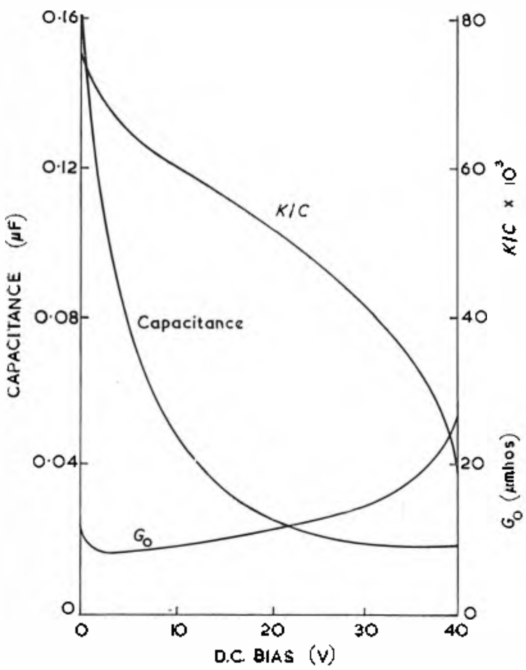


Fig. 3. Variation of rectifier parameters with applied bias

through the network is zero. In general:

$$(V_1/V_2) = 1+(G_b/G_a)+\frac{G_bY_1}{G_aY_2}+(Y_1/G_a)+(G_b/Y_2) \quad .. \ (3)$$

Now

$$Y_1 = G_1+j\omega C_1 = \omega C_1(D_1+j)$$

and

$$Y_2 = G_2+j\omega C_2 = \omega C_2(D_2+j)$$

If D_1 and $D_2 \leqslant 0{\cdot}1$, then it may be shown that:

$$\omega \simeq \sqrt{\left(\frac{G_aG_b}{C_1C_2}\right)} \quad .... \ (4)$$

and the attenuation A at this frequency is given by:

$$A \simeq 1+(G_b/G_a)+\frac{G_bC_1}{G_aC_2}+(\omega C_1/G_a)\,.\,D_1+(G_b/\omega C_2)\,.\,D_2 \quad .... \ (5)$$

In the present arrangement $G_a \simeq G_b$ and $C_1 \simeq C_2$, hence:

$$A \simeq 3+D_1+D_2 \quad .... \ (6)$$

Thus an added term appears in the expression for the attenuation consisting of the sum of the dissipation factors of the two capacitive elements. An additional factor influencing the variation of network loss with applied bias is the fact that the admittances Y_1 and Y_2 will not necessarily have identical values. Though these variations are small, the amplitude of oscillation will change until a loop

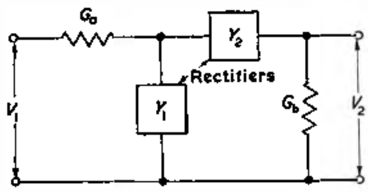
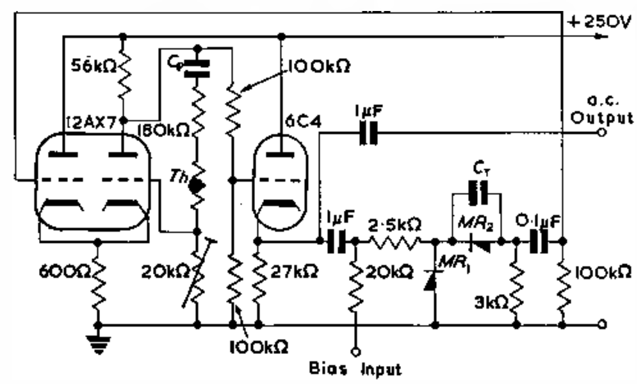


Fig. 4. (left) Phase-shift network

Fig. 5. (below) Circuit diagram of the oscillator



gain of unity is again achieved. This change may be quite large depending on the type of amplitude limiting employed. A long time-constant limiter such as a thermistor or a lamp would maintain a constant output but would limit the maximum rate of frequency modulation. An instantaneous limiter is to be preferred and is used in the present system. In order to avoid the high distortion which would be introduced by this method some form of compensation is required in the oscillator circuit.

The Oscillator Circuit

A nominal centre frequency of 1kc/s was chosen for the oscillator. This decision was influenced to some extent by the characteristics of the particular rectifiers available. The complete circuit diagram is shown in Fig. 5. The first stage is a cathode-coupled amplifier having an effective voltage gain of approximately eight with zero phase shift. The stage gain may be controlled by negative feedback from the anode to the grid of the second section of the 12AX7. This stage is direct-coupled by a resistance voltage divider to a 6C4 cathode-follower, which drives the phase-shift network. The output signal is also taken from the cathode-follower. The rectifiers MR_1 and MR_2 are D5H6MN1 six section selenium stacks produced by the Syntron Company. The six elements in series were necessary in order to obtain a sufficiently high shunting resistance to enable the phase-shift network to be driven from the cathode-follower. The higher network loss at low and high values of bias may be partially compensated for by C_p and C_t respectively. At low frequencies the amount of negative feedback is reduced due to the increased reactance of C_p, resulting in an increased gain. At high frequencies, C_t, in parallel with rectifier MR_2 (admittance Y_2), reduces the ratio G_bC_1/G_aC_2,

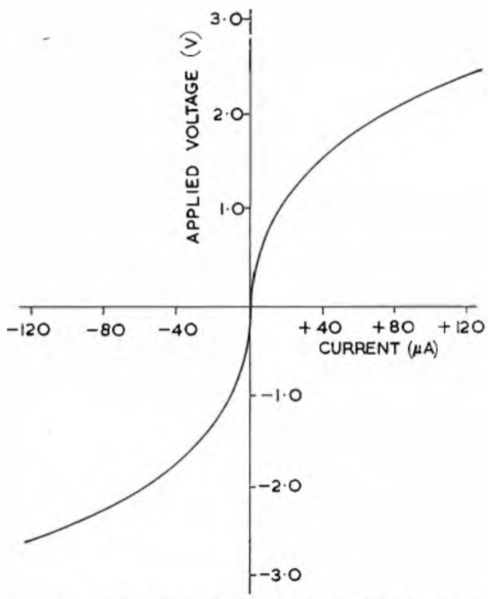


Fig. 6. Static characteristic of silicon carbide resistor

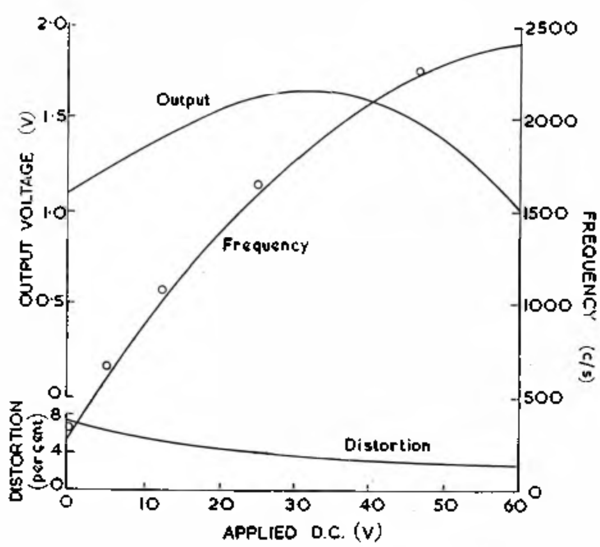


Fig. 7. Oscillator performance

thus decreasing the network loss. With the present rectifiers, complete compensation is not possible although constant amplitude may be obtained over a limited frequency range at the expense of much greater amplitude variations outside this range.

Limiting

The amplitude of oscillation is limited by the non-linear silicon carbide resistor *Th* connected in the negative feedback path. The d.c. characteristic of the element used is shown in Fig. 6. It can be seen that when the peak a.c. voltage across the non-linear resistor increases the effective a.c. resistance falls and the amount of negative feedback is increased. The 180kΩ resistor in series with this element modifies its characteristics so that the output amplitude is maintained at approximately 1·5V with minimum distortion. If the value of the resistor was reduced, more effective limiting would be obtained at the expense of appreciably higher distortion.

Performance of the Oscillator

Fig. 7 shows the measured performance of the oscillator. A frequency range from 300 to 2 400 c/s was obtained for a signal change of 60V. It should be noted that the effective bias applied to the rectifiers is less than the signal, due to the voltage drop in the 20kΩ isolating resistor. The output voltage varied from about 1 to 1·5V r.m.s. while the harmonic content was 3 per cent at the upper end and 7 per cent at the lower end of the frequency range. The open circles are calculated points based on expression (4) in the analysis. The value of G_a used includes the effective output resistance of the cathode-follower.

The frequency stability of the oscillator was checked using a stabilized d.c. supply for the bias. During the first half-hour after initially switching-on the frequency fell by approximately 5 per cent and thereafter remained constant to within 1 per cent. No doubt large changes in ambient temperature would result in greater deviations in frequency and may prove to be a limiting factor.

Conclusion

The wide frequency range which can be obtained with this oscillator, and its relative cheapness and simplicity indicate its suitability for many types of data transmission in which extreme accuracy of response is not required.

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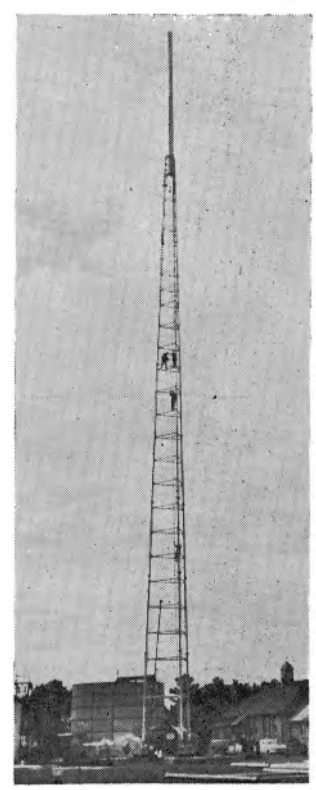
A New Type of Radio Tower



Two towers of light alloy construction have been designed and manufactured by Saunders-Roe Structures Ltd, a subsidiary of Saunders-Roe Ltd and a member of the Westland Group of Companies for erection on a site overseas where a guyed mast could not be used and where the base area of the mast had to be of minimum size.

The towers, one of which is illustrated, are of novel design. They are each 366ft high in the fully extended condition and may be adjusted during erection over a range of 16ft to suit the characteristics of the transmitting installation.

Each side of the triangular base of the tower is only 20ft. The diameter of each of the tubular legs at the base measures only 7¼in graded to 5in at the top, the weight being about 7 tons.

BOOK REVIEWS

Fundamente der Elektronik

By G. Rose. 214 pp. 431 figs. Demy 8vo. Verlag für Radio-Foto-Kinotecknik GmbH; Berlin. 1959. Price DM18.50.

THE purpose of this book is to provide an introduction to the basic principles, components and circuits of electronics. It is especially intended for workers in those fields into which electronics has expanded during the past few years.

In an introductory chapter the various types of electronic component are classified, and a general indication is given of their applications. In chapters 2 to 4 the functioning of the components is explained in more detail, and in each case typical circuits are given. Chapter 2 is concerned with hard valves, namely thermionic diodes, triodes and cathode-ray tubes. Chapter 3 deals with gas-filled valves. The more recent additions to electronics, for example, semiconductor rectifiers and point and junction transistors are described in chapter 4. Photocells, secondary emission tubes, and magnetic amplifiers are also included in this chapter. In a final chapter actual applications are examined more thoroughly. The choice is somewhat arbitrary, but on the whole some of the more recent electronic applications in, for example, control engineering, electronic computing, and medical electronics have been considered.

The presentation of the subject is clear and concise, and the text is not confused by unnecessary mathematics. The book appears to fulfil its purpose very well and should prove an excellent elementary reference work for students of electronics.

G. D. BERGMAN

Fluctuation Phenomena in Semiconductors

By A. Van der Ziel. 164 pp. 17 figs. Demy 8vo. Butterworths Scientific Publications. 1959. Price 35s.

THE author of this volume is a well-known authority on the various aspects of noise in electronic devices, and presents here a survey of the present knowledge of the subject as it applies to semiconductors. It is based on the work of a number of investigators in this field, and the author has endeavoured to present the various sections of the book in such a way that different explanations of the same phenomena are compatible.

The book is intended mainly for the specialist, and it is assumed that the reader is familiar with the diffusion equations for semiconductors, and the concept of donors, acceptors, traps, and recombination centres. A knowledge of the mathematics of probability and correlation functions is necessary in order fully to appreciate the theories of the different noise phenomena; a survey of these methods is given in Chapter 3.

After discussing the characterization of noise in electrical networks, the author presents a detailed account of generation-recombination noise for the case of thermal equilibrium, and then discusses 'flicker' noise in semiconductors, and noise in photoconductors. A short chapter on noise in semiconductor resistors is followed by an extensive treatment of shot and flicker noise in diodes and junction transistors, covering three chapters. The final chapter is devoted to considerations of low-noise circuits, and embraces photocells, transistor amplifiers, and diodes as mixers and parametric amplifiers. Throughout the book, the author has endeavoured to keep before the reader the various problems that need further clarification, in the hope that further research may be stimulated.

The book is very well written and clearly printed, and contains extensive bibliographies at the end of each chapter.

E. HUGHES

Proceedings of the National Electronics Conference, Volume XIV

1 074 pp. 738 figs. Medium 8vo. National Electronics Conference Inc., Chicago. 1959. Price $7.50.

THIS volume gives the texts of the one hundred papers presented at the 14th National Electronics Conference held at Chicago in October 1958. The papers are arranged in 25 groups, each group corresponding to a conference session. The important subjects of transistors, microwaves and servomechanisms had two sessions apiece, the remaining topics each having one session. In this way the contents of the volume give a balanced picture of developments over the whole field of electronics.

Two of the transistor papers are concerned with high-current fast-rising pulses. In the first paper, circuits depending on the phenomena of avalanche multiplication and voltage punch-through are discussed. Currents of up to 30A with rise-times down to 4mμsec are obtainable from ordinary transistors. The second paper describes a radar modulator using transistors and magnetic-cores, the modulator delivers a peak power of 24kW to the magnetron at a pulse-width of 1μsec. Further transistor papers deal with the use of the low collector-emitter impedance of a saturated transistor to replace a diode clamp and considers some theoretical aspects of pulse-width control circuits which allow large powers to be controlled with a modest dissipation within the transistor.

Two of the papers on solid-state deal with electro-luminescence, in one case for a numerical display and in the other for a device known as the 'luministor'. This consists of an excited phosphor optically coupled to a photo-conductor, it forms the basis for several novel circuits. The response-time of the luministor is slow, but it can yield amplifier and two state circuits of very low cost. Other sections describe developments in subjects such as audio, television, aerials and computers together with two groups of papers which deserve special mention. The first of these is concerned with the electronic side of the U.S. satellite programme and the second deals with technical writing. It is only in recent years that the presentation of technical information has been studied for its own sake and this group of papers discusses the outlook of the professional institutions (A.I.E.E. and I.R.E.), the business firm and the general press to the technical manuscript.

The book is well produced and the papers are, in the main, well supported by references to prior work. The volume will undoubtedly find its place, beside its predecessors, on many library shelves.

V. H. ATTREE

Reading German for Scientists

By H. Eichner and H. Hein. 187 pp. Demy 8vo. Chapman & Hall Ltd. 1959. Price 30s.

GERMAN is the mother tongue of both authors who have taught German to students in Canada. Starting from the premise that a reading knowledge of German is required for a Ph.D. in almost all the English-speaking world, they have set out to provide just that in preference to teaching how to speak or write German.

A draft of the book has been tried out in Canadian courses. Grammar and syntax are dealt with in such a compressed form that students may find it difficult to digest it without the help of a tutor or teacher. Chapters like the one on vocabulary building and the analysis of the typical long compound words will benefit even those with a fair knowledge of the language. The authors state that they have dealt with 1 400 words of which 900 have to be memorized while 150 are derivates and 350 have a similarity to English terms.

A strong bias to chemistry can only be explained by the background of the authors. While some of the chemical reading matter is quite up to date, physics and nuclear science are represented by texts published in 1926 and 1937 respectively. The section on "wireless communication" consists of paragraphs from a 1940 publication and a seven-page chapter on "The electron in physics and technology" was originally published in 1948.

As far as scientific German is concerned, the reading matter has another disadvantage: The terminology has changed completely during the last 20 years or so and has been 'Germanized' in many cases. The new terms are the ones which will puzzle the student more than anything else.

One also misses chapters on automation, nuclear energy, transistors and their influence on instrumentation. Space for this material could be found by omitting earlier chapters with their endless string of names and dates.

E. R. FRIEDLAENDER.

Digital Computing Systems

By S. B. Williams. 229 pp. 158 figs. Medium 8vo. McGraw-Hill Publishing Co. Ltd., New York, London. 1959. Price 60s.

MR. WILLIAMS was one of the earliest workers in the field of automatic digital computers, being the designer of the first truly automatic machine, the Bell System model 1 computer. This was a relay machine, rather than an electronic one, and throughout the book the author's greater familiarity with electromechanical devices is evident.

The publisher's claim that the book provides 'a broad technical introduction to digital computing systems, especially written for readers with engineering backgrounds . . .' is regretably hardly justified. It must be a major criticism of any book with such objectives that it fails to trace a clear path through the basic principles of its subject. By page 14, for example, after a short introduction to calculating machines (mainly semi-automatic desk types) operation of a computer is 'explained' using terms such as octal code, magnetic tape, matrix plane and Flexowriter.

Throughout the book the uninitiated reader is faced with similar difficulties, as well as by unbalanced treatment of detailed topics. For example, paper tape reading and punching equipment receives a lengthy description, in contrast with which the treatment of the Williams type of c.r.t. storage can only be called pathetic, especially when the elegance and lucidity of Prof. Williams' original papers is recalled.

An unduly large number of errors was also noted. Apart from mis-spelling of xerographic and styli, and the use of modulus 2 for modulo 2, it is stated that a magnetostrictive delay line operates by virtue of a magnetic wave travelling down the line, and that close packing and high tape speeds on a magnetic tape storage unit are primarily to give a large reading signal. Many other examples could be quoted.

The book does have the virtue of presenting a picture of contemporary American input/output and other electromechanical devices, but even here one is left with the impression of a miscellaneous collection of facts culled from manufacturers' literature.

The author frequently makes reference to the book 'High Speed Computing Devices' published in 1950 and has had in mind the production of a book supplementing this one and presenting the latest developments. The need for such an introductory work certainly exists, but the present work cannot be said to meet this requirement.

A. C. D. HALEY

Semiconductors and Transistors

By D. M. Warschauer. 450 pp. 177 figs. Medium 8vo. McGraw Hill Publishing Co. Ltd., New York, London. 1959. Price 50s. 6d.

THIS is a well-written book at an elementary level. Its main characteristic is the lucidity of the explanations given of basic solid state phenomena and of the characteristics of semiconductor devices. About one-third of the book is devoted to a largely non-mathematical exposition of the underlying principles of the conduction process in semiconductors. This section includes a chapter entitled 'Measurement of Semiconductor Characteristics'; Hall effect and cyclotron resonance measurements are discussed in addition to the usual measurements of resistivity and of lifetime and mobility of minority carriers. Chapters then follow on 'pn Junctions', 'Transistors' and 'Transistor Construction'. Apart from the final chapter on 'Non-Transistor Devices' which deals briefly with thermistors, photoconductive cells, etc., the remainder of the book is devoted to circuit aspects of transistors.

Small-signal analysis is essentially confined to the zero-frequency condition and introduces the resistance, conductance and hybrid parameters. The usual formulae are derived for current, voltage and power gain for the three methods of connexion. It is unfortunate that in figures 10.5 and 13.7 the common-base collector characteristics have been incorrectly drawn, particularly as figure 10.5 purports to refer to an 'actual transistor'. Criticism may also be made of section 12.5 dealing with 'Low-Frequency Response' in which the effect of the input resistance of the transistor in determining the small-signal emitter-base voltage is ignored. Also, the impression is given in section 10.10 that there is an abruptness associated with the cut-off frequency f_{α}, just as its unfortunately-chosen name implies. There is a chapter on 'Amplifiers' and another on 'Switches and Oscillators'. These are descriptive rather than analytical and serve to illustrate general principles.

There are no references to original work in the text. In a 'Selected Bibliography' the titles of twelve books on transistors and solid state physics are given. There are examples at the end of most chapters: answers are not included.

Despite the shortcomings referred to, this is a useful book. It can be recommended as a first reader on semiconductors.

F. J. HYDE

From Microphone to Ear

By G. Slot. 246 pp. 102 figs. Demy 8vo. Philips Technical Library, Eindhoven. Cleaver-Hume Press Ltd., London. 1959. Price 21s.

IN this book Mr. Slot's exceptional capacity for detail takes precedence over the more customary approach via examples of commercial equipment. It is written for the man who wants to know the exact reasons, quantities and relative values of fundamentals so commonly glossed over. An author has to have very special gifts to be able to present the elementary physical, electrical and musical considerations underlying the design of disk and tape reproducers in such a way that they appeal equally to the experienced engineer and the tyro; this book is not only very readable, it is fascinating.

After a historical introduction and a short description of the manufacture of disk records, Mr. Slot presents an excellent discussion on pick-ups, styli and tonearms. There is more detailed information in this section than in any other book so far to hand. Not everyone understands the design of shaded pole motors, turntables, drives or record chargers. These are lucidly explained. The section on amplifiers devotes as much space to undesirable properties as to their virtues, and there are some good circuits including a transformerless power amplifier.

The excellent chapters on loudspeakers, enclosures and listening rooms are well linked, and although wide range electrostatic loudspeakers are hardly mentioned the rest of the material is comprehensive in the extreme. Many measurements in these chapters, all of which have been carried out on Philips equipment, are illuminating. In a short review it is not possible to describe all the subject matter in detail, but the following are examples of some points of especial interest.

Response corrections for 23 different makes of record; graph of needle wear as a function of playing time; calculations for pulley diameters for different record and different motor speeds; degree of permissable 'wow' ranging through imperceptible to intolerable; permissable degree of intermodulation distortion in amplifiers; distribution of sound over a loudspeaker cone; critical loudspeaker frequency plotted against baffle size; chart for computing winding data for chokes in crossover networks; table relating top note loss to gap dimensions for various tape speeds; critical frequencies at which tape echo occurs for various tape thicknesses and speeds.

Stereophony is in the news; Mr. Slot deals with it very well, perhaps the most informative section being on pick-up data, well supported by illustrations. In chapter XI, the author explains the real meaning of high fidelity very neatly, considering that what means 'hi-fi' to one may merely be a source of irritation to another. This section ends with a list of records for testing various features of the listening system and some simple gear for measuring power output and frequency response.

The next chapter covers tape recording and here again, the author exhibits his unerring instinct for describing those details hardly ever to be found in this kind of book. The last chapter describes the techniques used during a recording session. Finally, an appendix gives details of an ingenious single-ended amplifier and some loudspeaker cabinets.

The book is one of the Philips Technical Library series and is published in English, German, French, Dutch, Spanish and Italian. The price is a remarkable illustration of value for money.

Mr. Slot is to be congratulated on his thorough, clear and at times entertaining handbook. It is a real encyclopaedia in miniature and does describe the cause and effect of all the little points, both good and bad, which (one feels) some authors consider beneath them. The style is refreshing and a tribute should be paid to the English translator, Mr. Harker.

A. DOUGLAS

Semiconductors

By R. A. Smith. 482 pp. 137 figs. Medium 8vo. Cambridge University Press. 1959. Price 65s.

THIS book is a welcome addition to the literature of solid state physics. The semiconductor field has grown so rapidly in the past few years that even in the 450 pages devoted here to the physics of semiconductors, some important topics have received only scant attention and some none at all. Dr. Smith has selected his material carefully, however, and has presented in coherent form a unified treatment of the subject. Although wave-mechanical considerations are introduced, the treatment generally used is based on the concept of mobile charged particles characterized by effective mass tensors. Some 350 pages are devoted to theoretical exposition. Topics covered include the three-dimensional energy band picture, impurities and imperfections, the statistical thermodynamics of carrier concentrations, transport phenomena including magneto-resistance, thermal effects and optical effects. It is pleasing to find a few pages devoted to thermomagnetic effects, which in many books receive only a brief mention in a concluding sentence at the end of a section dealing with Hall effect. The d.c. theory of pn junctions and of n^+-n and p^+-p junctions is presented for low injection-level conditions and brief attention is paid to high-frequency effects. It is a pity that a discussion of intermediate and high injection-level operation is not included as this is more common in practical devices. There is a chapter devoted to methods of determining the properties of semiconductors: principles rather than practices are stressed. Two subsequent chapters deal with the physical properties of elementary and compound semiconductors respectively. Finally, there is a short chapter on practical applications of semiconductors.

There is no doubt that this book will appeal primarily to the physicist. Parts of it will be useful for final year undergraduates, although it is more suitable as a text book for post-graduate courses. It is an excellent source book for specialist workers, both physicists and engineers. Although there are no detailed theories of the physics of operation of semiconductor devices in it, the basic material will provide an up-to-date background for the understanding of such theories. It is a book of quality which can be warmly recommended.

F. J. HYDE

Reports on Progress of Physics Volume XXII

Edited by A. C. Stickland. 634 pp. 188 figs. Crown 4to. The Physical Society, London. 1959. Price 63s.

THE survey of physics provided by these periodical reviews continues to be admirably well-balanced and impressively well-documented. Perhaps a wider than average range is covered in the present volume, articles on nuclear processes at one extreme being set in cosmic perspective by H. Bondi's lucid survey of 'Relativity' at the other, and J. Pollard's critical treatment of 'Progress in Vacuum Technology' reminding us of the debt owed by so much progress in 'purer' physics to that difficult art.

Apart from this latter paper, the electronic engineer may find himself immediately interested in 'Experimental Investigations of the Ionospheric E-Layer' by B. J. Robinson, and perhaps also 'Radio-frequency Spectroscopy of Excited Atoms' by G. W. Series.

Robinson's is one of the first surveys to include the recent access of information from the technique of rocket research. As such it is both especially interesting and especially vulnerable, since it is expressly intended only to bring the subject up to date 'before the analysis of the new data (from IGY material) begins.'

It may best be regarded therefore as an expert 'guide to form' for those who want to follow the ensuing race intelligently; but although the author concentrates particularly on the strengths and weaknesses of rival theories, he provides also a historical and theoretical setting and a body of established facts which the general reader—particularly the radio engineer—will find of more lasting value. Perhaps most surprising to the uninitiated reviewer was the great complexity of fine structure revealed in the E-Layer by high-resolution ionosondes.

G. W. Series's paper is concerned mainly with the results obtained by different methods of existing atomic resonances in the microwave and radio frequency range, rather than with the experimental techniques as such. Enough details are given, however, to indicate the ingenuity of experimentation in this field and the importance of the contribution now being made in this way by electronics to the study of atomic and molecular structure.

Among the remaining contributions special interest attaches to a Russian paper on 'Physical Problems of Thermo-electricity' by A. F. Joffe and L. S. Stil'bans. This deals chiefly, as might be expected, with the advance in knowledge of thermoelectric effects in semiconductors, which has been gained largely from work in the Soviet Union. Criteria of the efficiency of thermoelectric generators, refrigerators and heaters are carefully defined and evaluated. The reader is warned however (p. 188) that questions of the choice of materials and the production of optimum carrier concentration 'have been as yet very little studied;' and the slenderness of the bibliography attests the openness of this exciting field to further development.

Misprints are commendably few, apart from a persistent substitution of 'nuclear' for 'nucleon' which abruptly appears to haunt the titling of one paper for 18 pages before it is finally exorcised.

D. M. MACKAY

Mullard Circuits for Audio Amplifiers

136 pp. Crown 4to. Mullard Ltd. 1959. Price 8s. 6d.

THIS book has been prepared by the Technical Service Department of Mullard Ltd and covers 12 of the most popular Mullard circuits including the latest designs for stereophonic amplification. In addition to all the necessary circuit information, practical chassis layouts are suggested for each amplifier. Numerous diagrams and photographs are included to give the constructor the greatest possible guidance.

The book also contains much useful theoretical information, the first four chapters dealing with: considerations of monaural and stereophonic amplifying systems, the sources of distortion in recorded sound, high quality amplifier systems, and construction and assembly.

Symbolic Logic & Intelligent Machines

By E. C. Berkeley. 189 pp. 54 figs. Medium 8vo. Reinhold Publishing Corporation, New York. Chapman & Hall Ltd, London. 1959. Price 52s.

THE main purpose of this book is to make clear the nature of symbolic logic and the properties of intelligent machines and to show how these two subjects are related. It is also its purpose to explain the application of symbolic logic in the design of machines that behave intelligently. The design and principles of both small and large machines are included, along with specific problems confronting these machines.

Short News Items

Marconi's Wireless Telegraph Company announce that an agreement has been concluded between the Company and the Government of India (Ministry of Defence) for the manufacture under licence in India of v.h.f. multi-channel radio terminals and repeaters, and ancillary equipment. Under the terms of the agreement Marconi's are to supply all necessary technical assistance for indigenous manufacture. The v.h.f. multi-channel radio equipments to be manufactured in India are the HM100 series developed by Marconi's for use in terrain unsuitable for the construction of line or cable routes. They are designed to carry up to 48 telephone channels, any of which may be sub-divided to give either 18 or 24 telegraph channels.

The Timken Roller Bearing Company and The Fafnir Bearing Company of New Britain, Connecticut, U.S.A., have announced that an agreement has been concluded by which a recently formed subsidiary of the latter, Fafnir Bearing Company Ltd, Wolverhampton, purchased the business of Fischer Bearings Company Ltd, a Timken subsidiary. The sale of Fischer, which was until recently owned by the former British Timken Ltd, results from the Timken decision to confine their interest to the manufacture of Tapered Roller Bearings throughout the world. They will however, continue for the present, to distribute FBC bearings for replacement purposes in Britain, through their subsidiary Timken Stockists, and in Australia, South Africa and Canada through their branches in those countries.

One of the latest E.M.I. colour television cameras left the Hayes headquarters of E.M.I. Electronics Ltd recently *en route* for China. With the camera went control equipment and a Rank-Cintel large screen colour projector, which together form a complete closed circuit installation. It is the first colour television equipment to be exported to China by a British company. Designed specially for a wide range of industrial, medical and scientific applications, as well as for broadcast studio use, the colour camera incorporates several outstanding features. The camera is of an entirely new British design, and uses three vidicon tubes, and a novel optical system to give an improved quality even under difficult lighting conditions. It carries a lens turret with four positions, and the optical system has been designed so that the maximum amount of light falls on the photo-conductive surfaces of the vidicons. Notable for lightness and small size, it also has the advantage of being cheaper and easier to operate than the more elaborate Image Orthicon camera. Capable of producing broadcast quality simultaneous colour television signals on either 405, 525 or 625 line standards, the camera is suitable for use in compatible television systems.

The Salfords Research Laboratories of Mullard Ltd. recently held an exhibition of engineering technology associated with the development of new electronic components. Organized by the Engineering Division, the Exhibition was designed to show the many and widely differing techniques evolved in the production of the modern intricate electronic component. These included electro-chemical and photographic processes, the use of plastics both in modelling and fabrication, and the use of epoxy and polyester resins for encapsulation.

Telefunken 'Microport' transmitters were used by artists during performances of 'Die Czardasfürsten' in the recent Berlin Festival Week. Wearing concealed microphones, the principal performers were able to move freely on a stage of 4 500m^2 in area. In all, four sets working on two channels were used.

Mullard Limited are offering The Television Society a new yearly premium of £20 in order to encourage the presentation of original papers on various aspects of television. This premium will be awarded by the Council of The Television Society to the author of the best paper submitted during the year and subsequently published in their journal.

The British Group for Computation and Automatic Control of the British Conference on Automation and Computation held their Annual General Meeting on 17 December 1959. The Executive Committee was reconstituted for 1960 under the chairmanship of Mr. J. F. Coales, O.B.E., M.A. The Secretariat of the Group continues to be provided by The Institution of Electrical Engineers. The following is a revised list of the member organizations which, on and from 1 January 1960, constitute Group B:

Association of Certified and Corporate Accountants, British Computer Society, Chartered Institute of Secretaries, Institute of Actuaries, Institute of Bankers, Institute of Cost and Works Accountants, Institute of Fuel, Institute of Petroleum, Institute of Physics, Institution of Civil Engineers, Institution of Electrical Engineers, Institution of Mechanical Engineers, Institution of Production Engineers, Iron and Steel Institute, Office Management Association, Royal Aeronautical Society, Society of Instrument Technology. Observer: Department of Scientific and Industrial Research.

One substantial activity which Group B has undertaken, in collaboration with Group A, is the sponsoring of the British contribution to the First Congress of the International Federation of Automatic Control (I.F.A.C.) which is being held in Moscow from 27 June to 6 July 1960. As the number of papers submitted by British authors exceeded the time and space allocated by the organizers of this Congress, arrangements are being made also for a British Convention on this subject to be held in London in September 1960, at which the surplus of suitable papers will be presented. Further details concerning the I.F.A.C. Moscow Congress may be obtained from the Honorary Secretary, Group B, B.C.A.C., c/o The Institution of Electrical Engineers; the arrangements for the British Convention referred to above are in the hands of the Honorary Secretary, Group A, B.C.A.C., c/o The Institution of Mechanical Engineers by whom further announcements will be made.

The Semiconductor Division of the General Electric Company Limited and Semiconductors Limited of the Plessey Group of Companies, have entered into an arrangement whereby each technical sales force will handle information on the products of both organizations.

The third International Exhibition of Electronic Components will be held in Paris on 19-23 February 1960 at the Parc des Expositions at the Porte de Versailles. A floor area of 14 000m^2 has been provided for this Exhibtion which is an increase of 20 per cent on last year. The total number of exhibitors will be 400, of whom a considerable number will be from overseas. The Exhibition is being organized by the Fédération Nationale des Industries Electroniques of 23 Rue de Lubeck, Paris 17, from whom further particulars may be obtained.

The Society of Instrument Technology of 20 Queen Anne Street, London W.1, is to hold a one-day Symposium in London on 1 September 1960 on Rocket and Satellite Instrumentation. The papers proposed are as follows:

(1) Outline of the Investigations planned for the British Space Programme.
(2) A Design Study on a Communication Satellite.
(3) A Digital Data Reduction System for use in the static testing of Rocket Motors.
(4) Transducers for Rocket Motor Testing.
(5) The Measurement, Transmission and Recording of Data in the Skylark.
(6) Ultra-violet Measurements from Satellites.

The object of the Symposium is to indicate the fields of new scientific knowledge which have been opened up by experiments in the upper regions of the atmosphere and in outer space and to describe the kind of instrumentation which has either already been used or will be required for future research.

The Electronics and Communications Section of The Institution of Electrical Engineers, in association with the International Federation for Medical Electronics, is organizing the Third International Conference on Medical Electronics which will be held at Olympia from 21 to 27 July 1960. The Conference is planned to bring together members of the medical and electronic engineering professions so that each will gain a better understanding of the problems of the other. In conjunction with the Conference, the Institution is promoting an International Scientific Exhibition which will be held at Olympia at the same time as the Conference, and where the research organizations, universities, hospitals and industrial organizations from all over the world who are working in this important field, can display their latest developments. The exhibition is being organized by Industrial Exhibitions Ltd, 9 Argyll Street, London W.1. The Conference will be open to all interested persons, and those who would like to have Registration Forms and further particulars, or who are interested in submitting a paper should write to the Secretary, The Institution of Electrical Engineers, Savoy Place, London W.C.2.

A course on Induction Heating to be held in Cambridge on 1 to 2 March 1960 is being organized by the Process Heating Division of Pye Ltd. This course, which will be similar to the one held in December last, will cover the theory of induction heating and the design of coils for hardening, annealing, tempering, brazing and soft soldering. Further details of the course may be obtained from Mrs. E. Raeburn, Pye Process Heating, 28 James Street, Cambridge.

The twelfth National Aeronautical Electronics Conference, NAECON, will be held on 2-4 May 1960 at the Biltmore and Miami-Pick Hotels, Dayton, Ohio. The theme of this year's Conference is 'Electronics probes the Universe' and the technical papers portion of the programme will be limited to four simultaneous sessions. Exhibition space will be located in two hotels where aeronautical equipment concerns will exhibit their latest equipment designs. The Conference is being sponsored by the Dayton section of the I.R.E. and the Professional Group on Aeronautical and Navigational Electronics (PGANE). The Publicity Chairman is J. S. Hoorigan of 5387 Maple Cliff Ct, Dayton 15, Ohio.

The number of Television licences in Sweden in November 1959, reached 500 000 which corresponds to 66 sets per 1 000 inhabitants. There are at present 23 television transmitters in operation in the country, covering about 60 per cent of the entire population. A further 19 stations, plus two slave transmitters, will be added before 1 July 1960. This will bring the number of Swedes living in areas within the reach of television transmitters up to 5·3 million of the country's total 7·4 million population. A continued extension of the network is planned for the next few years and by 1964-65 some 90 to 95 per cent of the population will be covered.

Graduate Courses in pure and applied sciences are now being made available at the University of Birmingham from 1 October 1960. These courses are primarily intended for Honours Graduates of approved Universities and normally lead to either the degree of M.Sc. or the Diploma of Graduate Studies after a twelve-month course of full-time study. Subjects covered include Mathematical Physics, Reactor Physics and Technology, Information Engineering and Theory of Electrical Machines. Further particulars may be obtained from the Registrar, The University, Birmingham 15.

De Havilland Propellers Ltd. have formed a new group to apply their experience in System Engineering to the development of equipment for the automatic control of industrial processes and manufacture of all kinds. The Group has a great fund of practical knowledge to draw upon from within the Company in such diverse and complex fields as advanced electronics, measuring techniques, dynamics analysis, hydraulic and pneumatic engineering and servo systems. The Group will place particular emphasis on the design of integrated systems, and here the comparatively new techniques of System Engineering will be used, backed by digital computer investigations where necessary, to produce the optimum design at any level of complexity. There are now many industries which can make use of automation and the System Engineering Group is already holding discussions with the engineering, oil refining, chemical, food, sugar, brewing and paper industries, and many others. Use will be made of the advanced electronic techniques that have been developed by the Company. Digital systems will be used, when possible, to yield accuracy, simplicity and low cost. Most designs will incorporate transistors and ferrites, since these need little space or electrical power and present investigations on their reliability show them to be equal or superior to many older devices.

The death is announced of Philip Ray Coursey at his house at Headley Down, Hants, on 3 January 1960. P. R. Coursey, who was born in 1892, was educated at University College, London, and became assistant to Sir Ambrose Fleming at University College, London. During the first World War he was Admiralty Inspector of Wireless Telegraphy in H.M. Auxiliary Patrol and in 1919 became Technical Research Assistant at H.M. Signals School. He became Chief Engineer of the Dubilier Condenser Company Limited in 1923, a position which he held until 1931 when he was appointed Technical Director of the Company and retained that position with distinction until his retirement in 1957 when he was appointed Technical Consultant to the Company in addition to retaining his Directorship.

A new microwave link between Moncton and St. John in New Brunswick, Canada, and operating in the 2 000 Mc/s frequency band, was commissioned on 12 December 1959. This link is being used primarily for telephony transmissions, but can be switched over to transmit television signals, when required. The latest type of G.E.C. transmission equipment is employed, in which built-in metering facilities and plug-in panels are used to speed installation and minimize maintenance. Work is already in progress to extend the link from Moncton to Campbellton and from St. John to Halifax and Sydney. The equipment is to be manufactured by The General Electric Company of England and installed by Canadian General Electric Company for the New Brunswick Telephone Company and the Maritime Telegraph and Telephone Company.

For some three years Mining and Chemical Products Limited, of 86 Strand, London W.C.2, who are producers of the metals Bismuth and Indium, have been carrying on research in order to provide both these and other elements to a very high degree of purity suitable for the electronic industry. Their range covers Gallium, Arsenic, Tellurium and in addition the semiconductor compounds manufactured from these elements. Such materials as Bismuth Telluride and Indium Antimonide are now available in adequate quantities for use in the development of new electronic devices.

LETTERS TO THE EDITOR

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

Regulated Power Supplies

Dear Sir,—The comments by Thompson[1] on the article by Collins and Smith[2] point out the disadvantages of returning the pentode anode load to the unregulated supply voltage. In their reply, Messrs. Collins and Smith state their preference for increasing loop gain, rather than stabilizing the anode load return voltage, for alleviation of this problem.

I would like to draw attention to a particularly simple and effective method of stabilizing the anode load return voltage, which essentially eliminates the disadvantages in question. Though this method was originally described in connexion with transistor regulated power supplies[3], the principle is equally well applicable to valve circuits. The basic arrangement is shown in Fig. A,

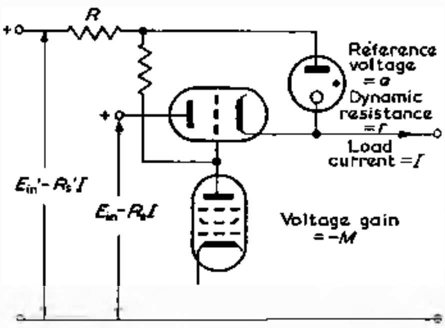


Fig. A. Basic arrangement of power supply

in which it is seen that the pentode anode load is fed from a voltage clamped to the regulator output by a reference tube. This circuit performs essentially as that of Collins and Smith's Fig. 5 is claimed to do, and is described by their equations. More detailed analysis shows that the equation for the regulator output voltage contains the additional correction terms

$$\frac{r}{R}\frac{1}{M}E_{in}' + \frac{1}{M}e$$

and that the equation for the output resistance contains the additional correction term

$$\frac{r}{R}\frac{1}{M}R'_s$$

where R_s' is a 'transfer' resistance defined as the change in the auxiliary supply voltage E_{in}' with change in the main supply current I.

These correction terms represent but a small degradation from 'ideal' performance. Further discussion, and simple methods of obtaining both supply voltages E_{in} and E_{in}' from the same power transformer winding, may be found in the original paper[3].

Yours faithfully,

R. D. Middlebrook,

California Institute of Technology,
Pasadena, California.

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Approximate Methods for Calculating the Behaviour of Square Loop Magnetic Cores in Circuit

Dear Sir,—In Mr. Brown's article "Approximate Methods for Calculating the Behaviour of Square Loop Magnetic Cores in Circuits" in the July 1959 issue, the switching of rectangular hysteresis-loop magnetic cores under various conditions is considered.

The basic equation arising from the domain wall motion concept is being presented in the article under discussion in the form:

$$\frac{dB}{dt} = S\,(H - H_c) \quad \ldots\ldots\ldots\ldots (5)$$

$\frac{dB}{dt}$ = the rate of the flux density change,
S = the switching constant,
H = the constant field provided by the step current,
H_c = the coercivity.

From the physical point of view, equation (5)* is true only in the case of some peculiar shapes of domains[1] and for switching times of order 1µsec or more, when the accelerating term in the domain wall differential equation[2] is negligible. Nevertheless, in the practical situations being considered by Mr. Brown, the utilization of equation (5)* for the prediction of the switching time of a core gives fairly good results. Therefore, equation (9) in Mr. Brown's article relating the switching time of a core resistively loaded with the parameters of the whole circuit, is true. This equation is, however, of rather theoretical interest and cannot be extended to the loaded conditions, in which the core under consideration is being used as a driver for other cores which are highly non-linear resistances with switching times depending on several factors.

The switching time of a core being loaded by other cores under constant magnetizing current conditions can be evaluated by introducing the equivalent resistance concept[3]:

$$R_{eq} = \frac{0{\cdot}8\pi\phi_s \cdot 10^{-8}}{S' \cdot l} \cdot N^2\,(1 - H_o/H)\,[\Omega] \quad \ldots\ldots\ldots\ldots (1)$$

where ϕ_s = the saturation flux of a core (Mx)
H = the magnetizing constant field (Oe)
H_o = the threshold field beyond which the irreversible changes of magnetization occurs, H_o nearly equals the coercivity H_c.
l = the main magnetic path length (cm).
N = numbers of turns.

The coefficient S' is being determined by the well-known relation:

$$S' = \tau\,(H - H_o) \quad \ldots\ldots\ldots\ldots (2)$$

τ = the switching time of a core.

This constant is related to Mr. Brown's S coefficient by:

$$S' = \frac{2\,\phi}{qS} \quad \ldots\ldots\ldots\ldots (3)$$

q = the cross-sectional area of a core.

For practical purposes one requires, of course, the switching time of the driven

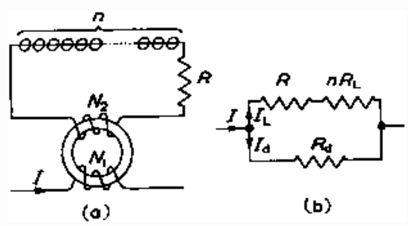

Fig. 1. Driven cores and equivalent circuit

cores (Fig. 1) to be not longer than that of the core being a driver. If one assumes, for simplicity, equality of both switching times and then takes into account the relations (1) and (2) for determining the equivalent resistances R_d (driving core) and R_L (driven core), and then considers the equivalent circuit given in Fig. 1(b), then:

$$N_{2(min)} = \frac{n\phi_{sL} + \dfrac{1{\cdot}25\;R\;S'\;l_L\;10^8}{\pi\;(1 - (H_o/H_L))}}{\phi_{sd}} \quad \ldots\ldots\ldots\ldots (4)$$

$N_{2(min)}$ = the number of secondary turns needed for the satisfactory operation of the whole circuit,
n = number of cores to be driven,
ϕ_{sd} = saturation flux of driving core,
ϕ_{sL} = saturation flux of driven core,
R = the whole series ohmic resistance in the driven circuit (Fig. 1),
H_L = the driving field of a core being the load,
l_L = the mean magnetic path length in the driven core. This formula gives the lowest number of

* *The formulas marked by the asterisk are the original as being presented in article discussed.*

secondary turns of driving core necessary for the driven core to be completely switched in the required time τ_L, which gives the numerical value of H_L by the use of equation (2). The whole magnetizing current needed can be determined in the same way:

$$I = \frac{2{\cdot}5\, H_L}{\pi N_1} [N_2 l_L + l_d] \quad [A] \ldots \ (5)$$

where: N_1 = number of primary turns,
l_d = the mean magnetic path length of driving core (cm),
H_L = the field previously determined (Oe).

It is of practical importance that the real ampere-turns applied should be taken somewhat larger than calculated from equation (5) with regard to safety margin.

As a matter of fact Mr. Brown has not given any formula useful for the calculation of equations (4) and (5) although he tended to solve the problem. I regret that with regard to saving of space the details of derivations must be omitted and the discussion must be shortened.

The problem of driving the cores from an imperfect current source can be solved by the application of Kirchhoff's law to the circuit given in Fig. 2:

$$V - n\, d\phi/dt - i\,R = 0 \ \ldots\ldots \ (6)$$

n = number of cores to be driven,
i = the driving current (instantaneous value),
R = the whole series resistance of the circuit,
V = the constant e.m.f. of the applied voltage source,
ϕ = the magnetic flux in a particular core (instantaneous value).

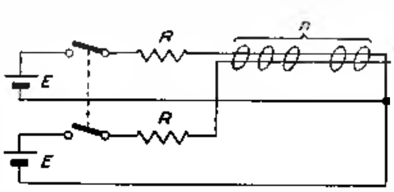

Fig. 2. Cores driven from imperfect source

Taking the general relationship arising from considerations performed by Papoulis and Chen[1] in the form:

$$S' = \int_0^{\tau} [H(t) - H_0]\, dt \ \ldots\ldots\ (7)$$

or:

$$S_1' = N \int_0^{\tau} [i(t) - I_0] dt \ \ldots\ldots \ (7')$$

N = number of turns, in the case under consideration $N = 1$,
$H(t)$ = the arbitrary magnetizing field being the function of time t,
$i(t)$ = the driving current related to $H(t)$ by well-known equations,
τ = the switching time of the core.

The switching time, τ, can be determined by integration of equation (6) and utilizing equation (7) for the case of double coincidence switching:

$$\tau = \frac{R S_1' + 4N\phi_s}{2V - I_0 R} \ \ldots\ldots\ldots\ldots \ (8)$$

If we substitute the numerical values from Mr. Brown's article, we simply obtain:

$$\tau = [1.12 + 0{\cdot}027\, n]\, 10^{-6}\text{sec} \ \ldots \ (8')$$

The numerical value of S_1' was evaluated by the use of data given in Fig. 9 of Mr. Kaposi's article[4].

It is worthy to note that equation (8) is indeed the same as equation (17) in Mr. Brown's article except for the term containing co-efficient S_1' which should be determined by the use of equation (3). Nevertheless, it can be easily proved that the switching times evaluated by the use of equation (8') are in fairly good agreement with those obtained experimentally and presented in Fig. 7(b) of Mr. Brown's paper, contrarily to results presented in the article. I believe therefore the discrepancy between the results obtained by calculation and those obtained by measurement cannot be due exclusively to a simplified theoretical approach. Maybe the switching constant S has been evaluated experimentally by a method not assuring the adequate accuracy of measurement. Nevertheless, it does not justify generally the simplifications made, where more correct relationships are known. The agreement between the equations based on simplifying assumptions and those obtained by the use of more correct methods may be sometimes casual; in the case under consideration it can be due to the averaging effect of integration performed.

Yours faithfully,

A. Goral,
Poland.

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3. Sands, E. A. The Behaviour of Rectangular-Hysteresis Loop Magnetic Materials under Pulse Conditions. *Proc. Inst. Radio Engrs.* 40, 1246 (1952).
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The Author Replies:

Dear Sir,—In reply to the three main sections of Mr. Goral's letter:—

(1) Mr. Goral gives no reasons to support his statement that his equation (3) is more fundamental than his (1). Papoulis and Chen start from (1) and this, with ψ(B)=constant, is my own starting point. The "constant" S' depends on the switching mechanism and is more generally a function of H.

(2) The constant S' of Goral's equation (3) depends on the core size and is not a true constant of the material. Therefore if cores of different size are used for switch core and storage core S' is different for these two cores and, as a result, Goral's equation (6), $H_d = H_L$ is not valid.

(3) Mr. Goral has concentrated on Fig. 8(b) and τ, and ignored Fig. 8(a). It is possible to make 8(b) a better fit by using a different value for S. This affects the intercept only (term P of equations (14) and (15) of my article) but not the slope (term Q). But by using such a different value the agreement of Fig. 8(a) will be destroyed.

Yours faithfully,

D. A. H. Brown,
Ministry of Aviation,
Great Malvern, Worcs.

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A Simple Time Marker

Dear Sir,—It is often desirable to use the mains supply as a timing marker when taking continuous photographic records of oscilloscope traces. At the higher film speeds a 50c/s trace is so elongated that it is difficult to determine the exact peak of each wave with certainty.

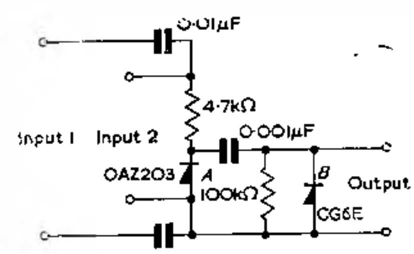


Fig. 1. Time marker circuit

The circuit shown in Fig. 1 will provide sharp pulses of 1/50sec intervals (Fig. 2 is from a film taken at 75in/sec). The mains supply is taken to input 1. The Zener diode (A) produces a square wave which is then differentiated to produce pulses of the form shown. Diode (B) is included to suppress the 'negative-going' pulses.

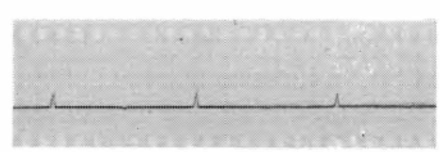


Fig. 2. Pulse unit output

Input 2 is included to allow phasing with a variable frequency supply if this is desired. The advantage of using a capacitor dropper with mains supply is to reduce heat dissipation in the circuit and to isolate the output from the mains.

The component types and values are not critical and can be chosen to suit, but the input capacitors should be able to withstand the supply voltage (1 000V d.c. types were used).

Yours faithfully,

J. J. Thomas and C. H. J. Daft,
The English Electric Company Limited,
Stafford.

ELECTRONIC EQUIPMENT

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

(Voir page 123 pour la traduction en français: Deutsche Übersetzung Seite 128)

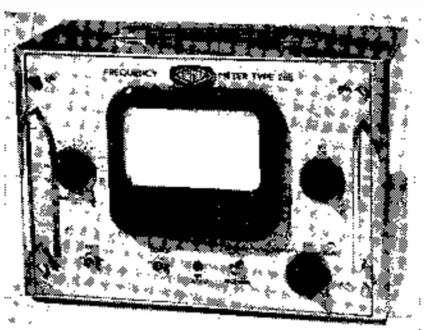
FREQUENCY METER

(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire

The frequency meter type 265 will measure the frequenecy of a repetitive waveform in the range 10c/s to 100kc/s to an accuracy of ±1 per cent. The input waveform is squared and limited, and applied to an integrating circuit, the voltage across which is indicated on a large, clear, panel meter. The circuits are so arranged that the meter reading is independent of the amplitude or waveform of the input voltage.

One application of the instrument is as an electronic tachometer. The speed of a rotating shaft may be measured directly by arranging that the shaft makes



or breaks a contact at each revolution and applying a voltage, thus interrupted, to the frequency meter. Alternatively, by employing the Airmec optical probe type 283 in conjunction with the frequency meter, detection of movement is carried out by reflection or interruption of a light beam, and speed measurements may be obtained without loading the equipment under test in any way.

The frequency meter is fully transistorized, light, and portable, and may be powered from a.c. mains or the internal batteries. Power supply and signal connexion for the optical probe are available at a rear socket. An internal calibration check is provided by means of a stable 5kc/s oscillator, and the meter is calibrated directly in frequency and revolutions per minute. An external meter or recorder may be fed from a socket on the front panel.

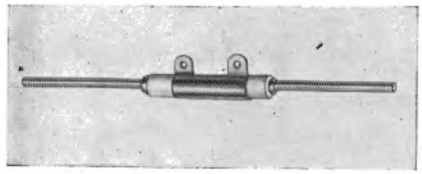
EE 18 751 *for further details*

SLEEVE TRIMMER CAPACITORS

(Illustrated above right)

Erie Resistor Ltd, 1 Heddon Street, London, W.1

In response to a demand from industry for an inexpensive stable low loss ceramic dielectric trimmer capacitor of simple robust construction and light weight, suitable for point-to-point wiring, on or between tag boards, across wafer



switches, and so on, Erie have developed type CT.20, which fulfils these requirements, and is suitable for a diversity of applications. The capacitance range covered is 1pF to 6pF, and adjustment is by positioning the sleeve with an insulated trimming tool the sleeve being provided with lugs for this purpose.

Noteworthy features are that both electrodes can be non-earthy—since there is no obligatory earthy mounting arrangement—that there is scope for mechanical coupling, by virtue of the design of the lugs provided for adjustment, and that there is also scope for printed circuit use, due to the light weight and wire terminations.

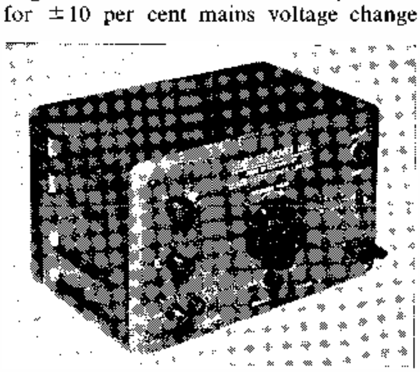
EE 18 752 *for further details*

STABILIZED L.T. POWER UNITS

(Illustrated below)

Lexor Electronics Ltd, 25 Allesley Old Road, Coventry

Two new miniature transistorized stabilized power units have been announced by Lexor Electronics. These are the L.T.12-05 and LT.24-05 having output voltage ranges of 5·5 to 14V and 18 to 28V respectively at output currents of 0 to 500mA d.c. floating. Output voltages are continuously variable over the stated ranges by either knob or pre-set control. Regulation is better that ±0·15 per cent for ±10 per cent mains voltage change



and/or zero to full load. Basic noise and ripple is less than 2·5mV peak-to-peak and the units are continuously rated for an ambient working temperature of 40°C.

The equipments are fully transistorized and mounted in enclosed miniature instrument cases. Mains input can be adjusted to suit 200 to 250V, 50c/s—units are fitted with mains and output fuse protection and voltmeter monitor sockets.

EE 18 753 *for further details*

ELECTRONIC ENGINEERING

will occupy a stand at

LE SALON DE LA PIÈCE DÉTACHÉE

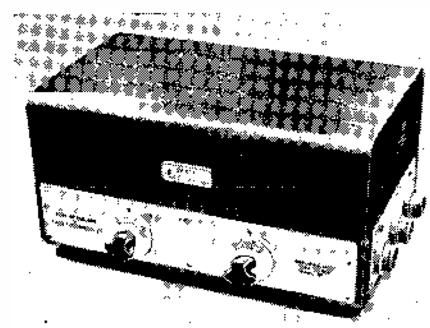
to be held in Paris from 19 to 23 February, and visitors will be welcome

PULSE AMPLIFIER

(Illustrated below)

Ekco Electronics Ltd, Southend-on-Sea, Essex

Ekco Electronics have announced a



new general-purpose high-stability pulse amplifier, type N640, with a wide frequency response. It is designed for use with scintillation counters without built-in amplifiers and will drive ratemeters and scalers requiring an input of 5V or more.

A six-position gain selector is incorporated and the low impedance output is positive-going for a negative-going input. A simple modification enables negative-going output to be obtained if so desired. Power supplies are obtained from the associated ratemeter or scaler. It has a gain approximately constant from 50c/s to 200kc/s and the upper half-power frequency is 1Mc/s.

EE 18 754 *for further details*

LONG PLAYING TAPE RECORDER

(Illustrated above right)

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

The Paraphone is designed to operate unattended indefinitely, to give a total of 40 hours recording or replay before repeating. A tape speed of 3¾in/sec has been selected to give a frequency response of 60c/s to 6kc/s within the 3dB points and 5¾in reels containing 1200ft (365m) of thin based tape give 1 hour's recording per track. In order to achieve 40 hours a two track head is used on a cam assembly giving 20 positions for the heads on a 1in tape. The

method of operation is such that 20 tracks of recording are made with one head and then a further 20 tracks are made with the second head.

Two capstan assemblies with counter-rotating flywheels are driven from a single capstan motor and move the tape in the direction appropriate to the selected track—even numbered tracks having the tape travelling from left to right across the heads and odd numbered tracks from right to left. A photo-electric cell actuates solenoids to change tape direction and move the heads to the next track.

A feature of the machine is the auto-start position which causes the machine to fast spool to the correct end of the selected track, then to reverse and play that track.

Push-buttons give start, fast spooling in either direction and stop in addition to the auto-start position. Operating the 'start' button will cause the machine to start normally from the position at which it is stopped.

A three-position switch on the deck gives, on a replay only machine 'off', an output of 1mW at 600Ω and a 2W output at any desired impedance. In the latter position an internal speaker can receive the full 2W output or give a low output for monitoring purposes. If the power is removed from the machine it cannot be restarted until this switch is returned to the off position and then reset to either of the other two positions. On a record-replay machine there is a four position switch, giving in the fourth position 'record'.

All the electronic circuits are transistorized and supplied via a transistor voltage stabilizing unit. The machine can be supplied for operation from a 28V d.c. supply (for aircraft use, etc.), or, by the addition of a separate power supply unit, from any standard mains supply.

EE 18 755 *for further details*

STANDARD SILICON SOLAR CELLS

(Illustrated above right)

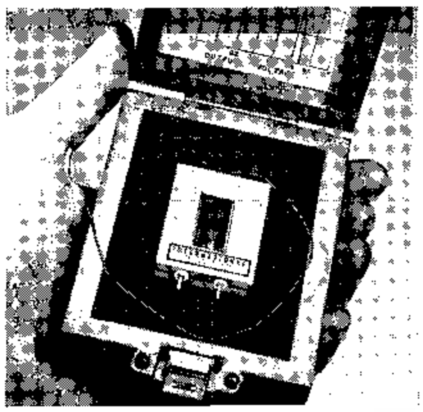
International Rectifier Co. (Great Britain) Ltd, Hurst Green, Oxted, Surrey

To provide accurate radiation measuring standards with which to check the efficiencies of production silicon solar cells, the International Rectifier Co. announces the availability of secondary standard silicon solar cells.

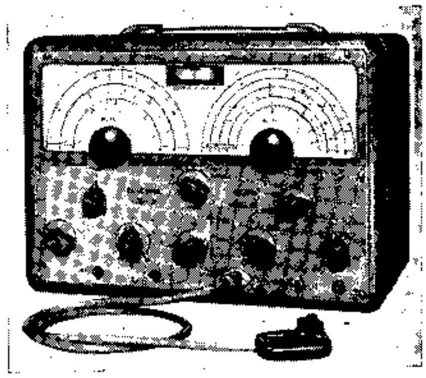
These cells are specifically designed for calibration of artificial light sources in terms of solar energy radiation. They may be considered to be the solar cell analogue to the pyroheliometer standard. Manufacturers using them will be able to evaluate incoming solar cells for efficiency, without the need of a pyroheliometer or other complex light-measuring device.

Each cell is selected to be of at least 9 per cent conversion efficiency, and is accurately calibrated to its exact efficiency rating. Before each cell is packaged, a calibration curve is made showing the efficiency of the unit at certain radiation intensities and specified temperature conditions. This calibration curve is mounted permanently to the instrument case, and accompanies the standard cell to the customer.

These calibrated cells are mounted in a rugged, shock resistant housing and protected by coated optical glass windows. They have thermocouple leads attached for measurements of cell temperature.



EE 18 756 *for further details*



A.M.-F.M. SIGNAL GENERATOR

(Illustrated above right)

Taylor Electrical Instruments Ltd, Montrose Avenue, Slough, Buckinghamshire

This signal generator is designed mainly for the servicing of a.m. and f.m. high frequency receivers and the i.f. stages of television receivers. The 61A provides, in conjunction with an oscilloscope complete facilities for the sweep alignment of the r.f., i.f. and discriminator or ratio detector stages of a.m. and f.m. receivers. The speedy and accurate method of ratio detector alignment using simultaneous a.m. and sweep signals is available.

The a.m. generator covers the range 4 to 120Mc/s in 5 bands, all on fundamentals. Calibration accuracy is ± 1 per cent and modulation depth 35 per cent at 400c/s. The f.m. and sweep section also operate on fundamentals, avoiding the confusion and poor frequency accuracy inherent in the frequency difference method. The f.m. and sweep bands cover frequencies from 4 to 12Mc/s and 70 to 120Mc/s in three bands. Calibration accuracy of the f.m. generator is ± 2 per cent. The a.m. generator may be used as a marker during sweep alignment. A variable phasing network is incorporated and retracing blanking is provided.

Deviation on f.m. is variable up to 100kc/s from the mean carrier frequency at a modulation rate of 400c/s. The sweep covers a total bandwidth of 1Mc/s at power line frequency. A crystal calibrator circuit with switch selection of any one of three internally mounted crystals, is incorporated. Crystals which may be selected in the frequency range 1 to 11Mc/s can be supplied.

All the facilities offered in the 61A model can be selected by internal switching, no cross connexions external to the instrument being necessary. There is choice of c.w., a.m., f.m., sweep plus c.w. or sweep plus a.m. by internal switching. When using the sweep facility, the marker level may be continuously varied.

A five-section attenuator with a non-inductive potentiometer provides attenuation over the whole frequency range. Radio frequency leakage has been reduced to a minimum by careful shielding of the attenuator and oscillator sections.

The new Taylor edgewise meter is used for monitoring the r.f. output.

An added facility in the servicing of audio stages is provided by a variable level 400c/s signal.

EE 18 757 *for further details*

ANTI-STATIC FLUID

Sionic Ltd, Cromwell House. Wood End Green Road. Hayes, Middlesex

The Sionic anti-static fluid has been developed for use in all fields where dust pick-up through static attraction is a major problem. It is stated that test results have shown that both positive and negative charges exist on the surface of moulded plastics. The highest charges are of positive polarity and are created by separation of the moulding from the metal mould. The charges of lower potential may be either positive or negative and may be the result of mouldings or hot processings, fabricating, wrapping, polishing or simply normal handling and storing.

Air-borne dust is attracted to and is held by these electrical charges. There appears to be no relation between dust pick up and the size, shape or flow pattern of any moulding. The type of mould lubricant used and the method of production, however, do seem to influence dust pick up.

The life of the anti-static coating under normal conditions is approximately one year, or more. The coating itself is a mono-molecular layer, which means it is so thin that it cannot be measured by ordinary means. It is completely transparent and in many cases renders the treated material more pleasant to the touch.

The fluid is equally suitable for use on cathode-ray tube screens and meter glasses and is available either in a 16oz Aerosol or in bulk form for industrial use.

EE 18 758 *for further details*

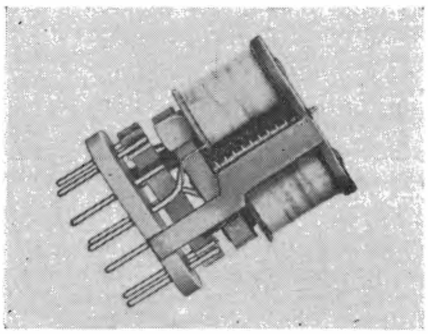
SUB-MINIATURE RELAY

(Illustrated below)

Distributed in U.K. by: Counting Instruments Ltd, 5 Elstree Way, Boreham Wood, Hertfordshire

This hermetically sealed relay, type RZO, is manufactured by Telefon Fabrik Automatic A/S, Copenhagen, Denmark. It measures only $\frac{3}{4}$in × $\frac{3}{4}$in × $\frac{1}{4}$in yet is rugged enough to withstand extremes of temperature, heavy shocks and severe vibration.

The RZO is a double coil unit with two changeover contacts, operated by a rotary balanced armature. The contacts are made of gold plated hard silver.



To eliminate the risk of contact contamination caused by gassing from the coil forms, these are made from a selected ceramic material.

All connexions in the relay, except inside the coils are welded and thus resistant to heat, shocks and vibration. The standard versions are fitted with hook terminals or plug-in type base.

Before sealing, the relay is evacuated under heat and filled with dry nitrogen.

The illustration shows the relay removed from its container.

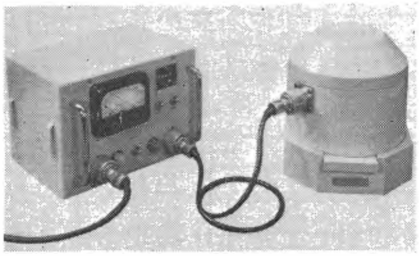
EE 18 759 *for further details*

CONTAMINATION MONITOR

(Illustrated above right)

Plessey Nucleonics Ltd, Weedon Road, Northampton

A new general purpose monitor, with a range of accessories permitting considerable versatility, has been developed by Plessey Nucleonics. In addition to the monitoring of radioactive contamination, its applications include the assay of solids and measurement of activity in large or small samples of liquid. It is particularly suitable for such high sensitivity measurements as are needed for the



assessment of radioactivity in evaporated water samples.

The monitor, type PNI 1065, is fully transistorized. It is mains operated for normal factory, hospital or laboratory use, but there is alternative provision for running from a 12V battery. E.H.T. is adjustable over a range suitable for most Geirger Müller and scintillation counting applications; a scintillation counter pre-amplifier is included.

The count rate is indicated on a rate-meter with a range of from 0 to 1 000 counts/sec; for low rates there is an electromechanical count register with a maximum operating speed of 25 impulses/sec. The range of accessories being produced for use with the monitor includes lead castles, a liquid probe and a beta/gamma hand probe.

EE 18 760 *for further details*

PARAMETRIC AMPLIFIER DIODE

(Illustrated below)

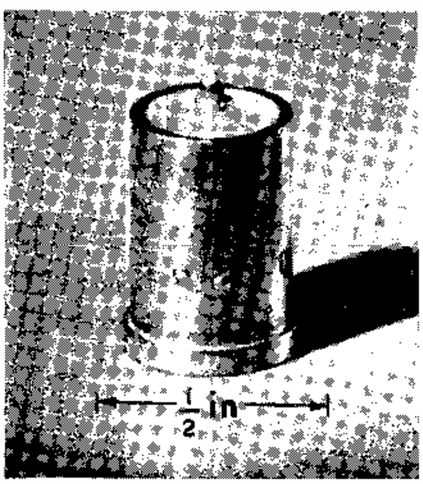
The General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

The EW97/1, a variable capacitance diode designed for use in parametric amplifiers at frequencies up to S-band, is now in production by G.E.C. The company claims that it is the first device with these qualities to be made available in the U.K.

The diode is suitable for use in radar and other communications systems. Designed for high power dissipation, it needs no elaborate precautions against surges.

It is mounted in a coaxial structure for direct insertion into coaxial and waveguide circuits and it has the low series inductance of 0·5mμH. A reverse polarity version (EW98/1) is also available.

Because of its very low forward impedance and very high reverse impedance,



the device can be used as a microwave switch. Another application is as a frequency multiplier—an important aid towards the future design of microwave receivers based entirely on solid state devices.

The inter services prototype references are: VX3314 (EW97/1), and VX3333 (EW98/1).

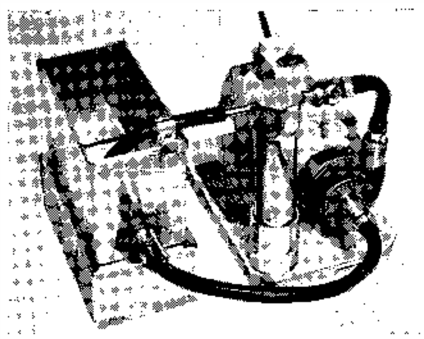
EE 18 761 *for further details*

ENCLOSED RELAYS

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

A range of small, robust d.c. relays has recently been announced by Radiospares Ltd. The relays are fitted with d.p.c.o. solid silver contacts rated at 250V, 2A a.c. (non-inductive). The connexions are to a tag panel mounted on top of the cadmium plated steel case. The tags, together with the connexions to the coils are protected by a varnish-flux against corrosion and to ensure easy tinning. The relays have moulded bobbins with varnish impregnated windings with a resistance of either 1kΩ or 12kΩ for nominal 12V or 50V operation. The relays are intended primarily for general laboratory or industrial use.

EE 18 762 *for further details*



ULTRASONIC CLEANER FILTER

(Illustrated above)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

Due to the effectiveness of ultrasonics, the cleaning solution becomes loaded with suspended dirt and has to be renewed at frequent intervals, depending on the state of the components being cleaned. Apart from the time lost in changing the solvent, the repeated renewal of solutions which are still active can be very expensive.

The type 1181 filter unit has therefore been introduced by Dawe Instruments Ltd to provide for the continuous removal of solid matter from ultrasonic cleaning baths of their own or other design. Fluid is drawn from the bottom of the cleaning bath by a centrifugal pump and returned to the top of the bath by way of a filter (centre of illustration) having a robust sintered element of stainless steel, which is capable of removing all solids down to a size of below 2 microns from 8 gallons of liquid per hour. Seven coarser filter elements can be substituted to give flow rates as high as 60 gallons per hour. Changing elements either for cleaning or to obtain a different pore size is effected in less than a minute by unscrewing the cylindrical filter housing from its cap to disclose the filter element, which can then itself be unscrewed and removed. The filter housing is conveniently held in a spring clip on the wooden base board.

All metal parts in contact with the cleaning fluid are of stainless steel. The interconnecting tubing is of tough, extruded nylon. All parts are thus easily cleaned and, if necessary, sterilized.

Larger filters are available which can also provide for temperature control of the cleaning fluid. Although designed primarily for use with ultrasonic cleaners, the new filter units are well suited for other filtration applications.

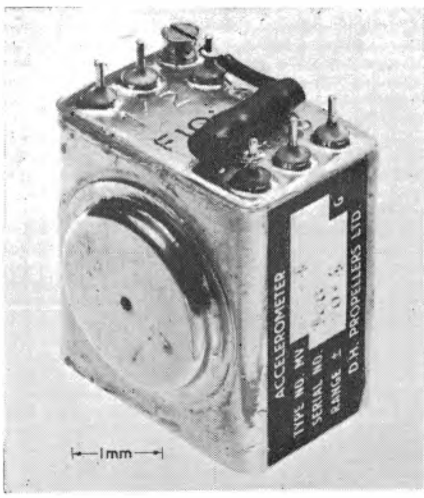
EE 18 763 *for further details*

ACCELEROMETER

(Illustrated below)

De Havilland Propellers Ltd, Hatfield Aerodrome, Hatfield, Hertfordshire

The MV300 accelerometer is the result of a development programme aimed at



the production of a high-performance instrument of low cost and size. The MV300 is the third generation in a series of designs, on each of which considerable production and operational experience has been acquired. A feature of this instrument is its high sensitivity of 25V full scale output. A selection of models is available with sensitivities ranging from 0·5V/g to 8V/g when working into a 0·5MΩ load, with natural frequencies from 34 to 137c/s.

The seismic mass of the instrument is a laminated armature which forms the moving part of a differential transformer a.c. pick-off. The armature is supported by a leaf spring system, geometrically similar to Watt's straight line mechanism, which ensures rectilinear motion for the centre of gravity and virtually eliminates cross coupling errors and the effects of transverse accelerations. The overall hysteresis is less than 0·7 per cent of the amplitude and cross coupling errors are less than 0·2 per cent. Stops, capable of withstanding very high shock loads and sustained overloads of the order of 100g are fitted.

Being an inductive device the resolution errors are nominally zero. An internally mounted demodulator and filter unit, potted in epoxide resin, is fitted and provides a d.c. output voltage with low ripple content.

The accelerometers have a three-point mounting which prevents distortion of the case due to attachment to uneven surfaces, and the instrument frame is attached to the case by a further strain-free mounting as an added precaution. The units are hermetically sealed in a cover which uses only soldered and glass-to-metal seals. They are oil filled for damping purposes, and a flexible metal diaphragm is provided to allow for thermal expansion. Silicone oil is used as a damping medium on account of its comparatively low temperature coefficient of viscosity.

The instrument is intended for use between −60° to +100°C, but is tested beyond these limits during manufacture to determine the stability of the zero error. The maximum zero error is 1·5 per cent of full scale and its maximum temperature coefficient is 2·5mV/°C. Normally a sensitivity tolerance of ±5 per cent is adequate, but for special purposes this figure can be reduced.

Provision is made for the adjustment of both zero error and sensitivity externally should the instrument suffer accidental damage; this simplifies repair work, the case being a completely soldered unit. These adjustments are performed by changing the values of miniature fixed resistors.

EE 18 764 *for further details*

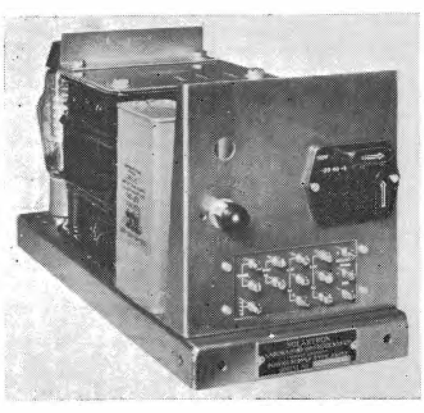
CHOKELESS POWER SUPPLY UNITS

(Illustrated below)

The Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

Designed specifically for continuous duty pulse loading applications, the new range of chokeless power supply sub-units announced by the Solartron Electronic Group can withstand pulsing from zero to full load without loss of control.

This attribute has been achieved by use of entirely new control circuits which, while maintaining a stability factor of better than 400 to 1, enables the very low source impedance to be held up to a frequency of 0·5Mc/s.



The range comprises three sub-units of 100mA maximum load capacity (AS 951, 952, 953) and three of 200mA (AS 955, 956, 957), output voltages covering from 100 to 400V. The voltage of any sub-unit can be varied over ±50V from nominal, also reactive loads may be applied, without affecting the specified performance.

In common with other Solartron power supplies, this chokeless series is conservatively rated throughout and provides the normal complement of a.c. heater supplies, both centre-tapped and untapped.

EE 18 765 *for further details*

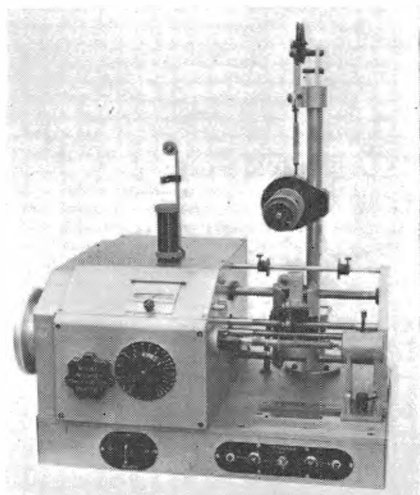
COIL WINDING MACHINE

(Illustrated below)

Westool Ltd, St. Helens Auckland, Co. Durham

A new bench type winding machine complete with meter and drive, and a total dimension of approximately 24in long by 24in wide is now being manufactured by Westool Ltd.

The machine is fitted with a five-figure



re-set revolution counter and can be operated up to a maximum speed of 4 000rev/min. The rate of acceleration from zero to maximum speed is determined by means of a plastic handwheel on the front of the machine which operates a potentiometer controlling the Warner electric clutch coupled to the motor.

Coils can be wound up to 5in in diameter and 3in in length in gauges from 30 to 50 s.w.g. The reel carrier spindle is suitable for reels up to 6in diameter but for the winding of fine gauges from 47 to 50 s.w.g. a small 'over end feed' is also supplied.

The traversing mechanism incorporates two Warner electric brakes coupled to two worm wheels in constant mesh with two lead screws. The traverse is set by means of adjustable stops which actuate microswitches energizing and de-energizing the brakes giving positive and instantaneous reversal of direction.

The wire spacing (turns per inch) is set by means of a handwheel moving against a calibrated scale engraved in s.w.g. sizes or in decimals of an inch.

EE 18 766 *for further details*

Meetings this Month

THE BRITISH COMPUTER SOCIETY LIMITED

Date: 18 February. Time: 6.15 p.m.
Held at: Northampton College of Advanced Technology, St. John Street, London E.C.1.
Lecture: The Development of the Use of a Computer for Production Control.
By: R. G. Hitchcock.

Belfast Branch

Date: 11 February. Time: 7.15 p.m.
Held at: 29 University Square, Belfast.
Lecture: The Computer in a University.
Nomination by: Professor Bates.

Glasgow Branch

Date: 1 February. Time: 6.45 p.m.
Held at: Royal College of Science & Technology, Glasgow.
Lecture: Problems Involved in Applying a Computer to Commercial Work.
By: H. W. Gearing.

Hull Branch

Date: 10 February. Time: 7.30 p.m.
Held at: Hull Chamber of Commerce, Samman House, Bowlalley Lane.
Lecture: "R.N.I.E."—Random Number Indicator, as used with Premium Savings Bonds.
By: W. E. Thomson.

Leicester Branch

Date: 11 February. Time: 7 p.m.
Held at: Room 104, Leicester College of Technology.
Lecture: Initial Experience with a National-Elliott 405.
By: J. G. Grover.

Liverpool Branch

Date: 23 February. Time: 6.30 p.m.
Held at: Mathematical Institute, The University, Liverpool.
Lecture: The Payroll Application at Newton Chambers & Co. Ltd.
By: J. F. Body.

Manchester Branch

Date: 9 February. Time: 7 p.m.
Held at: Manchester College of Science and Technology, Sackville Street.
Lecture: Production Planning by Computer.
By: R. B. Baggett.

Middlesbrough Branch

Date: 2 February. Time: 7.30 p.m.
Held at: Constantine Technical College, Borough Road, Middlesbrough.
Lecture: Computer Techniques for the Recording of, and Reference to, Data.
By: Dr. A. S. Douglas.

Newcastle Branch

Date: 2 February. Time: 7 p.m.
Held at: University Computing Laboratory, 1 Kensington Terrace, Newcastle.
Lecture: Data Transmission in Relation to Computers and Data Processing Systems.
By: W. S. Ryan.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1, commencing at 6.30 p.m.

Computer Group

Date: 9 February.
Lecture: Drift Correction of d.c. Amplifiers.
By: D. Leighton Davies.

Medical Electronics Group

Date: 24 February.
Lecture: The Unification of Electronic Clinical Instruments.
By: Dr. F. D. Stott.

South Midlands Section

Date: 2 February. Time: 7 p.m.
Held at: Winter Gardens, Malvern.
Lecture: Electronics in Medical and Biological Research.
By: W. J. Perkins.

South Wales Section

Date: 10 February. Time: 6.30 p.m.
Held at: Welsh College of Advanced Technology, Cardiff.
Lecture: Low Noise Low Drift d.c. Amplifiers.
By: V. L. Devonald.

South Western Section

Date: 24 February. Time: 7 p.m.
Held at: School of Management Studies, Unity Street, Bristol.
Lecture: Industrial Magnetic Recording and Playing Machine Design.
By: J. Elliot.

West Midlands Section

Date: 10 February. Time: 7.15 p.m.
Held at: College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Electronic Reading.
By: I. W. Merry.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Electronics and Communications Section

Date: 8 February.
Discussion: Present Views on Ground-Wave Propagation.
Opened by: G. Millington.
Date: 24 February.
Lecture: Applications of Microwaves.
By: Professor A. L. Cullen.

Medical Electronics Discussion Group

Date: 5 February. Time: 6 p.m.
Discussion: Computers in Medical Use.
Opened by: R. A. Buckingham and Dr. J. M. Tanner.

Irish Centre

Date: 18 February. Time: 6 p.m.
Held at: Physical Laboratory, Trinity College, Dublin.
Lecture: Television Systems.
By: H. DeLacy.

North-Eastern Centre

Date: 3 February. Time: 7 p.m.
Held at: Conservative Club, Newcastle-on-Tyne.
Joint Meeting with the Illuminating Engineering Society.
Lecture: An Outline of Television Studio Lighting Problems.
By: R. W. Koplick.
Date: 8 February. Time: 6.15 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-on-Tyne.
Lecture: Silicone Electrical Insulation.
By: J. H. Davis.
Date: 22 February. Time: 6.15 p.m.
Held at: As above.
Lecture: The Characteristics and Protection of Semiconductor Rectifiers.
By D. B. Corbyn and N. L. Potter.

North-Eastern Measurement and Electronics Group

Date: 29 February. Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle-on-Tyne.
Lecture: Radio Aspects of the International Geophysical Year.
By: R. L. Smith-Rose.

North Midland Centre

Date: 25 February. Time: 6.30 p.m.
Held at: Y.E.B. Offices, Ferensway, Hull.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

North Midland Utilization Group

Date: 16 February. Time: 6.30 p.m.
Held at: Leeds and County Conservative Club, South Parade, Leeds 1.
Lecture: The Design of Electromagnetic Pumps for Liquid Metals.
By: D. A. Watt.

North-Western Centre

Date: 2 February. Time: 6.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: Digital Computer Developments at Manchester University.
By: T. Kilburn.

North-Western Electronics and Communications Group

Date: 10 February. Time: 6.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: Television Recording: A Survey of the Problems and the Methods Currently in Use.
By: J. Redmond.

North-Western Measurement and Control Group

Date: 16 February. Time: 6.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: Transistors—Their Theory, Manufacture and Application.
By: R. W. A. Scarr and R. A. Setterington.

South-East Scotland Sub-Centre

Date: 16 February. Time: 7 p.m.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Lecture: Assessment of Speech Communication Links.
By: D. L. Richards and J. Swaffield.
Date: 24 February. Time: 7 p.m.
Held at: As above.
Lecture: Ultra-Sound Image Camera.
By: C. M. Smyth and J. Sayers.

South Midland Centre

Date: 1 February. Time: 7.30 p.m.
Held at: Winter Gardens, Malvern.
Lecture: Electronic Aids to Banking and Commerce.
By: R. Feinberg.

South-West Scotland Sub-Centre

Date: 17 February. Time: 6 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow C.2.
Lecture: Survey of Stereophony.
By: T. Somerville.
Date: 23 February. Time: 6 p.m.
Held at: Royal College of Science and Technology, Glasgow.
Lecture: Ultra-Sound Image Camera.
By: C. N. Smyth and J. Sayers.

South Midland Electronics and Measurement Group

Date: 22 February. Time: 6 p.m.
Held at: James Watt Memorial Institute, Birmingham.
Informal Lecture: Silicone Transistors—Their Manufacture and Fields of Application.
By: J. T. Kendall.

Rugby Sub-Centre

Date: 3 February. Time: 6.30 p.m.
Held at: College of Technology and Arts, Rugby.
Lecture: Radio Aspects of the International Geophysical Year.
By: R. L. Smith-Rose.
Date: 17 February. Time: 6.30 p.m.
Held at: As above.
Lecture: Progress in Permanent Magnet Materials.
By: J. E. Gould.

Western Centre

Date: 8 February. Time: 6 p.m.
Held at: South Wales Institute of Engineers, Cardiff.
Lecture: The Development of Communication, Indication and Telemetering for the British Grid.
By: G. A. Burns, F. Fletcher, C. H. Chambers and P. F. Gunning.

THE INSTITUTION OF PRODUCTION ENGINEERS

East and West Ridings Centre

Date: 8 February. Time: 7 p.m.
Held at: Hotel Metropole, King Street, Leeds 1.
Lecture: Computer Control of Machine Tools.
By: M. Monk.

Midlands Centre

Date: 17 February. Time: 7 p.m.
Held at: James Watt Memorial Institute, Birmingham.
Lecture: The Application of Computers to Production Control.
By: R. G. Hichcock.

THE RADAR & ELECTRONICS ASSOCIATION

Date: 9 February. Time: 7.30 p.m.
Held at: Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: Transistors—Today and Tomorrow.
By: E. Wolfendale.

THE TELEVISION SOCIETY

All London meetings will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London W.C.2, commencing at 7 p.m.

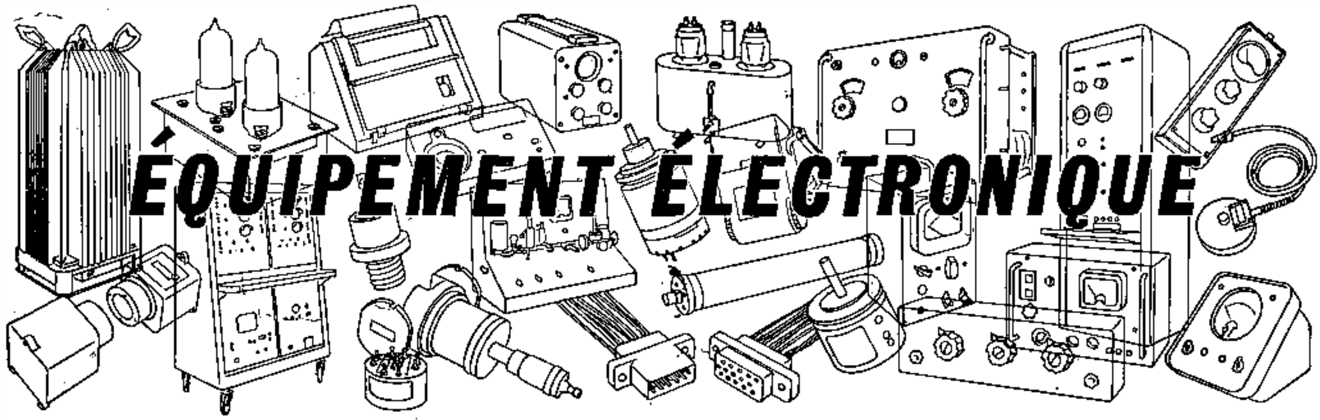
Date: 5 February.
Lecture: Automatic Control Systems in Television Receivers.
By: K. E. Martin and M. L. Mothersole.
Date: 25 February.
Lecture: Science on Television.
By: A. J. Garratt.

PUBLICATIONS RECEIVED

REPORT ON U.K. TRADE WITH SCANDINAVIA AND FINLAND summarizes hitherto unpublished talks which took place last May at a series of conferences between the F.B.I. and Commercial Representatives of British Embassies in Denmark, Norway, Sweden and Finland. The conferences held in London, Birmingham and Glasgow were organized to discuss the scope of the Scandinavian market for British goods. Federation of British Industries, 21 Tothill Street, London S.W.1. Price 4s. postage free.

PLATINUM METALS REVIEW is published by Johnson, Matthey & Co. Ltd, 78 Hatton Garden, London E.C.1, who are smelters and refiners of the platinum metals from the Rustenburg Platinum Mines of South Africa. The main article in the January issue surveys the most important of these applications and outlines some newer developments. Contributed by Dr. Eric Preston, the well-known consultant to the glass industry, the article emphasizes that the use of platinum equipment in the melting and working of glass has always resulted in an economy in manufacture, due entirely to the production either of a purer glass or of a glass free from defects of various kinds. Other articles in this issue deal with the palladium plating of printed circuits, the constitution and properties of the iridium-platinum alloys, and the behaviour of platinum-faced titanium and tantalum electodes for cathodic protection in fresh waters.

(Pages 123 to 132 French and German section)



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 118 à 121*

FRÉQUENCEMÈTRE

(Illustration à la page 118)

Airmec Ltd, High Wycombe, Buckinghamshire

Le fréquencemètre type 265 peut mesurer la fréquence d'une forme d'onde à répétition dans la gamme de 10 Hz à 100 kHz avec une précision de ±1%. La forme d'onde d'entrée est carrée et limitée, et elle est appliquée à un circuit intégrateur, dont la tension qui le parcourt est indiquée sur un grand panneau clair. Les circuits sont disposés de telle façon que la lecture du fréquencemètre est indépendante de l'amplitude ou forme d'onde de la tension d'entrée.

L'instrument peut servir, entre autres applications, de tachymètre électronique. La vitesse d'un axe tournant peut être mesurée directement en faisant commuter un contact par l'axe à chaque tour et en appliquant la tension, résultant de cette interruption, au fréquencemètre. En variante, on peut aussi, en utilisant le palpeur optique Airmec type 263 en relation avec le fréquencemètre, détecter un mouvement par la réflection ou l'interruption d'un faisceau lumineux. On peut, en outre, effectuer des mesures de vitesse sans surcharger aucunement l'équipement sous essai.

Le fréquencemètre est entièrement transistorisé, léger et portatif, et il peut être alimenté sur secteur c.a. ou par piles intérieures. L'alimentation et la connexion de signal pour le palpeur optique sont fournis par une prise arrière. Le contrôle d'étalonnage intérieur se fait par un oscillateur stable de 5 kHz, le fréquencemètre étant étalonné directement en fréquence et tours par minute. Un appareil de mesure ou enregistreur extérieur peut être alimenté à partir d'une prise sur le panneau avant.

EE 18 751 *pour plus amples renseignements*

CONDENSATEURS SHUNT D'ÉQUILIBRAGE

(Illustration à la page 118)

Erie Resistor Ltd, 1 Heddon Street, London, W.1

Pour répondre à la demande, de la part de l'industrie, d'un trimmer diélectrique céramique à faibles pertes, stable et peu coûteux, de fabrication simple et robuste ainsi que d'un poids réduit, propre au câblage point par point, sur plaquettes à cosses ou entre celles-ci, ou encore entre commutateurs plats, la société Erie Resistor a réalisé le trimmer CT20, qui remplit toutes ces conditions et convient à une diversité d'applications. La gamme de capacités s'étend de 1 pF à 6 pF et le réglage s'effectue en guidant la douille avec un outil d'équilibrage isolé, la douille étant pourvue à cet effet de pattes de fixation.

Le trimmer CT20 se distingue, entre autres, par le fait que les deux électrodes peuvent être non à la masse (puisqu'il n'y a pas de dispositif de montage à la terre obligatoire), par la possibilité de couplage mécanique, par le dessin des pattes se prêtant au réglage et, enfin, par la possibilité d'emploi de circuits imprimés, grâce au poids réduit et aux terminaisons des fils.

EE 18 752 *pour plus amples renseignements*

BLOCS D'ALIMENTATION STABILISÉE BASSE TENSION

(Illustration à la page 118)

Lexor Electronics Ltd, 25 Allesley Old Road, Coventry

Deux nouveaux blocs d'alimentation stabilisée à transistors viennent d'être présentés par la société Lexor Electronics Ltd. Ce sont les modèles LT-12-05 et LT-12-24-05, dont les gammes de tension de sortie sont de 5,5 à 14V et de 18 à 28V respectivement, à des courants de sortie de 0 à 500 mA c.c. flottant. On peut varier continûment les tensions de sortie au-dessus des gammes précitées, soit par bouton soit par commande préréglée.

Le réglage est supérieur à ±0,15% pour un changement de tension secteur de ±10% ou pour zéro à pleine charge. Le bourdonnement et l'ondulation de base sont inférieurs à 2,5 mV de crête à crête et les blocs sont prévus pour une température ambiante de service de 40°C.

Les blocs sont entièrement transistorisés et montés dans des coffrets fermés miniature. L'entrée secteur peut être ajustée pour 200 à 250 V, 50 Hz. Les blocs sont protégés par fusibles secteur et sortie ainsi que par des prises de contrôle de voltmètre.

EE 18 753 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera un stand au SALON DE LA PIÈCE DÉTACHÉE, qui se tiendra à Paris du 19 au 23 février 1960. Les visiteurs seront les bienvenus.

AMPLIFICATEUR D'IMPULSIONS

(Illustration à la page 118)

Ekco Electronics Ltd, Southend-on-Sea, Essex

La société Ekco Electronics vient d'annoncer la réalisation d'un nouvel amplificateur universel à grande stabilité, type N640, à réponde de fréquence étendue. Prévu pour l'emploi avec les compteurs à scintillations sans amplificateurs incorporés, il peut entraîner des compteurs et des appareils de mesure nécessitant une entrée de 5V et davantage.

Un sélecteur de gain à six positions est incorporé à l'amplificateur, et la sortie à faible impédance est à direction négative. On peut, cependant, sur simple modification, obtenir une sortie à direction négative, si nécessaire. L'alimentation s'obtient du compteur ou de l'appareil de mesure s'y rattachant. L'amplificateur a un gain approximativement constant de 50 Hz à 200 kHz et la fréquence supérieure à demi-puissance est de 1 MHz.

EE 18 754 *pour plus amples renseignements*

MAGNÉTOPHONE À ENREGISTREMENT PROLONGÉ

(Illustration à la page 119)

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

Le Magnétophone Paraphone a été étudié pour fonctionner indéfiniment sans qu'on s'en préoccupe et pour fournir un total de 40 heures d'enregistrement ou de jeu renouvelé avant de devoir répéter. Une vitessse d'avance de la bande de 95 mm/sec. a été prévue afin de donner une réponse de fréquence de 60 Hz à 6 kHz dans les limites des points de 3 dB, cependant que les bobines de 140 mm contenant 365 mm de ruban à base mince donnent une heure d'enregistrement par piste. Pour obtenir 40 heures, une tête à deux pistes est employée sur un assemblage à cames donnant 20 positions pour les têtes sur un ruban de 25,4 mm. Le mode de fonctionnement est tel que 20 pistes d'enregistrement sont effectuées avec une tête, suivies de 20 autres pistes avec la seconde tête.

Deux assemblages à cabestan avec volants à contre-tours sont actionnés à partir d'un seul moteur à cabestan et déplacent le ruban magnétique dans la direction appropriée à la piste choisie, même les pistes numerotées dont le ruban se déplace de gauche à droite à travers les têtes et les pistes à nombres impairs se déplaçant de droite à gauche. Une cellule photoélectrique met en mouvement des solénoïdes pour le changement de direction du ruban et le déplacement des têtes vers la piste suivante.

Une particularité intéressante du magnétophone en question est la position de marche automatique qui fait bobiner rapidement l'appareil jusqu'à l'extrémité voulue de la piste choisie, puis renverser la marche et jouer la piste précitée.

Des boutons-poussoirs permettent la mise en marche et le bobinage rapide dans l'une ou l'autre direction et s'arrêtent, en plus, à la position de marche automatique. Lorsqu'on presse le bouton 'marche', l'appareil se met en marche normalement à partir de la position à laquelle il est arrêté.

Un commutateur à trois positions sur le pont donne, sur un magnétophone à jeu renouvelé seulement, l'arrêt, une sortie de 1 mW à 600 Ω et une sortie de 2 W à n'importe quelle impédance voulue. Dans cette dernière position, un haut parleur intérieur peut recevoir la sortie entière de 2 W ou bien donner une sortie basse pour des besoins de contrôle. Si le courant est supprimé de l'appareil, on ne peut le remettre en marche que lorsque le commutateur est remis à la position "arrêt", puis fixé à nouveau sur l'une ou l'autre des deux positions. Sur un appareil à rejouer des disques, il existe un commutateur à quatre positions, donnant 'disque' à la quatrième position.

Tous les circuits électroniques sont transistorisés et alimentés au moyen d'un stabilisateur de tension à transistors. L'appareil peut être livré pour fonctionner sur 28 V c.c. (pour usage d'avion) ou, en ajoutant un bloc d'alimentation à part, sur n'importe quelle alimentation secteur courante.

EE 18 755 *pour plus amples renseignements*

CELLULES SOLAIRES ÉTALONS AU SILICIUM

(Illustration à la page 119)

International Rectifier Co. (Great Britain) Ltd. Hurst Green, Oxted, Surrey

La société International Rectifier vient de présenter des cellules solaires au silicium constituant des étalons précis de mesure de radiation.

Ces cellules ont été spécifiquement étudiées pour l'étalonnage de sources de lumière artificielle en fonction de la radiation d'énergie solaire. On peut les considérer comme étant des cellules solaires analogues aux étalons pyrohéliométriques. Les fabricants qui s'en serviront pourront évaluer l'efficacité de cellules solaires sans avoir recours à un pyroheliomètre ou à tout autre instrument compliqué pour mesurer la lumière.

Toute cellule choisie doit avoir au moins 9% d'efficacité de conversion et elle doit être étalonnée avec précision à sa valeur nominale exacte d'efficacité. Avante d'emballer une cellule choisie, on établit une courbe d'étalonnage qui indique l'efficacité de l'élément à certaines intensités d'irradiation et à des conditions de température spécifiées. La courbe d'étalonnage est fixée de façon permanente au coffret et elle suit évidemment la cellule étalon chez le client.

Ces cellules étalonnées sont logées dans un solide coffret à l'épreuve des chocs et elles sont protégées par des fenêtres optiques à revêtement. Elles sont pourvues de fils à couples thermoélectrique pour les mesures de température de cellule.

EE 18 756 *pour plus amples renseignements*

GÉNÉRATEUR DE SIGNAUX À m.a./m.f.

(Illustration à la page 119)

Taylor Electrical Instruments Ltd. Montrose Avenue, Slough, Buckinghamshire

Ce générateur de signaux est essentiellement prévu pour le dépannage des récepteurs h.f. à modulation d'amplitude/ modulation de fréquence, ainsi que pour les étages moyenne fréquence des récepteurs de télévision. Le modèle 61A permet pleinement, utilisé en relation avec un oscilloscope, l'alignement du balayage des étages h.f., m.f. et discriminateurs—ou à circuit de détection simplifié—des récepteurs à modulation de fréquence et modulation d'amplitude. La méthode rapide et précise d'alignement du circuit de détection simplifié par l'utilisation de signaux m.a. et de balayage est également applicable.

Le générateur m.a. couvre la gamme de 4 à 120 MHz en 5 bandes, sur toutes les fondamentales. La précision de l'étalonnage est de ±1% et la profondeur de la modulation est de 35% à 400 Hz. La modulation de fréquence et la section de balayage fonctionnent également sur les fondamentales, évitant la confusion et la faible précision de fréquence inhérentes à la méthode de différence de fréquence. Les bandes de modulation de fréquence et de balayage couvrent des fréquences de 4 à 12 MHz et de 70 à 120 MHz en trois bandes. La précision de l'étalonnage du générateur à modulation de fréquence est de ±2%. Le générateur à modulation d'amplitude peut servir de marqueur durant l'alignement du balayage. Un réseau de mise en phase est incorporé au générateur, qui prévoit la suppression du retour du spot.

La déviation sur modulation de fréquence est variable jusqu'à 100 KHz à partir de la fréquence porteuse moyenne à un taux de modulation de 400 Hz. Le balayage s'étend sur une largeur de bande totale de 1 MHz à la fréquence du secteur électrique. Un circuit à calibrage piézoélectrique, avec choix, par interrupteur, de l'un des trois cristaux montés intérieurement, fait partie de l'appareil. Les cristaux qui peuvent être choisis dans la gamme de fréquences de 1 à 11 MHz sont fournis sur demande.

Tous les avantages offerts par le modèle 61A peuvent être choisis par commutation intérieure, aucune connexion croisée extérieure n'étant nécessaire. On peut ainsi choisir, par commutation intérieure, les ondes entretenues, la modulation d'amplitude/ modulation de fréquence, le balayage plus ondes entretenues ou le balayage plus modulation d'amplitude. Lorsqu'on utilise le dispositif de balayage, le niveau de mesure peut être à variation continue.

Un atténuateur à cinq sections avec potentiomètre non-inductif assure l'atténuation sur toute la gamme de fréquences. Les fuites de haute fréquence ont été réduites au minimum par le blindage minutieusement étudié de l'atténuateur et des sections de l'oscillateur.

Le nouvel appareil de mesure de champ Taylor est utilisé pour le contrôle de la sortie haute fréquence.

Enfin, le signal de 400 Hz à niveau variable constitue un atout supplémentaire pour le dépannage d'étages basse fréquence.

EE 18 757 *pour plus amples renseignements*

FLUIDE ANTISTATIQUE

Sionic Ltd, Cromwell House, Wood End Green Road, Hayes, Middlesex

Le fluide antistatique Sionic a été mis au point pour l'emploi dans tous les domaines où la poussière ramassée par suite de l'attraction statique crée un problème d'importance majeure. Les résultats d'essais effectués par les fabricants ont prouvé qu'il existe des charges positives aussi bien que négatives à la surface de plastiques moulés. Les charges les plus élevées sont de polarité positive et sont créées par la séparation du moulage du moule métallique. Les charges de bas potentiel peuvent être soit négatives et résulter de moulages ou de traitements à chaud, de la fabrication, de l'emballage, du polissage ou, simplement, de la manipulation normale et de l'emmagasinage.

La poussière dans l'air est attirée et retenue par ces charges électriques. Il semble qu'il n'y ait aucun rapport entre la poussière ramassée et la grandeur, la forme ou le débit particulier d'un moulage. Le type de lubrifiant employé et le mode de production influent, en revanche, sur la poussière ramassée.

La durée du revêtement antistatique, dans des conditions normales, est d'environ un an ou davantage. Le revêtement lui-même est une couche monomoléculaire, c'est à dire, qu'elle est si mince qu'on ne peut la mesurer par des moyens ordinaires. Elle est entièrement transparente et, dans de nombreux cas, elle rend la matière traitée plus douce au toucher.

Le fluide antistatique convient aussi

bien à l'emploi sur des écrans de tubes cathodiques que sur des fenêtres de compteurs et il est fourni soit sous forme d'Aérosol de 0,46 litre soit en vrac pour l'usage industriel.

EE 18 758 *pour plus amples renseignements*

RELAIS SUBMINIATURE

(Illustration à la page 120)

Distributeurs dans le Royaume-Uni: Counting Instruments Ltd, 5 Elstree Way, Boreham Wood, Hertfordshire

Ce relais hermétiquement scellé, type RZO, est fabriqué par Telefon Fabrik Automatics A/S, Copenhaque, Danemark. Il ne mesure que 19 mm × 6,35 mm, mais il est assez robuste pour pouvoir résister aux températures extrêmes, aux chocs violents et aux fortes vibrations.

Le RZO est un élément à double bobinage avec deux contacts commutateurs, actionnés par une armature à équilibre rotatif. Les contacts sont en argent dur doré.

Afin d'éliminer tout risque de contamination des contacts, causé par un dégagement gazeux des formes de bobinages, ceux-ci sont en céramique spéciale.

Toutes les connections du relais, sauf à l'intérieur des bobines, sont soudées, donc résistantes à la chaleur, aux chocs et aux vibrations. Les modèles classiques sont munis de bornes à manche ou d'une base du type à fiches.

Avant de sceller le relais, on y fait le vide par la chaleur et on le remplit d'azote sec.

La gravure montre le relais hors de son coffret.

EE 18 759 *pour plus amples renseignements*

DÉTECTEUR DE CONTAMINATION

(Illustration à la page 120)

Plessey Nucleonics Ltd, Weedon Road, Northampton

Un nouveau détecteur universel, doté d'accessoires lui assurant une souplesse considérable, vient d'être mis au point par la société Plessey Nucleonics. En plus de la détection de toute contamination radioactive, ses applications s'étendent à l'essai des matières solides et à la mesure de l'activité dans des petits et des grands échantillons de liquide. Il est particulièrement indiqué pour les mesures de haute sensibilité, telles que celles qui sont nécessaires à l'évaluation de la radioactivité dans des échantillons d'eau.

Le détecteur type PNI 1065 est entièrement transistorisé. Il fonctionne sur courant secteur pour l'usage normal dans les usines, les hôpitaux et les laboratoires, mais il peut aussi être employé sur pile de 12 volts. La très haute tension est réglable dans une gamme convenant à la plupart des applications à compteur de Geiger et à compteur à scintillation; il comporte d'ailleurs un préamplificateur pour compteur à scintillation.

Le régime de comptage est donné par un compteur à gamme de 0 à 1000 coups/seconde. Pour les régimes de moindre envergure, l'appareil est pourvu d'un enregistreur électromécanique dont la vitesse de fonctionnement maximum est de 25 impulsions/seconde.

L'assortiment d'accessoires fournis à l'usage du détecteur comprend un château de plomb, un palpeur liquide et un palpeur à main beta/gamma.

EE 18 760 *pour plus amples renseignements*

DIODE D'AMPLIFICATEUR PARAMÉTRIQUE

(Illustration à la page 120)

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

La diode à capacité variable EW97/1, prévue pour l'emploi dans les amplificateurs paramétriques à des fréquences allant jusqu'à la bande S, est maintenant produite par la General Electric Company, qui nous assure que c'est le premier dispositif doté de ces qualités, livrable sur le marché britannique.

La diode EW97/ convient à l'emploi dans les radars et autres systèmes de communication. Conçue pour une dissipation de grande puissance, elle ne nécessite aucune précaution compliquée contre les surtensions.

Elle est montée sur une structure coaxiale pour intercalation directe dans des circuits coaxiaux et de guide d'ondes, et son inductance basse-série n'est que de 0,5mμH. Il existe également un modèle à polarité inversée (EW98/1).

En raison de sa très basse impédance avant et sa très haute impédance arrière, cette diode peut être utilisée comme interrupteur pour hyperfréquences, ce qui constitue un appoint important à l'étude future des récepteurs pour hyperfréquences, entièrement basés sur des dispositifs solides.

Les références des prototypes sont: VX3314 (EW97/1) et VX3333 (EW98/1).

EE 18 761 *pour plus amples renseignements*

RELAIS PROTÉGÉS

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

Une gamme de robustes petits relais c.c. vient d'être réalisée par la Maison Radiospares Ltd. Ces relais sont dotés de contacts en argent massif à guipage deux couches papier, d'une puissance nominale de 250V 2A c.a. (non-inductif). Les connexions sont effectuées sur un panneau à cosses, monté sur un coffret en acier cadmié. Les cosses, ainsi que les connexions des bobinages, sont protégées contre la rouille par un vernis isolant. Ceci garantit la facilité du minutage. Les relais sont pourvus de bobinages enrobés avec des enroulements imprégnés de vernis, ayant une résistance de 1kΩ ou de 12kΩ pour un fonctionnement nominal de 12V ou de 50V. Les relais sont destinés principalement à l'usage industiel ou aux laboratoires généraux.

EE 18 762 *pour plus amples renseignements*

FILTRE NETTOYEUR À ULTRA-SONS

(Illustration à la page 120)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

En raison de l'efficacité des ultra-sons, la solution décapante se charge de crasse en suspension et elle doit être renouvelée fréquemment, selon l'état des pièces que l'on nettoie. En dehors du temps perdu à changer le dissolvant, le renouvellement répété de solutions encore actives peut s'avérer très onéreux.

L'élément à filtre type 1181 a donc été créé par la société Dawe Instruments Ltd. pour la suppression continue des matières solides dans les bains de nettoyage ultrasoniques, construits par eux-mêmes ou par d'autres maisons. Le fluide est tiré du fond du bain de nettoyage par une pompe centrifuge et il est dirigé vers le haut du bain au moyen d'un filtre (centre de la gravure) pourvu d'un robuste élément aggloméré d'acier inoxydable pouvant supprimer toute matière solide, jusqu'à une dimension de moins de 2 microns, pour un débit de liquide de 36,5 litres par heure. Sept éléments de filtre plus approximatifs peuvent être substitués au précédent afin de donner des régimes de débit atteignant 273 litres par heure. Le changement des éléments, soit pour le nettoyage soit pour obtenir une dimension de pore différente, peut être effectué en moins d'une minute, en dévissant le boîtier de filtre cylindrique de sa calotte afin de libérer l'élément de filtre qui peut, lui-même, être ensuite dévissé et retiré. Le boîtier de filtre est tenu par une bride de serrage à ressort sur une plaquette de bois.

Toutes les parties métalliques en contact avec le liquide de nettoyage sont en acier inoxydable. Le tube d'interconnexion est en nylon profilé fort solide. Toutes les pièces peuvent donc être facilement nettoyées et, si nécessaire, stérilisées.

Des filtres de plus grandes dimensions sont fournis sur demande, pouvant assurer le réglage de température du liquide de nettoyage. Quoiqu'étudiés, en principe, pour l'emploi avec les nettoyeurs à ultra-sons, les nouveaux éléments à filtre conviennent tout aussi bien à d'autres procédés de filtrage.

EE 18 763 *pour plus amples renseignements*

ACCÉLÉROMÈTRE

(Illustration à la page 121)

De Havilland Propellers Ltd, Hatfield Aerodrome, Hatfield, Hertfordshire

L'accéléromètre MV300 est le résultat d'un programme de développement visant à la production d'un instrument de grand rendement à bas prix et d'un format réduit. C'est la "troisième génération" d'une série de dessins, dont chacun a permis de gagner une expérience considérable dans le domaine de la production et du fonctionnement. Une des particularités de cet instrument est sa grande sensibilité s'élevant à 25 volts de sortie totale. Il existe un choix de modèles dont les sensibilités vont de 0,5V/g à 8V/g lorsqu'elles fonctionnent avec une charge de 0,5 MΩ, avec des fréquences naturelles de 34 à 137 Hz.

La masse séismique de l'instrument est un induit laminé formant la partie mobile d'un dispositif sélectif c.a. à transformateur différentiel. Cet induit est supporté par un système à ressort à lames, géométriquement semblable au mécanisme à ligne droite de Watt, ce qui garantit le mouvement rectilinéaire pour le centre de gravité et élimine pratiquement les erreurs de couplage croisé et les effets d'accélérations transversales. L'hystérésis globale est inférieure à 0,7% de l'amplitude et les erreurs de couplage croisé sont inférieures à 0,2%. L'instrument est muni, en outre, de pièces d'arrêt pouvant résister à des charges de choc très élevées et à des surcharges soutenues de l'ordre de 100g.

L'instrument étant inductif, les erreurs de résolution sont nominalement de zéro. Un élément démodulateur et filtre, monté intérieurement et noyé dans la résine d'époxyde, fournit une tension de sortie c.c. à ondulation limitée.

L'accéléromètre MV300 a un montage à trois points qui empêche la déformation du coffret si ce dernier est fixé sur des surfaces inégales, et le cadre de l'appareil est fixé au coffret par une autre monture indéformable par surcroît de précaution. Les éléments sont hermétiquement scellés sous carter utilisant exclusivement des scellements soudés et à verre sur métal. Ils sont remplis d'huile pour l'amortissement des impacts et un diaphragme en métal flexible permet la dilation thermique. L'huile de silicium est utilisée comme fluide d'amortissement en raison de son coefficient de viscosité à température relativement basse.

L'appareil est prévu pour l'emploi entre −60° et +100°C, mais il a été éprouvé au delà de ces limites en cours de fabrication, afin de déterminer la stabilité de l'erreur zéro. L'erreur zéro maxima est de 1,5% de l'échelle totale et le coefficient de température maximum est de 2,5 mV/0°C. Normalement, une tolérance de sensibilité de ±5% est suffisante, mais cette marge peut être réduite pour des usages spéciaux.

En cas d'endommagement accidentel de l'appareil, on peut régler extérieurement tant l'erreur zero que la sensibilité. Ceci simplifie les travaux de réparation car le coffret est un ensemble entièrement soudé. Ces réglages s'effectuent en changeant les valeurs de résistances fixes miniatures.

EE 18 764 *pour plus amples renseignements*

BLOCS D'ALIMENTATION SANS BOBINES D'ARRÊT

(Illustration à la page 121)

The Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

Spécifiquement conçus pour les applications de charge d'impulsions continues, la nouvelle série de sous-ensembles d'alimentation sans bobines d'arrêt que vient de présenter le Groupe Electronique Solartron, peut soutenir des impulsions de zéro à pleine charge sans perte de contrôle.

Ceci a pu être réalisé par l'utilisation de circuits de commande entièrement nouveaux qui, tout en gardant un facteur de stabilité supérieur à 400:1, permettent de maintenir l'impédance de source très basse à une fréquence de 0,5 MHz.

La nouvelle série comprend deux sous-ensembles de 100 mA de capacité de charge maxima (AS 951, 952 et 953) et trois sous ensembles de 200 mA (AS 955, 956 et 957), les tensions de sortie s'étendant de 100 à 400 V. La tension de sortie de n'importe lequel des sous-ensembles peut être variée de ±50 V au-dessus de la tension nominale. On peut aussi appliquer des charges réactives sans affecter le rendement spécifié.

De même que les autres blocs d'alimentation Solartron, les caractéristiques nominales de cette série sans bobines d'arrêt ont été scrupuleusement établies et elle comporte l'effectif normal d'éléments de chauffage c.a., tant à prises centrales que sans prises.

EE 18 765 *pour plus amples renseignements*

MACHINE À BOBINER

(Illustration à la page 121)

Westool Ltd, St. Helens Auckland, Co. Durham

Une nouvelle machine à bobiner d'établi, complète avec appareil de mesure et organe d'entraînement, dont les dimensions hors-tout son d'environ 611mm × 611mm, est fabriquée maintenant par la société Westool Ltd.

Cette machine est munie d'un tachymètre à réenclanchement à cinq chiffres et elle peut être lancée jusqu'à une vitesse maximum de 4 000 tours/minute. Le taux d'accélération de zéro à la vitesse maximum est déterminé par un volant placé à l'avant de la machine. Le volant actionne un potentiomètre qui commande l'embrayage électrique Warner, accouplé au moteur.

La machine peut enrouler des bobines ayant jusqu'à 127mm de diamètre et 76mm de longueur, dont les fils vont de 0,315 à 0,025mm de diamètre (jauges anglaises 30 à 50). L'arbre du porte-bobine convient à des bobines ayant jusqu'à 152,5mm de diamètre, mais pour l'enroulement des fils de calibre réduit, soit de 0,051 à 0,025mm (jauges anglaises 47 à 50), un petit dispositif bobineur spécial est également fourni.

Le mécanisme d'amenage comporte deux freins électriques Warner, accouplés à deux engrenages hélicoïdaux constamment engrenés à deux vis de plomb. L'amenage est réglé au moyen de deux butées ajustables qui commandent des micro-contacts excitant ou désexcitant les freins, assurant une inversion de direction instantanée et positive.

L'espacement du fil (tours par pouce) est réglé par un volant se déplaçant sur une échelle étalonnée, graduée en jauges anglaises ou en décimales d'un pouce.

EE 18 766 *pour plus amples renseignements*

Résumés des Principaux Articles

Un aperçu de la télévision en couleur en Grande-Bretagne par R. D. A. Maurice

Résumé de l'article aux pages 68 à 73

Depuis 1953, la B.B.C. s'intéresse activement à la télévision en couleur et elle a accompli depuis lors un travail considérable en vue de déterminer l'adaptabilité des normes américaines du N.T.S.C. (National Television System Committee) au système britannique à 465 lignes. Un nombre de rapports et une monographie des services techniques de la B.B.C. ont été publiés sur ce travail, dont le présent article fournit un compte-rendu concis s'adressant particulièrement à ceux qui sont déjà versés dans la technologie de la télévision en couleur.

L'utilisation des calculateurs analogiques par F. C. Harbert

Résumé de l'article aux pages 74 à 77

La première partie de cet article en trois parties sur l'utilisation des calculateurs analogiques décrit le dispositif de calcul que nécessite la résolution d'équations algébriques simultanées, de problèmes de valeurs propres et d'équations polynômes, par l'emploi d'une méthode itérative.

La seconde partie décrira une méthode pour obtenir des solutions approximatives d'équations partielles différentielles, par la technique dite des différences finies.

La dernière partie, enfin, présentera les divers procédés de calcul nécessaires à la simulation du retard de transport, par l'emploi des approximations de Padé. Un exemple démontrant l'effet d'un pareil retard sur la stabilité d'un système à circuit fermé, et un moyen de parer à son effet déséquilibrant, complèteront cet article.

Les caractéristiques et les applications des Diodes Zéner (référence de tension) par J. A. Chander

Résumé de l'article aux pages 78 à 86

Les caractéristiques détaillées des diodes Zéner sont présentées dans cet article, qui donne également des courbes montrant les variations de résistance de pente c.a. et c.c., tant avec le courant qu'avec la tension, les variations de tension de rupture selon la température ainsi que les caractéristiques avant des diodes.

Le circuit général de l'alimentation stabilisée simple est analysé, de même que les restrictions dues aux limites de diodes et de circuits. La compensation de température des diodes est étudiée et quelques résultats sont mentionnés. La stabilité de la tension de sortie, selon les variations de charge et d'entrée, est indiquée en détail, en utilisant les caractéristiques susmentionnées. Après avoir examiné un stabilisateur à deux étages, l'auteur conclut par un exemple pratique d'une référence à haute stabilité utilisant des circuits relativement simples.

Un suppresseur de quadrature à transistors-thermistors par I. C. Hutcheon et D. N. Harrison

Résumé de l'article aux pages 87 à 91

Les effets de quadrature dans les servosystèmes c.a. sont supprimés si le signal résiduel amplifié est démodulé et utilisé pour actionner un signal à réaction négative, en phase de quadrature précise avec le courant alternatif principal. Les auteurs décrivent un suppresseur de ce genre, utilisant la pré-amplification à transistors et deux thermistors chauffés indirectement pour commander la réaction.

Le tube compte-décades EIT dans les circuits à comptage rapide par V. Radeka

Résumé de l'article aux pages 92 à 95

Les possibilités de comptage rapide au moyen des tubes compte-décades du type EIT sont examinées par l'auteur. Il étudie, à cet égard, le déplacement du faisceau électronique et calcule le temps voulu pour atteindre le point stable. Après avoir déterminé théoriquement e expérimentalement les limites de la précision de comptage, au regard de la sécurité de fonctionnement requise, il relève le fait qu'on peut employer le EIT jusqu'à environ 10^6 impulsions par seconde.

L'auteur présente ensuite un schéma de circuit d'un compteur de décades de sécurité à comptage rapide convenant à l'usage général. Le fonctionnement du circuit est entièrement indépendant de la forme d'onde d'entrée et des variations de tension. Une seule alimentation en c.c est nécessaire.

L'utilisation économique du stabilisateur série par D. J. Collins et J. R. Pearce

Résumé de l'article aux pages 96 à 97

Cet article souligne le fait que le stabilisateur série est peu rentable lorsqu'il est utilisé dans des conditions défavorables, vu qu'il exige alors des éléments série. Si le courant de charge demeure essentiellement constant, l'élément série peut être shunté par une résistance. Le circuit de base est décrit par les auteurs, ainsi qu'une solution pratique du problème par l'emploi d'une section intermédiaire à résistance de shunt.

Quelques paramètres de rendement des redresseurs de courant à jonction au silicium par D. R. Coleman

Résumé de l'article aux pages 98 à 102

Le rendement électrique des redresseurs de courant force au silicium, en fonction de la sortie de puissance d'une connexion de conversion, est conditionné par leurs caractéristiques et dépend de la température de jonction. Le contrôle de la température est examiné dans cette étude du point de vue de la résistance thermique, une méthode étant indiquée pour calculer la dissipation de puissance et les régimes de connexion des redresseurs au silicium. Mention est faite de certaines conditions nécessaires à un rendement transistoire satisfaisant.

Un tachymètre de haute précision à court terme par C. Reed

Résumé de l'article aux pages 103 à 105

L'auteur décrit un instrument pour mesurer la vitesse d'un alternateur, celle-là étant indiquée dans une gamme d'environ $\pm 10\%$ de vitesse nominale. La méthode employée résulte en une échelle linéaire et elle peut être adaptée pour d'autres vitesses et gammes nominales. La précision est maintenue par un oscillateur incorporé piloté par quartz et l'étalonnage est facilement contrôlé et réglé. Certains perfectionnements sont suggérés en conclusion.

Un amplificateur électrométrique à plusieurs sensibilités et à réaction variable par J. H. Leck et W. E. Austin

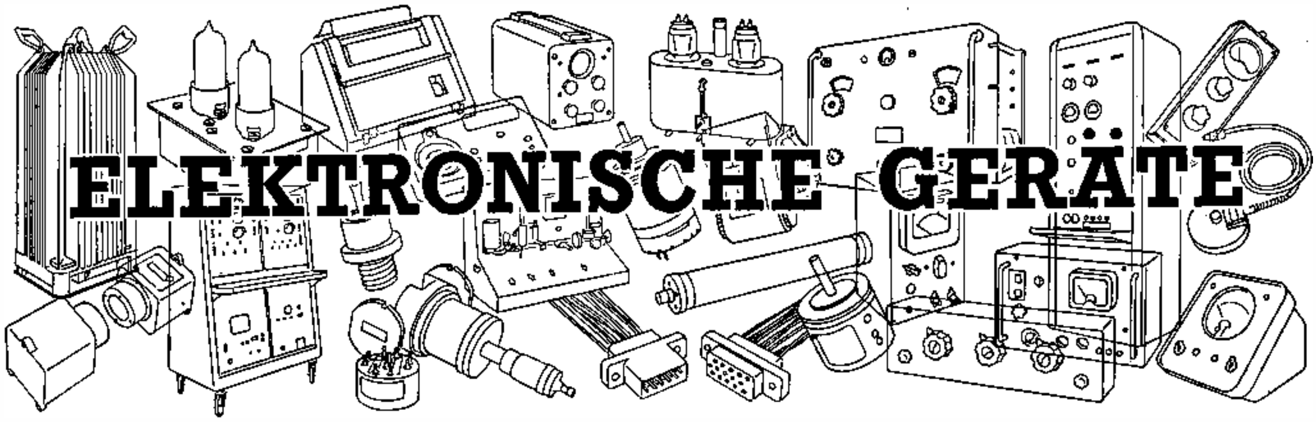
Résumé de l'article aux pages 106 à 107

L'amplificateur électrométrique simple décrit dans cet article a été trouvé à la fois précis et sûr pendant une période de douze mois an cours desquels il a été utilisé pour la mesure de courants ioniques positifs allant jusqu'à 10^{-15}A. Par l'emploi de lampes électrométriques miniature modernes conjointement avec un amplificateur c.c. à transistor de grand gain et à grande réaction négative générale, on allie l'avantage de la simplicité à une grande précision et une sensibilité adéquate. Les désavantages inhérents aux amplificateurs c.c. à transistor et leur température instable ont été surmontés dans cette application par l'utilisation d'un transistor au silicium au premier stade et par le fonctionnement avec entrée d'une source de courant constante.

Un oscillateur résistance-capacité accordé par W. D. Ryan et F. E. Hetherington

Résumé de l'article aux pages 108 à 110

La capacité variable, présentée par les redresseurs à rondelles au sélénium avec polarisation inverse, est utilisée dans un circuit à oscillateur résistance-capacité, du type à pont, afin de produire une basse fréquence variable dépendant du signal c.c. appliqué. Les caractéristiques des redresseurs appropriés, ainsi que le schéma du circuit de l'oscillateur et ses particularités de fonctionnement sont analysés dans cet article.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 118 bis 121

FREQUENZMESSER

(Abbildung Seite 118)

Airmec Ltd, High Wycombe, Buckinghamshire

Der Frequenzmesser Type 265 kann die Frequenzen sich wiederholender Wellenformen von 10 Hz ... 100 kHz mit ±1% Genauigkeit messen. Die Eingangswellenform wird in eine Rechteckform umgewandelt, begrenzt, und dann in eine Integrationsschaltung gespeist, deren Ausgangsspannung auf einem grossen, klar lesbaren Tafelmessgerät angezeigt wird. Die Schaltung macht die Instrumentanzeige von der Amplitude oder Wellenform der Eingangsspannung unabhängig.

Das Gerät kann z.B. als elektronisches Tachometer benutzt werden. Die Geschwindigkeit einer sich drehenden Welle kann direkt gemessen werden, wenn die Welle bei jeder Umdrehung einen Kontakt schliesst oder öffnet und die so unterbrochene Spannung dem Frequenzmesser zugeführt wird. Andererseits kann der optische Airmec Messkopf 283 mit dem Frequenzmesser Verwendung finden, wobei die Bewegung durch Ablenkung oder Unterbrechung eines Lichtstrahls gemessen wird, ohne dass das zu messende Gerät in irgendwelcher Weise belastet wird.

Der Frequenzmesser ist volltransistorisiert, leicht und tragbar. Er kann vom Wechselstromnetz oder eingebauten Batterien gespeist werden. Als Stromquelle und Signalverbindung für den optischen Messkopf ist ein Sockel auf der Geräterückseite vorgesehen. Zur Eigeneichung ist ein stabiler 5 kHz Oszillator eingebaut, und das Gerät ist direkt in Frequenz und Umdrehungen pro Minute geeicht. Auf der Fronttafel ist ein Anschluss für ein Aussenmessgerät oder einen Schreiber vorgesehen.

EE 18 751 *für weitere Einzelheiten*

TRIMMER-HÜLSENKONDENSATOR

(Abbildung Seite 118)

Erie Resistor Ltd, 1 Heddon Street, London, W.1

Erie hat die Type CT20 entwickelt, um Forderungen der Industrie nach einem preisgünstigen Trimmerkondensator mit stabilem, verlustfreiem, keramischem Dielektrikum einfacher, jedoch widerstandsfähiger Konstruktion und geringen Gewichtes gerecht zu werden, der sich zum Verdrahten zwischen Lötpunkten, zwischen Schaltbrettlötfahnen, an Drehschaltern usw. eignet. Type CT20 entspricht all diesen Anforderungen und hat die verschiedensten Anwendungsmöglichkeiten. Der Kapazitätsbereich liegt zwischen 1 ... 6 pF, und der Abgleich erfolgt durch Veränderung der Hülsenlage mittels eines isolierten Werkzeugs. Die Hülse ist für diesen Zweck mit Nasen versehen.

Bemerkenswert ist, dass keine der Elektroden an Masse zu liegen braucht, da keine zwangsläufig geerdete Befestigung vorgesehen ist; dass durch die Nasenkonstruktion die Möglichkeit mechanischer Kupplung besteht, und dass durch geringes Gewicht und Drahtanschlüsse Verwendung für gedruckte Schaltungen ermöglicht wird.

EE 18 752 *für weitere Einzelheiten*

STABILISIERTES NIEDERSPANNUNGS-NETZGERÄT

(Abbildung Seite 118)

Lexor Electronics Ltd, 25 Allesley Old Road, Coventry

Lexor Electronics kündigt zwei neue, stabilisierte Miniatur-Netzgeräte mit Transistorbestückung an. Typen LT.12-05 und LT.24-05 haben Spannungsbereiche von 5,5 ... 14 V bzw. 18 ... 28 V und liefern 0 ... 500 mA nicht geerdeten Gleichstrom. Die Ausgangsspannungen sind über die genannten Bereiche stufenlos durch Bedienungsknopf oder Voreinstellung regelbar. Die Regelung ist innerhalb ±0,15% für ±10% Netzschwankungen und/oder Null bis Vollbelastung. Geräusche und Restwelligkeit sind unter 2,5 mV Spitze-zu-Spitze, und die Geräte sind für Dauerbetrieb in Umgebungstemperaturen bis zu 40°C bemessen.

Die Geräte sind völlig transistorisiert und in Miniatur-Messgerätgehäuse eingebaut. Der Netzeingang kann zwischen 200 ... 250 V 50 Hz eingestellt werden. Sicherungen für die Netz- und Ausgangsseite, sowie Buchsen für ein Monitor-Voltmeter sind vorgesehen.

EE 18 753 *für weitere Einzelheiten*

IMPULSVERSTÄRKER

(Abbildung Seite 118)

Ekco Electronics Ltd, Southend-on-Sea, Essex

Ekco Electronics hat einen neuen, hochstabilen Universal-Impuls-Breitbandverstärker Type N640 angekündigt. Er wurde besonders für Szintillationszähler ohne eingebaute Verstärker entwickelt und kann Ratemeter und Untersetzer treiben, die 5 V oder mehr Eingangsspannung benötigen.

Ein sechsstufiger Verstärkungswähler ist eingebaut. Der impedanzarme Ausgang wird positiv, wenn der Eingang negativ wird. Durch eine einfache Änderung kann der Eingang auch negativ gemacht werden. Die Stromversorgung erfolgt durch das zugehörige Ratemeter oder den Untersetzer. Für den Bereich von 50 Hz . . . 200 kHz ist die Verstärkung annäherungsweise konstant, und die obere Frequenz für Halbleistung ist 1 MHz.

EE 18 754 *für weitere Einzelheiten*

LANGSPIEL-TONBANDGERÄT

(Abbildung Seite 119)

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

Das Paraphone kann für unbeschränkte Zeit ohne Bedienung laufen, wobei es insgesamt 40 Stunden aufnehmen oder wiedergeben kann, bevor es sich wiederholt. Eine Bandgeschwindigkeit von 9,5 cm/s wurde gewählt, um einen Frequenzgang von 60 Hz...6kHz in den 3 dB Grenzen zu erreichen. 140 mmø Spulen mit 365 m Band auf dünnem Träger geben 1 Stunde Aufnahme pro Spur. Ein zweispuriger Kopf in einem Nockenmechanismus kann 20 Kopfstellungen auf einem 25,4 mm breiten Band einnehmen. Im Betrieb werden 20 Spuren mit dem einen Kopf und die anderen 20 Spuren mit dem zweiten Kopf aufgenommen.

Zwei Bandtransport-Vorrichtungen mit gegenläufigen Schwungrädern werden von nur einem Motor angetrieben und führen das Band in der für die gewählte Spur zugeordneten Richtung: von links nach rechts über den Kopf für geradzahlige Spuren und von rechts nach links für ungeradzahlige Spuren. Eine fotoelektrische Zelle betätigt Magnetspulen für die Richtungsumkehrung und verschiebt den Kopf zur nächsten Spur.

Ein besonderes Merkmal des Gerätes ist die Selbstanlaufposition, in der die Maschine das Band schnell bis zum richtigen Ende für die gewählte Spur aufspult, dann die Richtung umkehrt und die Spur wiedergibt.

Anlauf, schneller Vor- und Rücklauf, Stop und Selbstanlaufposition werden durch Drucktasten eingestellt. Bei Bedienung der Anlauftaste läuft das Gerät normalerweise von derselben Stelle an, an der es gestoppt wurde.

Nur bei Einstellung auf Wiedergabe gibt ein dreistufiger Schalter am Laufwerk die Positionen „Gerät aus", 1 mW Ausgangsleistung für 600 Ω oder 2 W Ausgangsleistung für jede gewünschte Impedanz. Ein eingebauter Lautsprecher kann im letzteren Falle entweder mit den vollen 2 W gespeist oder mit geringerer Leistung für Überwachung betrieben werden. Bei Ausfall der Stromversorgung kann das Gerät nicht wieder anlaufen, bis der Schalter auf „Gerät aus" und dann auf eine der anderen beiden Stellungen zurückgestellt wird. Für Aufnahme- und Wiedergabegeräte ist eine vierte Schalterstellung für „Aufnahme" vorgesehen.

Alle elektronischen Kreise sind transistorisiert und werden über ein spannungsstabilisiertes Transistor-Netzgerät gespeist. Das Gerät kann für 28 V Gleichstrom (für Flugzeuge usw.) oder mit einem Zusatz-Netzgerät für normale Netzspannungen geliefert werden.

EE 18 755 *für weitere Einzelheiten*

SILIZIUM-SONNEN-EICHZELLE

(Abbildung Seite 119)

International Rectifier Co. (Great Britain) Ltd. Hurst Green, Oxted, Surrey

Die International Rectifier Co. gibt bekannt, dass indirekte Silizium-Sonnen-Eichzellen lieferbar sind, die als genaue Strahlungsnormalmasse zur Wirkungsgradprüfung hergestellter Silizium-Sonnenzellen dienen.

Die Zellen wurden besonders entwickelt, um künstliche Lichtquellen in Sonnenenergiestrahlung zu eichen. Sie können als Sonnenzellenanalog zum pyroheliometrischen Normal betrachtet werden. Industriezweige, die sie benutzen, können nunmehr den Wirkungsgrad gelieferter Sonnenzellen ohne Pyroheliometer oder komplizierte Lichtmessgeräte bestimmen.

Jede gewählte Zelle muss mindestens 9% Umwandlungswirksamkeit haben und ist genau für den tatsächlichen Wirkungsgrad geeicht. Vor der Verpackung wird für jede Zelle eine genaue Eichkurve angefertigt, die die Wirksamkeit für bestimmte Strahlungsintensitäten und Temperaturverhältnisse gibt. Diese Eichkurve ist fest im Instrumentgehäuse befestigt und geht mit der Eichzelle zum Kunden.

Diese Eichzellen sind in ein stabiles, stossfestes Gehäuse eingebaut und durch ein optisch beschichtetes Glasfenster geschützt. Thermoelementzuführungen zur Messung der Zellentemperatur sind vorgesehen.

EE 18 756 *für weitere Einzelheiten*

AM-FM-MESSENDER

(Abbildung Seite 119)

Taylor Electrical Instruments Ltd, Montrose Avenue, Slough, Buckinghamshire

Dieser Messender ist hauptsächlich für Service von AM- und FM-Empfängern und ZF-Stufen von Fernsehempfängern bestimmt. Mit einem Oszillografen zusammen ermöglicht der Messender 61A den Wobbelabgleich von HF-, ZF-, Diskriminator- und Ratiodetektorstufen von AM- und FM-Empfängern. Für den Abgleich von Ratiodetektoren kann die schnelle und genaue Methode angewandt werden, die gleichzeitig AM- und Wobbelsignale benutzt.

Der AM-Generator bestreicht die Frequenzen 4 ... 120 MHz in fünf Bereichen mit Grundfrequenzen. Die Eichgenauigkeit ist ±1% und der Modulationsgrad 35% für 400 Hz. Die FM- und Wobbelteile arbeiten auch mit Grundfrequenzen und vermeiden so die durch die Frequenzunterschiedsmethode entstehende Verwirrung und geringe Eichgenauigkeit. Die FM- und Wobbelbänder bestreichen die Frequenzen 4 ... 12 MHz und 70 ... 120 MHz in drei Bereichen. Die Eichgenauigkeit für den FM-Generator ist ±2%. Während des Wobbelabgleiches kann der AM-Generator als Frequenzmarkengeber benutzt werden. Ein einstellbares Gleichlaufnetzwerk ist eingebaut und Rücklaufauslastung vorgesehen.

Für FM ist der Frequenzhub bis zu 100 kHz von der mittleren Trägerfrequenz mit einer Modulationsgeschwindigkeit von 400 Hz variabel. Der Wobbelhub ist 1 MHz bei Netzfrequenz. Jedes der drei eingebauten Quarze kann durch Schalter für den Quarzeichkreis gewählt werden. Quarze im Frequenzbereich von 1 ... 11 MHz sind nach Wunsch lieferbar.

Alle im Modell 61A gebotenen Einsatzmöglichkeiten können durch eingebaute Schalter eingestellt werden und benötigen keine Verbindungen ausserhalb des Gerätes. Durch das Schalten können Dauerton, AM, FM, Wobbel mit Dauerton oder Wobbel mit AM gewählt werden. Bei Benutzung des Wobbelteiles kann das Markenniveau fortlaufend geändert werden.

Ein fünfteiliges Dämpfungsglied mit induktionsfreiem Potentiometer ist zur Dämpfung über den ganzen Frequenzbereich vorgesehen. HF-Verluste werden durch sorgfältiges Abschirmen des Messsender- und des Oszillatorteiles auf ein Minimum herabgesetzt.

Überwachung des HF Ausgangs erfolgt mittels des neuen Taylor Hochkant-Instruments. Durch den einstellbaren Pegel des 400 Hz Signals kann auch Service von NF Stufen ausgeführt werden.

EE 18 757 *für weitere Einzelheiten*

ANTISTATIK-FLÜSSIGKEIT

Sionic Ltd, Cromwell House, Wood End Green Road, Hayes, Middlesex

Die Sionic Antistatik-Flüssigkeit wurde zur Verwendung auf allen Gebieten entwickelt, wo Staubansammlungen durch statische Anziehung ein wesentliches Problem darstellen. Nach Angabe der Firma zeigen Versuche, dass sowohl positive als auch negative Ladungen auf der Oberfläche von Kunststoff-Formstücken vorhanden sind. Die höchsten Ladungen sind positiv und entstehen, wenn das Formstück von der Metallform getrennt wird. Die Ladungen niedrigerer Spannung sind entweder positiv oder negativ und können durch Pressen, Heissbearbeitung, Zusammenbau, Einwickeln, Polieren, oder einfach durch Handhabung und Lagern entstehen.

Freischwebender Staub wird durch diese elektrischen Ladungen angezogen und festgehalten. Anscheinend besteht zwischen der Staubansammlung und Grösse, Form oder Fliessdiagramm irgendwelcher Formstücke keinerlei Beziehung. Jedoch scheinen benutzte Gleitmittel und Produktionsmethoden die Staubansammlung zu beeinflussen.

Unter normalen Umständen bleibt ein antistatischer Überzug ein Jahr oder länger wirksam. Der Überzug besteht aus einer monomolekularen Schicht, d.h. sie kann ihrer geringen Dicke wegen nicht mit üblichen Methoden gemessen werden. Sie ist vollständig durchsichtig, und in vielen Fällen fühlen sich die behandelten Gegenstände angenehmer an.

Die Flüssigkeit ist ebenso für Bildschirme von Elektronenstrahlröhren und Instrumentgläser geeignet und wird in 460 cc Aerosol-Packungen oder in Grosspackungen für die Industrie geliefert.

EE 18 758 *für weitere Einzelheiten*

KLEINST-RELAIS

(Abbildung Seite 120)

Vertrieb in England durch Counting Instruments Ltd, 5 Elstree Way, Boreham Wood, Hertfordshire

Das luftdicht abgeschlossene Relais Type RZO wird von Telefon Fabrik Automatic A/S, Kopenhagen, Dänemark hergestellt.

Trotzdem die Abmessungen nur 19 × 19 × 6,35 mm betragen, ist es robust genug, um ausserordentlichen Temperatureinflüssen, schweren Stössen und forcierten Schwingungen zu widerstehen.

Das RZO ist ein Doppelspulenrelais mit zwei Umschaltkontakten, die durch einen ausbalanzierten Anker betätigt werden. Die Kontakte bestehen aus vergoldetem Hartsilber.

Die Spulenkörper werden aus ausgewählten keramischen Materialien hergestellt, um Kontaktverschmutzung durch Gasentwicklung zu vermeiden.

Alle Verbindungen im Relais, mit Ausnahme der in den Spulen liegenden, sind geschweisst und daher hitze-, stoss- und vibrationsbeständig. Die Normalausführung kann zum Einlöten oder Einstecken geliefert werden.

Vor dem Abdichten wird das Relais unter Wärmeeinfluss ausgepumpt und mit trockenem Stickstoff gefüllt.

Die Abbildung zeigt das aus dem Gehäuse entfernte Relais.

EE 18 759 *für weitere Einzelheiten*

VERSEUCHUNGSMONITOR

(Abbildung Seite 120)

Plessey Nucleonics Ltd, Weedon Road, Northampton

Plessey Nucleonics hat einen neuen Universal-Monitor mit einer Reihe von Zusatzgeräten entwickelt, die vielseitige Verwendung ermöglichen. Er kann ausser zur Überwachung von radioaktiven Verseuchungen auch zur Prüfung von festen Körpern und Aktivitätsmessungen in grossen und kleinen Flüssigkeitsmustern eingesetzt werden. Besonders eignet er sich für Hochempfindlichkeitsmessungen, wie sie für die Schätzung der Radioaktivität verdampften Wassers erforderlich sind.

Der Monitor Type PNI 1065 ist durchgehend transistorisiert. Für Normalbetrieb in Fabriken, Krankenhäusern oder Laboratorien ist er netzgespeist, kann aber in Sonderfällen auch von einer 12 V Batterie betrieben werden. Die Hochspannung ist über einen Bereich regelbar, der für die meisten Geiger-Müller- und Szintillationszählzwecke geeignet ist; ein Vorverstärker für Szintillationszähler ist vorhanden.

Die Zählrate wird auf einem Ratemeter mit einem Bereich von 0...1000 Impulse/s angezeigt; für niedrige Raten ist ein elektromechanischer Zähler mit einer Höchstgeschwindigkeit von 25 Impulsen/s vorgesehen. Zum Zubehör gehören Bleikammern, ein Flüssigkeitsmesskopf und ein Beta/Gamma- Handmesskopf.

EE 18 760 *für weitere Einzelheiten*

DIODE FÜR PARAMETRISCHE VERSTÄRKER

(Abbildung Seite 120)

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

GEC stellt jetzt die Diode EW97/1 her, die für parametrische Verstärker bis zu S Bandfrequenzen entwickelt wurde. Die Firma gibt bekannt, dass damit das erste Element mit diesen Eigenschaften in England zur Verfügung steht.

Die Diode ist für Radar- und andere Fernmeldesysteme geeignet. Da sie für hohe Energiezerstreuung konstruiert ist, sind keine sorgfältigen Vorkehrungen gegen Energiestösse erforderlich.

Das Element ist koaxial für direkte Einschaltung in koaxiale oder Hohlleiter aufgebaut und hat nur 0,5 nH Serienimpedanz. Ein Modell mit umgekehrter Polarität ist als Type EW98/1 lieferbar.

Die Diode kann auf Grund der sehr niedrigen Durchlassimpedanz und der sehr hohen Sperrimpedanz auch als Dezimeterwellenschalter benutzt werden. Verwendung als Frequenzvervielfacher wird wesentlich zur Weiterentwicklung von Dezimeterwellenempfängern beitragen, die ausschliesslich mit Festkörperelementen bestückt sind.

Die amtlichen Prototypbezeichnungen sind: VX3314 (EW97/1) und VX3333 (EW98/1).

EE 18 761 *für weitere Einzelheiten*

GEKAPSELTE RELAIS

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

Radiospares Ltd hat vor kurzem eine Serie kleiner, widerstandsfähiger Gleichstromrelais angekündigt. Die Relais sind mit doppelpoligen Umkehrkontakten aus gediegenem Silber ausgerüstet, die für 250 V 2 A Wechselstrom (induktionsfrei) bemessen sind. Die Verbindungen werden zu einem Schaltbrettchen oberhalb des kadmierten Stahlgehäuses herausgeführt. Spulenverbindungen und Lötfahnen sind durch ein Lackflussmittel gegen Korrosion geschützt, das auch das Verzinnen erleichtert. Die Spulenkörper der Relais sind gepresst, und die mit Lack getränkten Wicklungen haben einen Widerstand von 1 kΩ bzw. 12 kΩ für Nennspannungen von 12 V bzw. 50 V. Die Relais sind in erster Linie für Verwendung in Labor und Industrie bestimmt.

EE 18 762 *für weitere Einzelheiten*

FILTER FÜR ÜBERSCHALL-REINIGUNGSBÄDER

(Abbildung Seite 120)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

Durch die Wirksamkeit des Überschalls wird das Reinigungsmittel mit schwebendem Schmutz verseucht und muss, je nach dem Zustand der zu reinigenden Teile, mehr oder weniger häufig ausgewechselt werden. Ganz abgesehen vom Zeitverlust durch den Wechsel, kann die häufige Erneuerung noch aktiver Lösungen sehr teuer werden.

Dawe Instruments Ltd hat daher ein Filtergerät Modell 1181 eingeführt, um laufende Entfernung von Feststoffen aus Überschall-Reinigungsbädern eigener und fremder Konstruktion zu ermöglichen. Flüssigkeit wird vom Boden des Reinigungsbades durch eine Zentrifugalpumpe abgesaugt und über ein Filter (Mitte der Abbildung) der Oberfläche des Bades zugeführt. Das widerstandsfähige Filterelement besteht aus rostfreiem Sinterstahl und kann pro Stunde aus 36,5 Liter alle Festkörper bis unter 2μ herunter entfernen. Statt dieses Filterelementes können sieben gröbere Elemente eingesetzt werden, um Durchflussgeschwindigkeiten von 273 l/h zu erreichen. Auswechseln von Filterelementen zum Reinigen oder zum Einsatz einer anderen Porengrösse kann innerhalb einer Minute durchgeführt werden: das zylindrische Filtergehäuse wird aus seiner Kappe geschraubt, wonach das Filterelement selbst herausgeschraubt werden kann. Das Filtergehäuse ist in einer Federklammer auf einem Holzbrett bequem zugänglich befestigt.

Alle Metallteile, die mit der Reinigungsflüssigkeit in Berührung kommen, sind aus rostfreiem Stahl hergestellt. Die Rohrverbindungen bestehen aus zähem, stranggepresstem Polyamid. Alle Teile können daher leicht gereinigt und wenn nötig sterilisiert werden.

Grössere Filter können auch mit Temperaturregelung für die Reinigungsflüssigkeit geliefert werden. Die neuen Filtergeräte sind zwar hauptsächlich für Überschall - Reinigungsbäder bestimmt, können aber auch für andere Filterzwecke Verwendung finden.

EE 18 763 *für weitere Einzelheiten*

BESCHLEUNIGUNGSMESSER

(Abbildung Seite 121)

De Havilland Propellers Ltd, Hatfield Aerodrome, Hatfield, Hertfordshire

Der Beschleunigungsmesser MV300 ist das Ergebnis eines Entwicklungsprogramms für hochwertige, preisgünstige und kleine Instrumente. Der MV300 ist die dritte Generation einer Entwicklungsreihe. Für jedes der Geräte wurden wesentliche Herstellungs- und Betriebserfahrungen erworben. Bemerkenswert ist die hohe Empfindlichkeit dieses Gerätes mit 25 V Vollausschlag. Eine Auswahl von Modellen mit Empfindlichkeiten von 0,5 V/g ... 8 V/g bei Belastung mit 0,5 MΩ sowie mit Eigenfrequenzen von 34 . . . 137 Hz ist lieferbar.

Die seismografische Masse des Instrumentes besteht aus einem lamellaren Anker, der den beweglichen Teil eines Differenzial - Transformators in einem Wechselstrom-Wandler bildet. Der Anker wird durch ein Blattfedersystem gestützt, das geometrisch Watt's geradliniger Anordnung ähnlich ist und für den Schwerpunkt geradlinige Bewegungen ergibt. Dadurch werden praktisch Kreuzkupplungsfehler und Einflüsse durch Querbeschleunigung ausgeschlossen. Die Gesamthysterese ist unter 0,7% der Amplitude und der Kreuzkupplungsfehler unter 0,2%. Anschläge sind vorgesehen, die sehr hohe Stossbelastungen aushalten und Überlastung bis zu 100 g widerstehen können.

Da es sich um ein induktives Instrument handelt, sind die Auflösungsfehler nominell Null. Ein eingebauter Demodulator mit Filter ist mit Epoxydharz vergossen und liefert eine Ausgangsgleichspannung mit geringer Restwelligkeit.

Der Beschleunigungsmesser hat Dreipunktbefestigung, um bei Montage auf unebenen Flächen ein Verziehen des Gehäuses zu verhüten. Als weitere Vorbeugungsmassnahme ist der Instrumentrahmen im Gehäuse spannungsfrei befestigt. Die Einheiten sind luftdicht in einem Gehäuse verschlossen, das nur Glasdurchführungen und Lötsäume hat. Zwecks Dämpfung sind sie mit Öl gefüllt, und eine vorgesehene Metallmembrane erlaubt Wärmeausdehnung. Für die Dämpfung findet Silikonöl Verwendung, da es einen geringen Temperaturkoeffizienten für die Viskosität hat.

Des Gerät ist für Einsatz bei $-60°C$ ... $+100°C$ gedacht, wird aber während der Herstellung über diese Grenztemperaturen hinaus geprüft, um die Stabilität des Nullfehlers zu bestimmen. Der höchste Nullfehler ist 1,5% des Vollausschlags, und der höchste Temperaturkoeffizient ist 2,5 mV/°C. Im allgemeinen genügt eine Empfindlichkeits-Toleranz von $\pm 5\%$; für Sonderzwecke kann dieser Bereich aber reduziert werden.

Sowohl Nullfehler als auch Empfindlichkeit können von aussen nachgestellt werden, falls das Gerät durch Unfall Schaden erleidet; das vereinfacht die Reparatur, da das Gehäuse vollständig verlötet ist. Die Nachstellung wird durch Auswechseln fester Miniaturwiderstände gegen andere Werte erreicht.

EE 18764 *für weitere Einzelheiten*

DROSSELLOSE NETZGERÄTE

(Abbildung Seite 121)

The Solartron Electronic Group Ltd. 45 Thames Street, Kingston, Surrey

Die Solartron Electronic Group kündigt eine Serie drosselloser Netzgerät-Bausteine an, die besonders für Dauerbetrieb bei Impulsbelastung entwickelt wurden und Belastungsstösse von null bis zur vollen Belastung ohne Regelungsausfall vertragen können.

Diese Eigenschaft wird durch vollständig neue Regelungsschaltungen erreicht, die es ermöglichen, die Quellenimpedanz bis zu 0,5 MHz trotz eines Stabilisierungsfaktors, der besser als 400 : 1 ist, sehr niedrig zu halten.

Die Serie umfasst drei Bausteine für 100 mA Höchstbelastung (AS 951, 952, 953) und drei für 200 mA (AS 955, 956, 957) mit Ausgangsspannungen von 100... 400 V. Die Ausgangsspannung kann über einen Bereich von ± 50 V der Netzspannung eingestellt werden. Blindbelastungen können angeschlossen werden, ohne die Leistung zu beeinflussen.

Genau so wie die anderen Solartron-Netzgeräte ist diese drossellose Serie reichlich bemessen und gibt auch die übliche Wechselstrom-Heizversorgung mit oder ohne Mittelanzapfung ab.

EE 18765 *für weitere Einzelheiten*

SPULENWICKELMASCHINE

(Abbildung Seite 121)

Westool Ltd, St. Helens Auckland, Co. Durham

Westool Ltd stellt eine neue Tischwickelmaschine mit Zähler und Antrieb her, die nur ungefähr 610 mm lang und 610 mm breit ist.

Die Maschine ist mit einem fünfstelligen Tourenzähler mit Nullstellung ausgerüstet und kann mit einer Höchstgeschwindigkeit von 4000 U/m laufen. Die Anlaufgeschwindigkeit von Null bis zur Höchstgeschwindigkeit wird durch ein Bedienungshandrad an der Maschinenvorderseite reguliert, das das Potentiometer einer elektrischen Warner-Kupplung zum Motor einstellt.

Spulen bis zu 127 mm Durchmesser und 76 mm Länge können mit Drahtstärken von 0,025 ... 0,31 mm gewickelt werden. Der Abroller ist für Haspeln bis zu 152 mm Durchmesser geeignet. Für Drahtstärken von 0,025 ... 0,05 mm wird jedoch auch ein Feindrahtabroller geliefert.

Zwei elektrische Warner-Bremsen sind in die Transportvorrichtung eingebaut und mit zwei Schneckenrädern gekuppelt, die dauernd mit zwei Leitspindeln in Eingriff stehen. Der Transport wird mittels verstellbarer Anschläge eingestellt, die durch Mikroschalter die elektrischen Bremsen erregen oder abregen, wodurch eine genaue und sofortige Richtungsumkehr erzielt wird.

Die Steigerung wird mittels eines Handrades eingestellt, dessen Skala in Drahtlehren oder Zolldezimalwerten geeicht ist.

EE 18766 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Farbfernsehen in England von R. D. A. Maurice

Zusammenfassung des Beitrages auf Seite 68-73

Die British Broadcasting Corporation beschäftigt sich seit 1953 mit dem Farbfernsehen. In den vergangenen Jahren wurde viel an der Frage gearbeitet, ob sich das amerikanische N.T.S.C.-System dem britischen 465-Zeilen-System anpassen lässt. Einige Berichte und eine Sonderveröffentlichung der BBC Engineering Division wurden in diesem Zusammenhang veröffentlicht. Die vorliegende Arbeit gibt eine kurzgefasste Zusammenstellung für diejenigen, die mit der Farbfernsehtechnik schon vertraut sind.

Analogrechner-Technik von F. C. Harbert

Zusammenfassung des Beitrages auf Seite 74-77

Der erste Teil dieses dreiteiligen Beitrags über Analogrechner-Technik beschreibt Rechneranordnungen zur Lösung von gleichzeitigen algebraischen Gleichungen, Eigenwertproblemen und vielgliedrigen Gleichungen mit Hilfe einer Kettenmethode.

Der zweite Teil beschreibt eine Methode für Näherungslösungen partieller Differentialgleichungen mit begrenzten Differentialmethoden.

Im letzten Teil wird eine andere Rechenanordnung beschrieben, die für die Simulierung einer Transportverzögerung erforderlich ist. Padé's Annäherungen werden benutzt und durch ein Beispiel erläutert, das den Einfluss einer solchen Verzögerung auf die Stabilität eines Schleifensystems zeigt, sowie eine Technik, die diesem entstabilisierenden Einfluss entgegenwirkt.

Kennlinien und Einsatz von Zener-(Referenz-) dioden von J. A. Chandler

Zusammenfassung des Beitrages auf Seite 78-86

Die Kennlinien der Zenerdioden werden ausführlich besprochen und Kurven werden gezeigt, die die Schwankungen des Wechsel- und Gleichstromwiderstandes der Steilheit für Spannung und Strom, die Temperaturschwankungen der Zenerspannung und die Durchlasskennlinien der Dioden darstellen. Eine allgemeine Schaltung für eine einfache konstante Stromquelle wird unter Berücksichigung der durch die Dioden und die Schaltung auftretenden Beschränkungen besprochen. Temperaturausgleich für die Dioden wird erwogen und die Resultate gezeigt. Unter Benutzung der erwähnten Kennlinien wird die Stabilität der Ausgangsspannung in Bezug auf Schwankungen der Eingangsspannung und Belastung untersucht. Ein zweistufiger Stabilisator und ein praktisches Beispiel einer hochkonstanten Referenzspannung mit verhältnismässig einfachen Schaltungen werden gezeigt.

Eine Quadratursperre mit Transistor-Thermistor Gegenkopplung von I. C. Hutcheon und D. N. Harrison

Zusammenfassung des Beitrages auf Seite 87-91

In einem Wechselstrom-Servosystem können Quadratureffekte ausgeschaltet werden, wenn das verstärkte Restsignal entmoduliert und dann dazu verwandt wird, eine negative Gegenkopplung zu regeln, die gegenüber dem Hauptwechselstromträger um genau 90° phasenverschoben ist. Die Konstruktion einer Sperre mit Transistor-Vorverstärkung wird beschrieben, die zwei indirekt geheizte Heissleiter zur Regelung der Gegenkopplung benutzt.

Dekadenzählröhre EIT in Schnellzählkreisen von V. Radeka

Zusammenfassung des Beitrages auf Seite 92-95

Die Möglichkeiten, mit EIT Zählrörhen schnell zu zählen, werden untersucht. Die Ablenkung de' Elektronenstrahls wird betrachtet und die zur Erreichung des stabilen Punktes erforderliche Zeit kalkuliert. Die Grenzwerte für genaues Rechnen werden im Zusammenhang mit der geforderten Betriebssicherheit theoretisch und versuchsmässig bestimmt. Der Befund zeigt, dass die EIT bis zu ungefähr 10^6 Impulse pro Sekunde Verwendung finden kann. Eine für Allgemeinverwendung geeignete Schaltung für dekadische Schnellzähler wird angegeben. Das Arbeiten der Schaltung wird in keiner Form von Eingangswellenform oder Netzschwankungen beeinflusst. Nur eine Gleichstromspannung ist erforderlich.

Ersparnisse in Serienstabilisatoren von D. J. Collins und J. R. Pearce

Zusammenfassung des Beitrages auf Seite 96-97

In diesem Beitrag wird darauf hingewiesen, dass der Wunsch, die schlechten Bedingungen des Serienstabilisators zu überkommen, unwirtschaftliche Ansprüche an Serienelemente stellt. Wenn der Belastungsstrom im wesentlichen konstant ist, kann ein Widerstand parallel zum Serienelement geschaltet werden. Zur praktischen Lösung dieses Problems wird eine Grundschaltung beschrieben, in der ein Widerstand einem Pufferteil parallel geschaltet ist.

Leistungsparameter von Silizium-Leistungsflächengleichrichtern von D. R. Coleman

Zusammenfassung des Beitrages auf Seite 98-102

Als Ausgangsleistung eines Gleichrichteranschlusses betrachtet, bezieht sich die elektrische Leistung der Silizium-Leistungsgleichrichter auf deren Kennlinien und hängt, wie gezeigt wird, von der Kristalltemperatur ab. Temperaturregelung wird vom Gesichtspunkt des Wärmewiderstandes aus besprochen. Eine Methode zur Berechnung der Verlustleistung und der Anschlussbemessungen für Silizium-Gleichrichter wird beschrieben. Einige für einwandfreies Arbeiten bei Einschwingvorgängen notwendige Uberlegungen werden erwähnt.

Ein Tachometer mit hoher, kurzfristiger Genauigkeit von C. Reed

Zusammenfassung des Beitrages auf Seite 103-105

Ein Gerät für Geschwindigkeitsmessungen an einem Wechselstromerzeuger, das eine Nenngeschwindigkeit mit einem Bereich von $\pm 10\%$ anzeigt, wird beschrieben. Die angewandte Methode ergibt eine geradlinige Skala und kann anderen Nenngeschwindigkeiten und Bereichen angepasst werden. Die Genauigkeit wird durch einen eingebauten quarzgesteuerten Oszillator eingehalten, und die Eichung kann leicht geprüft und nachgeregelt werden. Verfeinerungen werden vorgeschlagen.

Mehrbereich-Elektrometerverstärker mit regelbarer Gegenkopplung von J. H. Leck und W. E. Austin

Zusammenfassung des Beitrages auf Seite 106-107

Der in dem Beitrag beschriebene einfache Elektrometerverstärker hat sich während einer zwölfmonatlichen Periode zur Messung positiver Ionenströme bis zu 10^{-15} A herunter als genau und zuverlässig bewährt. Die Verwendung einer modernen Miniatur-Elektrometerröhre zusammen mit einem transistorisierten Gleichstrom-Leistungsverstärker mit grosser negativer Gesamtgegenkopplung ergibt eine vorteilhafte Kombination von Einfachheit, hoher Genauigkeit und ausreichender Empfindlichkeit. Der dem transistorisierten Gleichstromverstärker innewohnende Nachteil der Temperaturunbeständigkeit wird durch Einsatz eines Silizium-Transistors in der ersten Stufe und Betrieb mit Eingaben von einer konstanten Stromquelle überkommen.

RC-Oszillator mit Spannungsabstimmung von W. D. Ryan und F. E. Hetherington

Zusammenfassung des Beitrages auf Seite 108-110

Die im Selen-Trockenscheibengleichrichter mit Vorspannung in Sperrichtung auftretende variable Kapazität findet in einem RC-Oszillator des Brückentyps Verwendung, um eine mit der angelegten Gleichspannung variable Niederfrequenz hervorzurufen. Die Kennlinien geeigneter Gleichrichter, die Oszillatorschaltung und ihre Betriebsdaten werden beschrieben.