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Commentary

IN March 1956 the Television Advisory Committee was asked by the Postmaster General of the day to advise on fundamental technical problems of television development in this country. More specifically, the Committee was instructed to look into the existing 405 line system to see if it were likely to remain adequate for the next 25 years and also to make recommendations for a compatible colour television system.

As is well known, the world's first high definition television system set up by the BBC in 1937 was based on the 405 line system, and it has served the country well for over 20 years. The choice was a wise one at the time but technical progress, together with different line standards in Europe, larger screens, and the demand for alternative programmes, have conspired to bring about its doom and before long the present 405 line system will be no more than a museum piece.

It was therefore a foregone conclusion that the Television Advisory Committee in its Report* which has just been published, would find that the potentialities of the 405 line system have now been exploited to the full and the system is inadequate for the next 25 years.

The choice of an alternative system, however, has involved the Television Advisory Committee in a considerable amount of labour over the past four years and has necessitated the setting up of a Technical Sub-Committee to investigate the many technical problems arising from the changeover to a new system.

There are a number of political and economic factors which are largely outside the terms of reference of the Committee, but which nonetheless have had to be taken into consideration. These are largely the subject for debate by the Government who must reach their decision before the present Television Act expires in 1964, but it is essential that the technical matters be thoroughly investigated before the Government embarks on a long-term television development policy.

To this end, a most comprehensive series of tests have been carried out by the BBC, ITA, Department of Scientific Industrial Research, the Post Office and members of the Radio and Electronic Industry.

The technical problem to be investigated can be stated very briefly. At present the BBC is able to cover 98 per cent of the population of the country using the five available channels in Band I (41 to 68Mc/s) while the ITA with four of the eight channels in Band III (173 to 216Mc/s) covers 94 per cent of the population. But the BBC and the ITA each radiate a single national programme on the present 405 line system—there are no extra channels available in Band I, but if Band III were exploited to the full the most that the Bands together could provide would be three programmes, two with a coverage of 98 per cent and one with 95 per cent coverage of the population. In other words, only one additional programme to the present system of a single BBC programme and a single ITA programme.

Clearly, the position is unsatisfactory and the use of other suitable bands has had to be explored since Band III is now extensively committed in the United Kingdom and elsewhere for essential communication services which cannot be displaced.

Use therefore must be made of Band IV (470 to 582Mc/s) and Band V (606 to 960Mc/s) and it is this aspect which has taken up so much of the Technical Committee's time. The main burden of the work has fallen on the BBC who set up a high power transmitter in Band V and along with the other interested parties carried out extensive propagation trials.

The results of these propagation tests and the studies made on the relative merits of the 405 and 625 line systems make it clear that the future television development programme for this country lies in the use of Bands IV and V with a 625 line system with an 8Mc/s channel in line with the existing European standard system.

The report states categorically, and few will disagree, that the use of Bands IV and V will be the last opportunity for us to change our line standard. If the 405 line standard is retained then we are committed to it indefinitely with all the complexities that are involved where programmes are exchanged.

What is much less clear is the manner and timing with which these changes will be brought into being. The 405 line system at present in use in Bands I and III must be retained for some years in order not to make the several million television receivers obsolescent since they cannot be converted for 625 line operation in Bands IV and V. But ultimately there must be a single system for all the Bands and if we are not to be the only one out of step with the European countries it must be a 625 line system.

On the question of costs, the Television Advisory Committee is understandably vague. It puts the cost of transmitting stations providing near national coverage for one programme at £15M and £25M for three programmes. The cost of new receivers to the general public they regard as of the greatest importance, since it already represents an annual expenditure of £150M. If single standards were employed on all bands, the increased cost over and above the present £150M would be £14M for a 405 line system and £18M for 625 line system.

The Committee has little to say on the progress of colour television beyond the fact that the proposed line standards will allow a fully compatible colour system when the time is ripe.

* *Report of the Television Advisory Committee 1960 H.M.S.O. Price 1s.*

A Display Unit for a Flight Simulator

By A. G. Barnes*, B.Sc. and W. F. Coulshed*, B.Eng.

English Electric Aviation Ltd use a flight simulator to assess aircraft characteristics in the design and development stages. A visual representation of the ground is presented to the pilot on a c.r.t. in the simulator cockpit. This display, based on the 'contact analogue' type of presentation has been developed at Warton with the particular requirements of the flight simulator in mind. The design limitations and electronic techniques used in generating the display are described in this article, together with a short account of the display in use.

(Voir page 460 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 465)

DURING the past decade, considerable developments have been made in the field of flight simulation. This has been brought about partly by the new computing techniques now available, and partly by the stimulus of demand. Modern high-speed aircraft are both complex and expensive to operate, and the use of flight simulators for crew training is now accepted practice. By such means, in fact, pilots can become familiar with an aircraft's handling characteristics, emergency drills, and let-down procedures, in a safe, convenient, and economical fashion.

A further extension in the use of flight simulation is that of basic research, whereby the stability and control of an aircraft at an early design stage may be determined with a pilot in the servo loop. Considerable experience has been accrued in the past five years by the design staff of English Electric Aviation Ltd of the requirements for such a simulator, as opposed to one used simply for training. For example, in a training simulator, emergencies such as engine failure must be included, but are unnecessary for research purposes. Moreover, the basic aircraft characteristics of the trainer are fixed to represent one particular aircraft, but must be readily interchangeable in the research simulator.

A further, and vital, difference is that of display. Most training simulators have an accurate representation of cockpit layout, and the pilot flies by reference to the instruments of the standard blind flying panel; this is inadequate for research purposes. In appraising the handling characteristics of an aircraft, the lateral and longitudinal short period oscillations are of supreme importance, and are best seen by reference to the horizon. To see the lateral oscillation on the standard instrument panel is difficult, because cross reference to several instruments, and subsequent interpretation, is necessary.

Initially, the research flight simulator had a display which consisted of a horizon line presentation with a yaw marker, produced on a 10in c.r.t., by which the pilot could see bank, pitch and yaw displacements. The basic limitations of such a display were soon apparent. Even the inclusion of one or two flight instruments did not persuade the pilots that the gap between simulation and flight was narrow, although good correlation between computed and 'recorded in flight' aircraft response was obtained.

As a consequence, the display to be described was developed to present the pilot with an easily distinguishable reference, bearing in mind the purpose and limitations of the simulator, and closely relating these to the cost and complication.

Form of the Presentation

Any display which simply consists of horizon and target

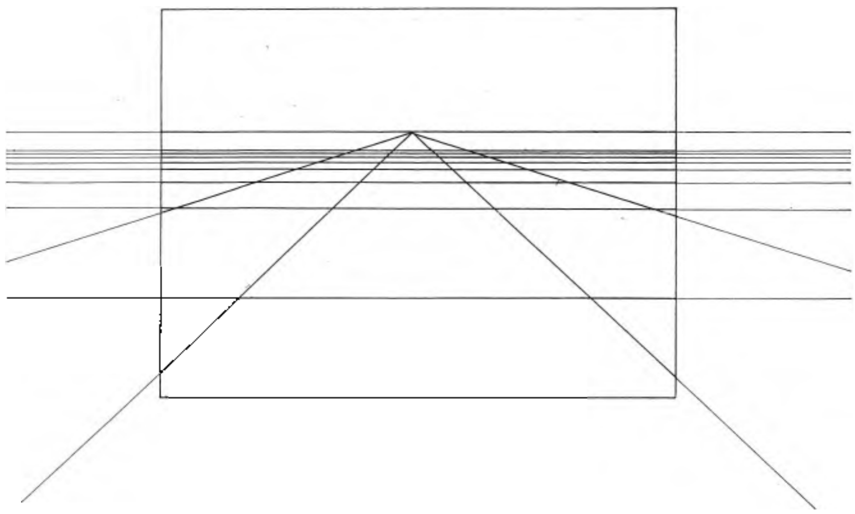


Fig. 1. 'Runway' perspective display

at infinity only shows angular information to the pilot. It is apparent that, when flying near the ground, this is quite inadequate, and in fact two further impressions must be given: one of movement, the other of perspective. Cdr. Hoover[1] (U.S.N.) links with these the texture of the presentation; this may be considered as the quality of the presentation, rather than an absolute quantity like perspective (which must be simulated correctly or not at all). In fact, the texture is simply the means by which the perspective and movement are seen.

Figs. 1 and 2 show two similar presentations which give an impression of perspective. It is readily apparent that if the horizontal lines of Fig. 1 are made to move downwards from the horizon at ever increasing speed, an impression of movement will be obtained. If the lines from each vanishing point of Fig. 2 are made to rotate in a non-linear fashion, a similar impression of movement is obtained. Moreover, because the lines give an impression of perspective, by varying the slope of the lines and the position of the vanishing points, both directional and positional orientation can be obtained. Hence not only can the pilot get roll, pitch and yaw information from this display, he can also deduce relative height, speed, and sideslip.

* *English Electric Aviation Ltd.*

Specification of the Display

The following requirements for the system represent a compromise between complete angular freedom in roll, pitch, and yaw, (which implies (a) an extremely complicated system to generate the picture, and (b) full Euler angle transformation between aircraft computor and display) and angular limits and computing approximations which would otherwise restrict the flight simulator work for which it could be used.

(1) Complete rolls in either direction are possible.

(2) Angles of pitch much in excess of 15° are not exceeded.

(3) Angles of yaw in excess of 20° are not exceeded.

(4) The assumed size of the squares determines the initial height. This can be pre-set at any of three initial values, 10 000ft, 1 000ft, and 100ft; the ratio of maximum to minimum height thereafter does not exceed 10:1.

With these limits in mind, it is possible to define the control which must be exercised over a presentation as seen in Fig. 2, when responding to the six degrees of freedom of an aircraft in motion. For convenience, pitch, yaw, sideslip, and vertical velocity as discussed below are measured relative to axes fixed in space (notational suffix sa) so that the two Euler transformations

pitch angle	$\theta_{sa} = \int(\theta \cos \phi - \dot{\psi} \sin \phi)$
yaw angle	$\psi_{sa} \doteq \int(\theta \sin \phi + \psi \cos \phi)$
vertical velocity	$w_{sa} = w \cos \phi - v \sin \phi$
sideslip	$v_{sa} = w \sin \phi + v \cos \phi$

must be made.

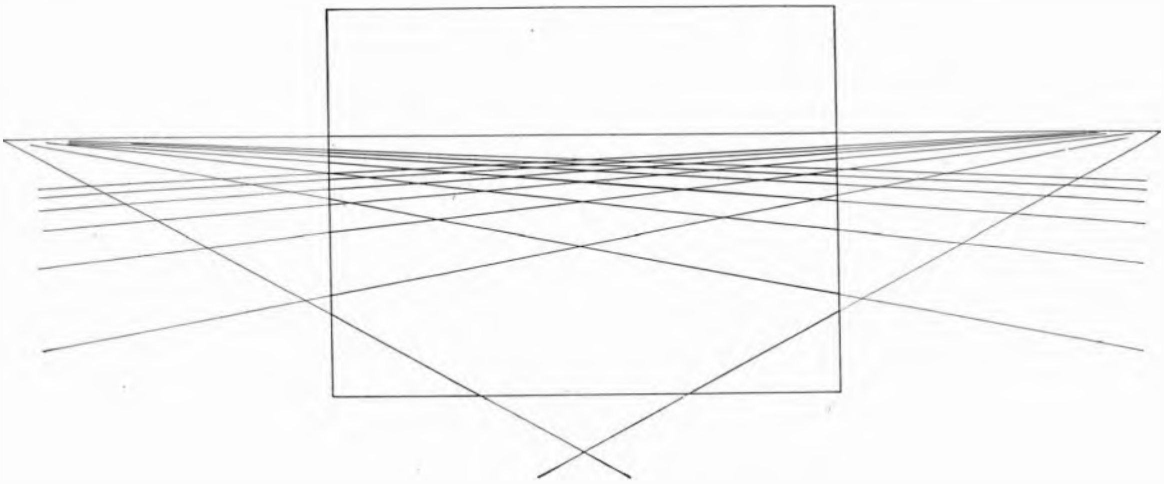


Fig. 2. ' Chessboard ' perspective display

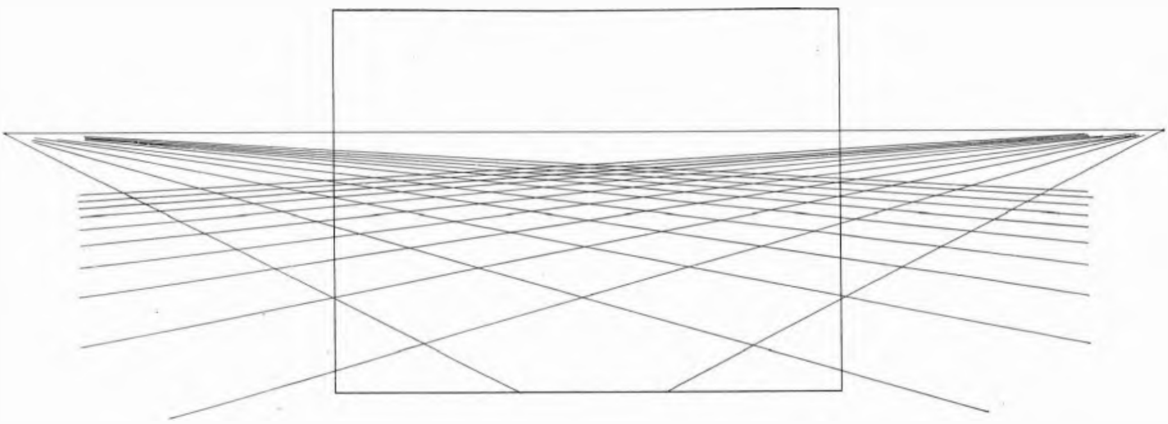


Fig. 3. Effect of height change

(a) *Roll.* Obtained by rolling the whole display about a fixed point.

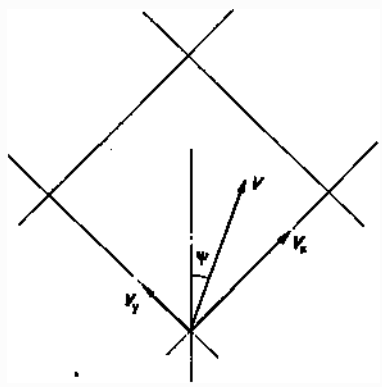
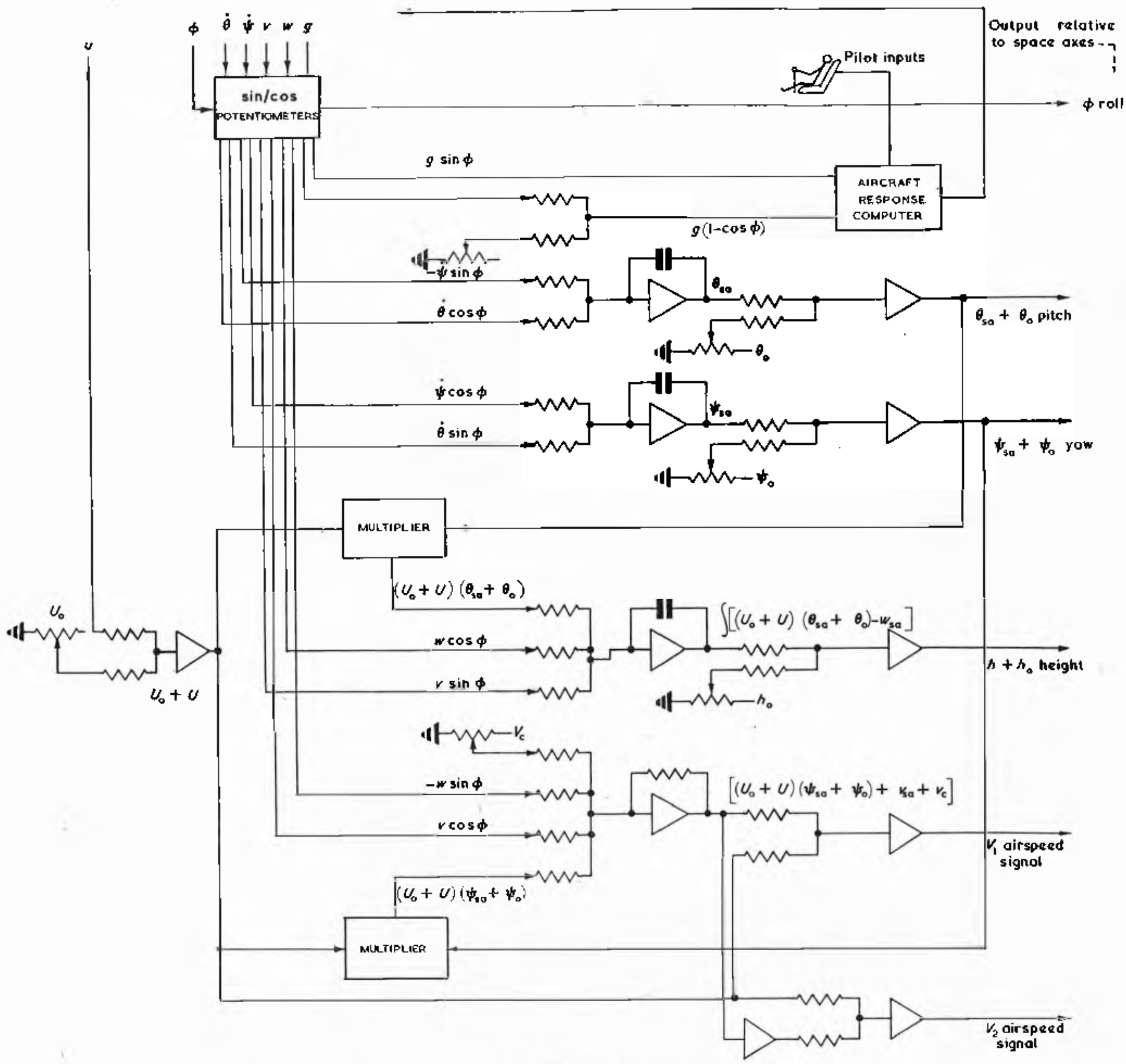


Fig. 4. Resolution for change in direction

(b) *Pitch.* Obtained by deflecting the whole display vertically (relative to the horizon).

(c) *Yaw.* Obtained by moving the vanishing points



Inputs		Setting		Computations
Roll angle	ϕ	Initial pitch	θ_0	$\theta_{sa} = [\dot{\theta}\cos\phi - \dot{\psi}\sin\phi]$
Rate of pitch	$\dot{\theta}$	Initial yaw	ψ_0	$\psi_{sa} = \int[\dot{\theta}\sin\phi + \dot{\psi}\cos\phi]$
Rate of yaw	$\dot{\psi}$	Initial forward velocity	U_0	$v_{sa} = [v\cos\phi - w\sin\phi]$
Incremental forward velocity	U	Crosswind velocity	v_0	$w_{sa} = [v\sin\phi + w\cos\phi]$
Side velocity	v	Initial height	h_0	
Downward velocity	w			

$$h = h_0 + \int[(U_0+U)(\theta_{sa}+\theta_0) - w_{sa}]$$
$$V_1 = (U_0+U)(1+\psi_{sa}+\psi_0)+(v_{sa}+v_c)$$
$$V_2 = (U_0+U)(1-\psi_{sa}-\psi_0)-(v_{sa}+v_c)$$

Fig. 5. Computation of signals for the display

along the horizon, distances measured from the centre of the display proportional to $\tan(45°+\psi)$ and $\tan(45°-\psi)$. In practice, the vanishing points are not generated, but an equivalent effect is produced by a method described later, which is valid for yaw angles less than $\pm 20°$.

(d) *Height.* The presentation in Fig. 2 is a slant view of perpendicular coplanar lines, and can be regarded as the appearance of square fields viewed from an aircraft.

Flying at a height of h miles, if the pilot views a point R miles in front of him, the angular depression of this point below the horizon is $\alpha = \tan^{-1} h/R$. (Assuming a 'flat' earth. The horizon seen from an aircraft is, of course, at a finite distance because of the earth's curvature, but investigation has shown that display errors assuming a flat earth only become appreciable at high altitudes—say 40 000ft). If the separation of the lines is varied according to this relationship the effect of change of height is obtained. This effect is seen in Fig. 3.

At a height of two miles (the maximum it is desired to simulate with this display) a total range of 100 miles is reasonable and to get a fair amount of information from the display, 20 lines are adequate. Thus the squares should have diagonals of five miles. This fixes the general requirement for the content of the picture. Of course, the same picture will represent equally well the case of $h = 0{\cdot}2$ miles and squares with a diagonal of 0·5 miles; this fact is used when switching from one height range to another.

The height h of an aircraft after time t, with the initial conditions h_o and θ_o is

$$\int_o^t [V(\theta_{sa} + \theta_o) - w_{sa}] + h_o$$

where θ_{sa} = pitch angle, w_{sa} = vertical velocity. This signal is computed from the aircraft response computer outputs, and is used as the height control.

(e) *Forward speed.* By rotating the set of lines from both sources at the same speed, each line moving faster as it comes down the picture, an impression of forward speed is obtained. The speed represented depends upon the size of the square which the display represents.

It is also clear that if one set of lines moves faster than the other, the apparent direction of motion will be no longer directly ahead.

Considering an aircraft flying with velocity V over square fields at an angle ψ to the diagonal (Fig. 4), the relative speed of each set of lines V_x and V_y will be

$$V_x = V\cos(45 - \psi) = \frac{V(\cos\psi + \sin\psi)}{\sqrt{2}} \simeq \frac{V(1+\psi)}{\sqrt{2}}$$

$$V_y = V\sin(45 - \psi) = \frac{V(\cos\psi - \sin\psi)}{\sqrt{2}} \simeq \frac{V(1-\psi)}{\sqrt{2}}$$

assuming angles of yaw less than 20°. Hence to change the apparent direction of motion by an angle ψ, one must increase the speed input to one set of lines by $V\psi$, and decrease the other by $V\psi$.

If an aircraft travelling at velocity V yaws through an angle ψ without sideslip, the change in direction of motion is also ψ; consequently when simulating yaw, the individual inputs to the speed controls V_1 and V_2 must be $V(1 + \psi)$ and $V(1 - \psi)$.

If V_o and ψ_o are initial conditions, and u and ψ_{sa} are computed increments of speed and yaw,

$$V_1 = (V_o + u)(1 + \psi_{sa} + \psi_o)$$
$$V_2 = (V_o + u)(1 - \psi_{sa} + \psi_o)$$

(f) *Sideslip.* Sideslip occurs when the flight path of the aircraft does not lie in the plane perpendicular to the aircraft's y axis, and is thus manifest as a change in direction of motion of the terrain. Since the display uses space axes, the effect of a cross-wind is similar: if v_{sa} is the sideways velocity of the aircraft, and v_c is the cross-wind component the angular change in direction of motion is $(v_{sa} + v_c)/(V_o + u)$.

Hence:

$$V_1 = (V_o + u)(1 + \psi_{sa} + \psi_o) + (v_{sa} + v_c)$$
$$V_2 = (V_o + u)(1 - \psi_{sa} - \psi_o) - (v_{sa} + v_c)$$

The manner in which outputs from the aircraft response computer must be transformed to provide signals for this type of display is shown in Fig. 5. Three servo multipliers are required, although an assumption of constant forward speed reduces this number to two. This computation does not present any great difficulties, but close integration with display generation circuits is necessary.

Display Generation—Electronic Techniques

The lines are generated using magnetic deflexion, so that by servo-driving the deflexion coils round the neck of the tube, roll angle may be impressed on the display. It is thus necessary to use time-sharing techniques to draw the two sets of lines. In order to avoid flicker effects, it is then necessary to have a basic line frequency around 2kc/s, which provides about 40c/s repetition of each set of lines (allowing some for 'frame' flyback, etc.). It has been found experimentally that when viewing such a display as this from 17in flicker becomes objectionable at lower frequencies.

The lines are drawn using the usual technique of simultaneous vertical (Y) and horizontal (X) sawtooth deflexion. The X sawtooth is generated by a Phantastron triggered at 2kc/s running at constant amplitude and having a sawtooth duration of about 100μsec. The screen waveform of this is used to gate a precision Y sawtooth generator which is also modulated in amplitude by the output of a circuit generating a voltage proportional to h/R. This h/R generator is triggered by every 24th line pulse so that 20 lines appear on the picture, allowing 4 for flyback and accommodation of the velocity input to be described later.

The h/R curve has a very high slope at close range, and in fact changes amplitude by a large amount during the line period (100μsec), and it is necessary therefore to sample the h/R voltage at the start of each line, and to

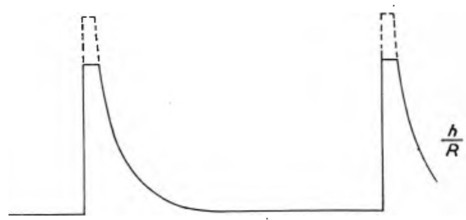


Fig. 6. The h/R waveform

hold the sample for the line duration so that constant input is provided at the modulating terminal of the Y sawtooth generator. Varying the scale or amplitude of the h/R curve proportional to h obviously gives the effect of change in height. This curve is not easy to generate, being hyperbolic, and is further complicated by the requirement of a 10:1 change in amplitude. Fortunately, the pilot of an aircraft cannot usually see more than about 15° below the aircraft longitudinal datum, so that one is not interested in zero range.

Therefore, provided one can generate the curve with sufficient accuracy at the lowest height of interest, and scale up the amplitude proportional to height, it does not matter if the system overloads at high height, low range, provided that the maximum value of h/R can be accommodated. The curve is actually approximated by the summation of three exponentials of different time-constants. This gives a reasonable approximation. Fig. 6 shows the h/R waveform.

To be able to display impressions of forward speed, it is necessary to move the lines on the picture. This is done by shifting the relative timing of the basic 2kc/s line pulses and the h/R trigger. A delay Phantastron is triggered every 24th line pulse (two Phantastron dividers divide by 6 × 4) and the trailing edge of this Phantastron screen wave is used to trigger the h/R generator.

If now the delay period is made to increase each frame by an amount proportional to the distance travelled by the aircraft in the period between frames, the lines will appear to move at the aircraft speed. The modulating signal for the delayed period is derived by integrating the forward speed signal from the main computer. To preserve a continuous picture, this integrator is 'zeroed' as soon as the delay has been increased by one period between lines so that one line on the picture will move down until it reaches the position from which the next lower line started, and then fly back to start again. The lines are so close together at the horizon that the movement up and down of the horizon line is not noticeable.

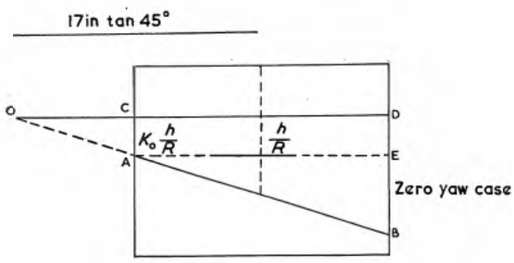


Fig. 7. The zero yaw case

Time sharing for the two sets of lines is accomplished as follows: First, an inverted X sawtooth wave is produced by an inverting amplifier, and an electronic switch driven from a bistable multivibrator triggered by the delay Phantastron output is used to switch alternately the X and inverted X sawtooth waves to the X deflexion amplifier.

Sideslip is produced by varying the velocity of one set of lines with respect to the other. In fact, two separate velocity integrators are used—one for each set of lines—(inputs from the main computer)—and the delay Phantastron is switched by a high-speed relay from one to the other. The relay is driven from the bistable multivibrator. Thus the delay Phantastron produces delay pulses alternately from one integrator and the other.

Further switching is necessary in the yaw system which must now be described. In Fig. 2, the impression is given that the picture is generated from the two vanishing points. The same effect can be (and is) obtained as seen in Fig. 7. Instead of generating the triangle OBD and only looking at the section AB of OB, the triangle ABE may be generated and the rest of the deflexion (AC) be provided by the sampled value of the h/R curve. A more general case is drawn in Fig. 8.

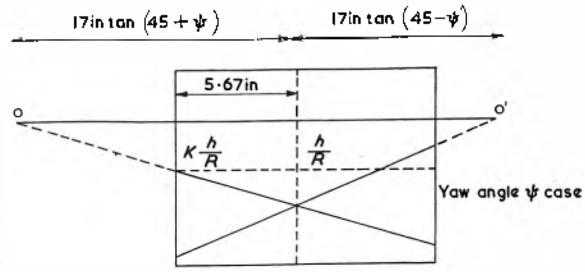


Fig. 8. The case for yaw angle ψ

By similar triangles:

$$\frac{17 \tan (45 + \psi)}{5{\cdot}67} = \frac{h/R}{(1 - K)h/R}$$

i.e.

$$\tan (45 + \psi) = \frac{1}{3(1 - K)}$$

i.e.

$$K = 1 - \frac{1}{3 \tan (45 + \psi)} = 1 - \frac{1 - \tan \psi}{3(1 + \tan \psi)}$$

$$= \frac{2 + 4 \tan \psi}{3(1 + \tan \psi)} = 2/3 \frac{1 + 2 \tan \psi}{1 + \tan \psi}$$

The sawtooth peak amplitude is constant for varying ψ at $2h/3R$. Thus, as well as reducing the amplitudes of sawtooth waves required, this system gives a method of varying yaw angle, by varying K. This is done by a servo function generator. The expression for K for the other set of lines is obtained by substituting $-\psi$ for $+\psi$, an electronic switch is used to switch the two Kh/R inputs to the vertical deflexion amplifiers.

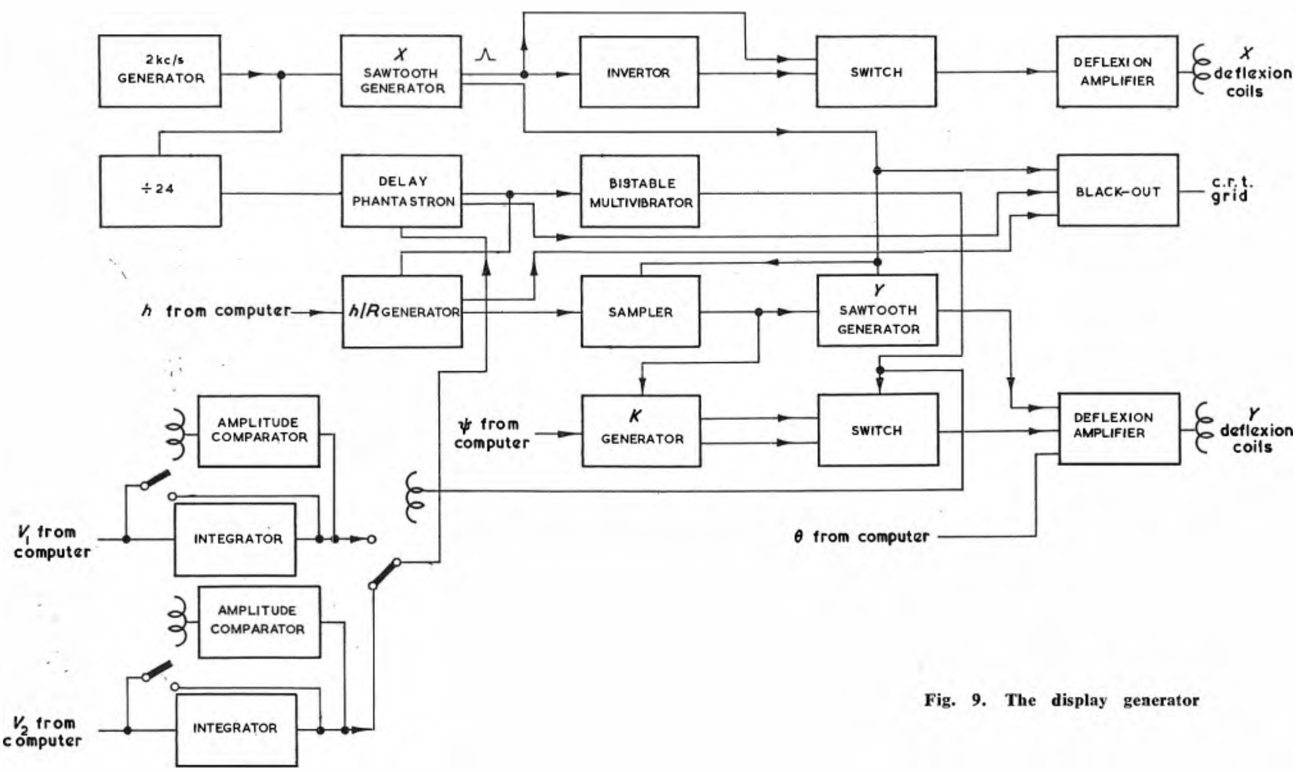


Fig. 9. The display generator

The generation of the K function would be more difficult for angles $\psi \to 45°$, but $\pm 20°$ variation in ψ is sufficient for the requirements of this simulator.

The remaining input, pitch angle θ, is limited to about $\pm 15°$, and vertical deflexion proportional to θ is sufficiently accurate.

Fig. 9 is a block diagram of the display system.

The Display in Use

As already seen, the display gives to the pilot information about the position, direction of motion, and orientation of his aircraft relative to the outside world. In fact, if the display were to be projected on to the windscreen of the aircraft, and its image collimated to appear to be at infinity, then the horizon of the display and the true horizon would always coincide. This means that no interpretation has to be done by the pilot to fly an aircraft or simulator using this method of presentation.

The value of this in a simulator for aerodynamic research is self-evident—simply by kicking the rudder in the simulator cockpit, the pilot is able to see on the display the horizon movements that would occur during a 'dutch roll*' in flight.

English Electric Aviation Ltd's flight simulator at Warton is so arranged that the primary factors affecting an aircraft's handling characteristics can be changed rapidly. This includes both variation in the aerodynamic response of the aeroplane simulated, and of the simulated feel on the stick. The latter is done by a mechanical system in which stick force, stick friction, circuit backlash, and inertia may be changed. To alter the aerodynamic characteristics simply means altering the appropriate terms in the equations of motion solved by the general purpose analogue computer LACE[2]. Simulation of power control characteristics, such as time lag, valve overlap, and rate limiting, is also done electronically.

One possible investigation might be to determine pilot comment about the handling of an aircraft when wing tip stores are added. This would involve increasing the inertia in roll in the equations of motion, by turning up the appropriate coefficient. A rapid comparison of the two configurations could be made by a number of pilots. Then, if some change in the aileron feel or control system was felt to be necessary, this could be investigated in a similar manner.

Over the past few years, a large number of useful investigations have been carried out. These include general work on desirable short-period characteristics, handling characteristics of the Lightning aircraft at high speed, feel system design for new aircraft, control circuit instabilities due to bob-weights, and stability and response augmentation and their effect on tracking errors. With the experience gained, work is proceeding to obtain information from the flight simulator on approach and landing, and also take-off characteristics. This kind of work, besides requiring the solution of much more complex and non-linear equations, obviously calls for a display which shows both movement and perspective.

A 'dutch roll' is the characteristic short period lateral oscillation of an aircraft, involving yaw, roll and sideslip, which is excited either by rudder application, or a lateral gust.

Conclusions

This display was designed for a particular purpose, and has been most useful. Other systems, using closed circuit television, have been developed elsewhere which can be used in a similar maner, and are potentially capable of much wider application. The Naish system[3] at I.A.P., R.A.E., is an example of this. The display which has been described, however, is both cheap and compact, since it does not involve any optical system. A further advantage is that the mechanical servos used are small, so that adequate frequency response to simulate high-performance aircraft is easily obtained.

Acknowledgment

The authors wish to thank English Electric Aviation Ltd for permission to publish this article.

REFERENCES

1. HOOVER, G. W. Integrated Instruments for Jet Transport. SAE reprint 40C. April 8-11th (1958).
2. GOMPERTS, R. J., RIGHTON, D. W. LACE (The Luton Analogue Computing Engine). *Electronic Engng. 29*, 380 (1957).
3. Simplying the Pilot's Task. *Aeroplane. 97*, 414 (1959).

Automatic Cheque Sorting

A demonstration of automatic cheque handling using E.M.I.'s Figure Reading Electronic Device—FRED—in conjunction with the Pitney-Bowes/National high-speed document sorter. was given recently.

In a typical day's voucher clearing in the United Kingdom over two million cheques are sorted for return to about 10 000 branches. Manual sorting methods are finding it difficult to keep up with the increasing numbers of cheques being used, so the banking industry is studying the advisability of installing electronic figure reading and sorting auipment.

FRED has been developed as a numeral reading machine specifically for use with bank cheques and for other similar business purposes.

All the data normally found on a cheque—such as serial number, branch number, customer's identity and amount—can be coded along the bottom edge using the FRED numerals.

Cheques, in sizes ranging from 6in by 3in to 8in by 4in can be sorted indiscriminately and other special sizes can be handled if required.

Accurate spacing of the characters along a line, or the precise positioning of any character relative to the leading edge of a cheque in the sorter is not necessary, and normal printing tolerances are sufficient.

The FRED system makes very few demands on the cheque printing industry, so its adoption would result in keeping cheque printing charges down to a minimum. E.M.I. has carried out extensive tests with hundreds of thousands of cheques from all the leading British cheque printing houses to confirm FRED's practicability.

Some of the data—such as customer's account number and amount—has to be imprinted on the cheques in the banks, so tests have also been carried out in collaboration with the leading manufacturers of imprinting machines.

A reading machine intended for use in business offices should be very reliable, fast, compact and of low cost. While it is relatively easy to meet these requirements with a bar code reader, since the detection of such a simple " pattern ' is a comparatively straightforward problem, it is much more difficult with complex shapes such as conventional Arabic numerals.

This difficulty has been overcome in the FRED system by building a five-element bar code into the design of the numerals. The resulting typefaces are visually similar to conventional designs—they are in fact based on an existing fount called Broadway—and are therefore easily recognized by the human eye. Furthermore, the machine is presented with what amounts to a simple 5-unit bar code, which, with magnetic ink printing, leads to very reliable machine recognition. The method of operation of FRED has been described in some detail in previous issues of this journal[1,2].

REFERENCES

1. 'FRED'—A Figure Reading Electronic Device. *Electronic Engng. 31*, 45, (1959).
2. YOUNG, D. A. Automatic Character Recognition. *Electronic Engng. 32*, 2 (1960).

The Role of Space Charge Waves in Modern Microwave Devices

(Part 1)

By G. D. Sims* and I. M. Stephenson*

This article is intended as an introduction to the properties of space charge waves and the way in which they play a part in the operation of certain modern electronic devices. Part 1 deals with the fundamental characteristics of space charge wave propagation and shows how the waves can be considered in terms of a transmission line analogy. Part 2 discusses how the space charge waves are modified by different launching conditions and different beam symmetries, while Part 3 indicates the part played by simple space charge waves, ion waves and cyclotron waves in backward wave oscillators, travelling wave tubes, parametric amplifiers and simple plasmas.

(Voir page 460 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 465)

SOUND propagates in air by means of longitudinal waves. This type of wave propagation, i.e. the passage of a series of compressions and rarefactions through the medium, is also experienced in numerous other practical fields, and is most easily envisaged when the medium is readily compressible. An electron gas possesses this property—if it is not too dense—and hence longitudinal waves can propagate through it. The propagation characteristics of the waves depend critically on the electron density and the reason why this should be so is readily apparent. If the electron gas is so tenuous that the influence of each electron on its neighbours is small, then clearly the motion can be analysed by considering the kinematics of a single electron. If on the other hand the gas is dense, each electron will be conscious of the fact that there are others around it and a single electron theory will yield inaccurate results. This argument can be summarized by saying that in any electron gas there will be some critical density at which the electrons start to behave as a medium. The waves propagated in the electron gas under these conditions are called 'space charge waves'. In other contexts they have been called electro-mechanical waves, the term arising from the fact that the energy process involved is one of continual interchange of energy at a point in the wave between kinetic and electric field forms.

Space charge waves are of significance in the discussion of all high frequency valves and have recently assumed a growing importance in the theory of plasmas. The theory of space charge waves was developed as long ago as 1939 by Ramo[1] and Hahn[2] working independently. It is only relatively recently, however, that their work has been related to the behaviour of microwave valves in particular. Theories of valve operation have, generally in the first instance, been of the ballistic (one electron) kind and the effects of space charge have invariably been ignored. Frequently these first theories have been extremely useful, but invariably as higher power valves have been developed, higher density electron beams and clouds have been used and in consequence 'space charge' effects have ceased to be negligible. The statement that space charge 'has ceased to be negligible' amounts to the same thing as the previous statement that a dense electron gas exhibits a medium like behaviour. It is thus logical in the case of finite space charge to endeavour to interpret the phenomena which arise in terms of the space charge waves of Ramo and Hahn.

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The object of this article is to try to state simply how this has been achieved in the case of the well-known types of microwave valve and the simple plasma. The design of fast wave couplers for parametric amplifiers will also be considered, a process which is entirely dependent on an understanding of the properties of space charge waves. This article is intended as an introduction to the more comprehensive works which exist on the subject as well as to give some indication of the limits of ballistic theory in dealing with electronic problems.

The general properties and characteristics of space charge waves will first be discussed in some detail and later the practical cases in which their application is of interest will be considered.

The General Nature of Space Charge Waves

Consider an electron beam emanating from a gun in which the electron density is such that space charge effects are negligible. If the beam is subjected to a longitudinal sinusoidal modulating voltage of small amplitude at some subsequent point it is fairly obvious that the resulting disturbance on the beam will take the form of a velocity/density modulation, travelling with the same velocity as the initial beam and with a wavelength equal to that of the modulating voltage. If the beam current (and hence the electron density) is now increased, space charge effects become of importance and the picture generally ceases to be so simple. The disturbance set up in the beam now has usually to be described in terms of a pair of waves. Theory shows that one wave of the pair propagates with a velocity greater than the mean electron velocity in the beam, while the other travels with a lower velocity than the beam. These waves, the simple space charge waves of Ramo and Hahn are referred to as the 'fast' and 'slow' waves respectively. It should again be emphasized here that the appearance of the two waves is purely a consequence of the electron density being high enough to cause the beam to behave as a medium in its own right. No external circuits are required except to provide the modulating voltage. The discussion of the properties of microwave valves therefore amounts to a discussion of the interaction of these natural waves of the electron medium with the natural waves which can propagate on different types of circuit brought into proximity with the beam.

In a similar manner fast and slow space charge waves will be set up by noise voltages at the electron gun itself. The noise at the cathode, which may be considered as an infinite number of sine waves of varying frequency, sets up pairs of space charge waves (each pair corresponding to one frequency component of the noise) which travel down the beam and because of their differing phase velocities each pair of waves will interfere one with the other producing periodic maxima and minima of noise at points along the beam. This phenomena was investigated experimentally by Cutler and Quate[3] who found that noise minima did occur but that the sharpness of the minima tended to be reduced by partition noise. The noise on the electron beam varied by about 20dB from the minima to the maxima, and the variation was periodic as would be expected from the space charge wave theory.

Here again it becomes clear that if the intensity of the noise developed in a beam varies sinusoidally along the beam, a knowledge of the space charge wave properties is of the utmost importance in determining those low noise regions in which it would be most suitable to impress external modulating signals in order to obtain the maximum signal-to-noise ratio.

List of Symbols

The following suffixes are used throughout the article:

o	refers to d.c. quantities
1	,, ,, a.c. ,,
o1	,, ,, values of a.c. quantities at an input ($z = 0$)
*	denotes complex conjugate
f	refers to fast wave only
s	,, ,, slow ,, ,,
e	= electronic charge
m	= electron mass
η	= e/m
μ_o	= permeability of free space
ϵ_o	= permittivity ,, ,, ,,
D	= displacement
E	= electric field
B	= magnetic flux density
H	= ,, field
J	= current density
ρ	= charge density
x,y,z,r,θ	= co-ordinates: also used as subscripts with field components
ω	= angular frequency
ω_p	= ,, plasma frequency $(=\eta\,\rho_o/\epsilon_o)^{\frac{1}{2}}$
β	= phase constant
β_o	= ω/u_o
β_p	= ω_p/u_o
k	= $\omega^2\epsilon_o\mu_o$
u	= electron velocity
v_p	= phase velocity
V	= potential (kinetic potential $V_1 = u_o u_1/\eta$)
Z_ω	= space charge wave impedance ($= u_o/\epsilon_o\omega$)
Z_B	= d.c. beam impedance ($= V_o/J_o$)

The following symbols are used in Part 3 only

Z_{cct}	= circuit impedance
C	= Pierce's travelling wave tube gain parameter
d	= travelling wave tube circuit attenuation parameter
b	= Travelling wave tube velocity parameter
ω_c	= cyclotron frequency ($= \eta B$)

Suffixes i and e appended to frequency of phase constant symbols refer them specially to ions or electrons respectively.

All other symbols are defined in the text.

Any disturbance on the beam then generally sets up a fast and a slow wave. The wave velocities for an infinite beam are derived in Appendix (1) and are given by:

$$v_p = \frac{u_o}{1 \pm (\omega_p/\omega)} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

A travelling wave tube gun for example operating at 1kV and supplying a current of $1A/cm^2$ produces an unmodulated beam of density 3×10^9 electron/cm^3. This corresponds to a plasma frequency of approximately 500Mc/s. At 10 000Mc/s, $\omega_p/\omega = 0{\cdot}05$ and hence the velocities of the fast and slow waves differ from one another by only 10 per cent and from the beam velocity by only 5 per cent. The difference is thus not large and whereas in some classes of problem this fact enables ballistic theory to be applied over a wider range than might be expected, in others, particularly in the design of fast wave couplers the smallness of the difference will be seen to be a considerable disadvantage.

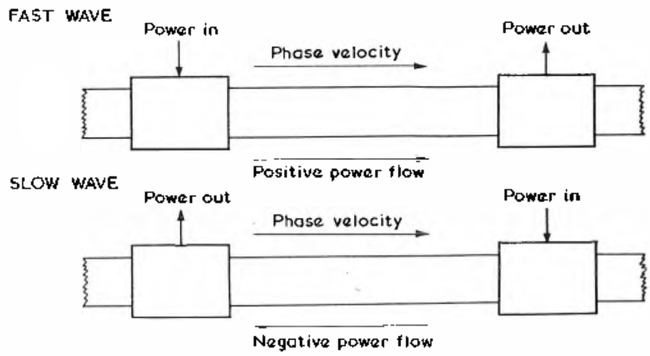

Fig. 1. Power flow in separate slow and fast waves

The electron beam is unmodulated before entering the input couplers and after leaving the output coupler. The electron flow is from left to right

Kinetic Power Flow

The waves differ not only in velocity but in that for the fast mode, the current and voltage in the wave are in phase, whereas for the slow mode, current and voltage are 180° out of phase. This leads to the apparently curious conclusion that the slow mode carries a negative power. This can be more clearly understood in terms of the following argument:

Suppose an electron beam of velocity u_o is subjected to a modulating voltage which superposes a velocity modulation u_1 on the beam. Any element of the beam will then possess a velocity $u_o + u_1$ which can be considered as having been attained by falling through a potential $V_o + V_1$. One may write:

$$e(V_o + V_1) = \tfrac{1}{2}\,m(u_o + u_1)^2 \quad \ldots\ldots\ldots (2)$$

and hence if $u_o \gg u_1$:

$$V_1 = (u_o u_1/\eta) \quad \ldots\ldots\ldots\ldots\ldots (3)$$

The quantity $u_o u_1/\eta$ is referred to as the 'kinetic voltage' of the beam.

The actual mean ballistic power carried by the electron beam, by analogy with normal transmission line theory will thus be:

$$\tfrac{1}{2}\,\mathrm{Re}[(V_o + V_1)\,J^*] \quad \ldots\ldots\ldots\ldots (4)$$

The first term of the product represents the mean d.c. power in the beam while the second:

$$\tfrac{1}{2}\,\mathrm{Re}(V_1 J^*) \quad \ldots\ldots\ldots\ldots\ldots (5)$$

represents a flow of power relative to the mean electron stream. This is referred to as the mean kinetic power flow. Since for the slow mode V_1 and J^* are out of phase, the mean kinetic power flow is negative while for the fast mode it is positive.

Power flow in the separate waves is illustrated in Fig. 1.

For a single component periodic modulating field, the total power flow in the beam will be given by[4]:

$$P_{\text{total}} = \tfrac{1}{2}\,\text{Re}[E x H^* - (m/e)\; u_o u_1 J^*] \;\ldots\ldots\; (6)$$

being the sum of the electromagnetic and kinetic power flows.

Because the velocity of the slow wave is less than that of the mean beam velocity the slow wave may be considered as propagating backwards with respect to the mean electron flow and with this idea in mind the concept of the negative power carried is not so difficult to grasp. It should be emphasized, however, that with respect to a stationary observer both fast and slow waves propagate forwards, in the same direction as the mean electron flow, unless $\omega < \omega_p$.

Wave Impedance

The continuity equation for any uniform one dimensional system states that:

$$\partial \rho / \partial t = -(\partial J / \partial z) \;\ldots\ldots\; (7)$$

and in this case where all a.c. quantities vary as $e^{j(\omega t - \beta z)}$ leads to the relation:

$$\rho_1 = (\beta/\omega)\, J_1 \;\ldots\ldots\; (8)$$

However:

$$J = J_o + J_1 = \rho u = (\rho_o + \rho_1)(u_o + u_1) \;\ldots\; (9)$$

whence:

$$J_1 \simeq \rho_o u_1 + u_o \rho_1 \;\ldots\ldots\; (10)$$

Thus from equations (8) and (10)

$$J_1 = \frac{\omega \rho_o}{\omega - \beta u_o} u_1 \;\ldots\ldots\; (11)$$

Now:

$$V_1 = (u_o u_1/\eta) = u_o/\eta \left(\frac{\omega - \beta u_o}{\omega \rho_o}\right) J_1 \;\ldots\; (12)$$

But $\omega - \beta u_o = \pm \omega_p$ the positive sign describing the fast mode and the negative, the slow one.

Therefore:

$$V_1 = (u_o/\omega)\left(\pm(\omega_p/\rho_o \eta)\right) J_1 = \pm(u_o/\epsilon_o \omega)\, J_1 \;\ldots\; (13)$$

Thus the ratio of kinetic voltage and current defines a wave impedance $|Z_\omega| = u_o/\epsilon_o \omega$ with which a positive sign has to be associated for the fast wave and a negative sign for the slow wave.

Space Charge Waves and Transmission Lines

It has been shown that when an electron beam becomes sufficiently dense, it behaves as a medium, capable of propagating pairs of waves whose velocities are greater and less than that of the beam. These waves are fully described by their kinetic voltage $V_1 = (u_o u_1/\eta)$ and their current J_1. In both cases the magnitude of the characteristic impedance is $u_o/\epsilon_o \omega$ being positive for the fast, and negative for the slow wave. Both waves propagate with an exponential factor $e^{-j\beta z}$ where β has the value $\beta_f = \beta_e - \beta_p$ for the fast wave and $\beta_s = \beta_e + \beta_p$ for the slow wave where $\beta_e = (\omega/u_o)$ and represents the propagation constant for negligible space charge and $\beta_p = (\omega_p/u_o)$ represents the amount by which this constant is modified by the presence of space charge for either wave.

Since the ratio of voltage and current is constant and since either wave has exponential dependence on z, each separate space charge wave will resemble a simple mode on a transmission line and the pair of modes (fast and slow) propagating on the beam can be represented in terms of two coupled transmission lines. In analysing the behaviour of a microwave valve for instance a suitable equivalent circuit from which to work will thus be one in which the valve circuit is coupled with a further transmission line representing either the fast or slow wave or both.

Boundary Conditions

So far it has been stated that in general when an electron beam is subjected to a sinusoidal modulation a pair of space charge waves will in general be set up. It is appropriate at this stage to examine how the total voltage and current, together with the amplitudes of the fast and slow modes depend upon the initial conditions.

Suppose that at the plane $z = 0$ at which modulation is assumed to take place:

The initial current is J_{10}

" " kinetic voltage V_{10}

" " $\left.\begin{matrix}\text{fast}\\\text{slow}\end{matrix}\right\}$ wave current amplitude is $\begin{cases}J_{fo}\\J_{so}\end{cases}$

" " $\left.\begin{matrix}\text{fast}\\\text{slow}\end{matrix}\right\}$ wave kinetic voltage amplitude is $\begin{cases}V_{fo}\\V_{so}\end{cases}$

Then:

$$V_{10} = V_{fo} + V_{so} = (J_{fo} - J_{so})\, Z_\omega \;\ldots\ldots\; (14)$$

$$J_{10} = J_{fo} + J_{so} = (V_{fo} - V_{so})\, 1/Z_\omega \;\ldots\; (15)$$

Missing out the exponential time factor, which is common to all terms, the total current at a distance z can be written as :

$$J_1(z) = J_{fo} \exp[-j(\beta_e - \beta_p)z] + J_{so} \exp[-j(\beta_e + \beta_p)z]$$
$$= [(J_{fo} + J_{so}) \cos \beta_p z + j(J_{fo} - J_{so}) \sin \beta_p z] \exp(-j\beta_e z) \;\ldots\ldots\; (16)$$

Using equations (14) and (15) this yields:

$$J_1(z) = [J_{10} \cos \beta_p z + j(V_{10}/Z_\omega) \sin \beta_p z]\; \exp(-j\beta_e z) \;\ldots\; (17)$$

Similarly the total kinetic voltage is found to be:

$$V_1(z) = [V_{10} \cos \beta_p z + jZ_\omega J_{10} \sin \beta_p z]\; \exp(-j\beta_e z) \;\ldots\; (18)$$

Since the transmission line analogy has been implicitly assumed in this analysis it is not surprising that these equations resemble others familiar to those working in the transmission line field.

From equations (14) and (15) it is observed that the amplitude moduli of the kinetic voltages for the fast and slow waves will be given by:

$$V_{fo} = \tfrac{1}{2}(V_{10} + Z_\omega J_{10}) \;\ldots\ldots\; (19)$$

$$V_{so} = \tfrac{1}{2}(V_{10} - Z_\omega J_{10}) \;\ldots\ldots\; (20)$$

in other words the mode amplitudes will be comparable for small values of the product $Z_\omega J_{10}$. A little manipulation shows that Z_ω can also be written in the form:

$$Z_\omega = 2(\omega_p/\omega)\, Z_B \;\ldots\ldots\; (21)$$

where Z_B is the d.c. beam impedance. Since at microwave frequencies ω_p/ω will usually be small, the product $Z_\omega J_{10}$ will be correspondingly small and in consequence V_{fo} and V_{so} will be of the same order of magnitude. If it is desired to preferentially launch the fast wave rather than slow mode ω_p and J_{10} must be as large as possible. Clearly then high space charge density is required for this as well as appreciable current modulation. Considerations of this kind are discussed in detail in a later section as they are of considerable importance in the design of couplers for electron beam parametric amplifiers.

Velocity Jumps

In theory it is possible to completely remove either the fast or the slow space charge mode by means of a suitably located velocity jump. If at some point Z_1, the beam voltage is suddenly increased, the space charge wave impedance Z_ω will increase proportionately.

Suppose that for $z < z_1$, the mode impedance is $Z_{\omega 1}$

" " " $z < z_1$, " " " " $Z_{\omega 2}$

From equations (17) and (18) the kinetic voltage and current at z are:

$$V(z_1) = [(V_{10}\cos\beta_p z_1 + jZ_{\omega 1}J_{10}\sin\beta_p z_1)]\exp(-j\beta_{e1}z_1) \quad \ldots\ldots\ldots\ldots (21)$$

$$J(z_1) = [(J_{10}\cos\beta_p z_1 + j(V_{10}/Z_{\omega 1})\sin\beta_p z_1)]\exp(-j\beta_{e1}z_1) \quad \ldots\ldots\ldots\ldots (22)$$

while beyond z_1 the voltage and current will be given by expressions of the form:

$$V(x) = [(V(z_1)\cos\beta_p x + jZ_{\omega 2}J(z_1)\sin\beta_p x)]\exp(-j\beta_{e2}x) \quad \ldots\ldots\ldots\ldots (23)$$

$$J(x) = [(J(z_1)\cos\beta_p x + j(V(z_1)/Z_{\omega 2})\sin\beta_p x)]\exp(-j\beta_{e2}x)$$

where $x = z - z_1$.

Now equation (23) may be re-written:

$$V = \tfrac{1}{2}(V(z_1) + Z_{\omega 2}J(z_1))\exp(-j(\beta_{e2}-\beta_p)x) + (V(z_1) - Z_{\omega 2}J(z_1))\exp(-j(\beta_{e2}-\beta_p)x) \quad \ldots\ldots (24)$$

and thus it is seen that if the magnitude and position of the jump are so chosen that either $(V(z_1) \pm J(z_1)Z_{\omega 2}) = 0$ then the disturbance beyond the jump will be in the form of a pure slow, or a pure fast wave according to whether the upper or lower sign is taken.

If the ratio of slow wave voltage to fast wave voltage at the input ($z = 0$) is written as:

$$V_{so}/V_{fo} = W e^{j\phi} \quad \ldots\ldots\ldots\ldots (25)$$

it is shown in Appendix (2) that the slow wave can be eliminated if:

(1) The position of the jump is chosen such that

$$\beta_p z_1 = \frac{\phi + n\pi}{2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (26)$$

where n is a positive or negative integer.

(2) $W < 1$, i.e. the initial fast wave amplitude is greater than the initial slow wave amplitude.

(3) $$|Z_{\omega 2}| = |Z_{\omega 1}|\left\{\frac{1 + (-1)^n W}{1 - (-1)^n W}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (27)$$

This latter condition fixes the magnitude of the velocity jump required and it is noted that the jump may be either upwards or downwards according to whether n is odd or even.

Conversely it may be shown that the fast wave is eliminable subject to the same conditions except that W must be greater than 1.

Appendix (1)

Theory of Space Charge Waves in a Uniform Infinite Electron Stream

In this consideration the negative space charge density due to electrons in the unmodulated beam is assumed to be neutralized by positive ions so that space charge divergence forces may be neglected.

Maxwell's equations for a region containing free charges state that:

$$\nabla \times E = -\dot{B} \ldots\ldots (A1) \qquad \nabla . D = \rho \ldots\ldots (A2)$$

$$\nabla \times H = \dot{D} + J \ldots . (A3) \qquad \nabla . B = 0 \ldots\ldots (A4)$$

For a weakly modulated stream:

$$J = J_o + J_1; \qquad \rho = \rho_o + \rho_1 \text{ etc.,}$$

where $J_1 \ll J_o$ $\qquad \rho_1 \ll \rho$

and all the quantities with suffix 1 vary as $\exp j(\omega t - \beta z)$. The equation of motion of an electron states that in the absence of magnetic field,

$$m\,(du/dt) = eE \quad \ldots\ldots\ldots\ldots\ldots (A5)$$

The continuity equation states that for an infinite stream propagating in the z-direction,

$$\partial\rho/\partial t = -(\partial J/\partial z) \quad \ldots\ldots\ldots\ldots\ldots (A6)$$

Also $J = \rho u = J_o + \rho_1 u_o + u_1\rho_o$ to a first order $\ldots\ldots$ (A7)

In this system, considering the a.c., z – component of electric field it is found from equations (A1) and (A3) that:

$$\nabla \times \nabla \times E_1 = -\nabla \times \dot{B}_1 = -j\omega\mu\,(j\omega\epsilon E_1 + J_1)$$

Thus, grad. div E_1 – div. grad $E_1 = \omega^2\epsilon\mu E_1 - j\omega\mu J_1$

and using equations (A2) and (A7) it is found that:

$$\nabla^2 E_1 + \omega^2\epsilon\mu E_1 = j\omega\mu(\rho_1 u_o + u_1\rho_o) + \text{grad}\,(\rho_1/\epsilon_o) \ldots . (A8)$$

Equation (A5) can be rewritten:

$$(\partial u_1/\partial t) + (dz/dt)\,(\partial u_1/\partial z) = \eta E_1$$

whence, since $dz/dt = u_o$ to a first order:

$$u_1 = \frac{\eta}{j(\omega - \beta u_o)} E_1 \quad \ldots\ldots\ldots\ldots\ldots (A9)$$

From equation (A6):

$$\rho_1 = (\beta/\omega)\,J_1 \quad \ldots\ldots\ldots\ldots\ldots (A10)$$

Substituting for J_1 in equation (A10) from equation (A7), and eliminating u_1 using equation (A9) it is found that:

$$\rho_1 = -j\beta\epsilon_o\left(\frac{\omega_p}{\omega - \beta u_o}\right)^2 E_1 \quad \ldots\ldots\ldots (A11)$$

Thus eliminating ρ_1 and u_1 between equations (A8), (A9) and (A11), and rearranging, the following equation is obtained for E_1, viz:

$$\nabla_t^2 E_1 + (k^2 - \beta^2)\left(1 - \left(\frac{\omega_p}{\omega - \beta u_o}\right)^2\right)E_1 = 0 \quad .. (A12)$$

where $k^2 = \omega^2\epsilon\mu$ and $\nabla_t^2 = (\partial^2/\partial x^2) + (\partial^2/\partial y^2)$ (or the equivalent transverse Laplacian in any other set of coordinates). Now in an infinite uniform stream $(\partial^2/\partial x^2) = (\partial^2/\partial y^2) = 0$ and hence either

$$(k^2 - \beta^2) = 0 \quad \ldots\ldots\ldots\ldots\ldots (A13)$$

or

$$1 - \left(\frac{\omega_p}{\omega - \beta u_o}\right)^2 = 0 \quad \ldots\ldots (A14)$$

The roots of these equations are:

$$\beta = \pm k \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (A15)$$

or:

$$\beta = \frac{\omega \pm \omega_p}{u_o} \quad \ldots\ldots\ldots\ldots\ldots (A16)$$

The first two waves are those which would exist in free space in the absence of the electrons and propagating in the positive and negative z-direction respectively with the velocity of light.

The second pair are the 'space-charge waves' of Ramo-Hahn and propagate with phase velocities

$$v_p = (\omega/\beta) = \frac{u_0}{1 \pm (\omega_p/\omega)} \quad \ldots\ldots\ (A17)$$

For microwave frequencies $\omega \gg \omega_p$ and thus one wave propagates with a velocity slightly greater than the mean beam velocity and one with a velocity slightly less than this. At low frequencies the phase velocity of the fast wave becomes negative.

Appendix (2)

Elimination of the Slow Wave

For the slow or fast wave to be eliminated it has been shown that:

$$V(z_1) \pm |Z_{\omega 2}|\, J(z_1) = 0 \quad \ldots\ldots\ (A18)$$

which leads to the expression:

$$|Z_{\omega 2}| = \pm \frac{V(z_1)}{J(z_1)} = \pm \frac{(V_{10}/J_{10}) + j|Z_{\omega 1}| \tan \beta_p z_1}{1 + j\,\dfrac{V_{10}/J_{10}}{|Z_{\omega 1}|} \tan \beta_p z_1} \quad \ldots\ (A19)$$

which is formally identical with the result obtained by considering the fast and the slow mode as transmission lines and calculating the terminal impedance of the two lines in parallel at an angular distance $\beta_p z_1$ from the input. This procedure amounts to matching the pair of equivalent lines before the jump to a single line after the jump.

For elimination of either wave $Z_{\omega 2}$ must have a real positive value. If the ratio of the input amplitudes of the fast and slow waves is written as:

$$V_{s0}/V_{f0} = W e^{j\phi}$$

Then

$$V_{10} = V_{f0}(1 + W e^{j\phi})$$
$$J_{10} = (V_{f0}/|Z_{\omega 1}|)(1 - W e^{j\phi})$$

and consequently:

$$V_{10}/J_{10} = \left(\frac{1 + W e^{j\phi}}{1 - W e^{j\phi}}\right) |Z_{\omega 1}| \quad \ldots\ldots\ldots\ (A20)$$

Substituting this expression in equation (A18) it is found that for $|Z_{\omega 2}|$ to be real—the essential condition for the elimination of either wave is:

$$\beta_p z_1 = \frac{\phi + n\pi}{2}$$

where n is a positive or negative integer. This gives the value:

$$|Z_{\omega 2}| = \pm |Z_{\omega 1}| \left(\frac{1 + (-1)^n W}{1 - (-1)^n W}\right)$$

for the magnitude of the space charge wave impedance. For this result to be meaningful it should be noted that $\dfrac{1 + (-1)^n W}{1 - (-1)^n W}$ must be $\left.\begin{matrix}\text{positive}\\ \text{negative}\end{matrix}\right\}$ if the $\left\{\begin{matrix}\text{slow}\\ \text{fast}\end{matrix}\right.$ wave is to be eliminated. Thus to eliminate the slow wave $W \not> 1$, i.e. the initial fast wave amplitude must exceed that of the slow wave and conversely.

REFERENCES

1. Ramo, S. The electron wave theory of velocity modulation tubes. *Proc. Inst. Radio Engrs. 27*, 757 (1939).
2. Hahn, W. C. Small signal theory of velocity modulated electron beams. *Gen. Elect. Rev. 42*, 258 (1939).
3. Cutler, C. C., Quate, C. F. Experimental verification of space-charge and transit time reduction of noise in electron beams. *Phys. Rev. 80*, 875 (1950).
4. Louisell, W. H., Pierce, J. R. Power flow in electron beam devices. *Proc. Inst. Radio Engrs. 43*, 425 (1955).

(To be continued)

Units for Current Noise

By P. L. Kirby*, D.Sc., F.Inst.P. and R. H. W. Burkett*, B.Sc., A.M.I.E.E.

The relationship between current noise power, the voltage applied to a resistor and the frequency band of measurement is discussed. If the noise power spectrum is 1/f it is shown that a practical unit for the expression of noise is μV/V in a decade. It is proposed that such a standard be adopted. The errors arising from the assumption of a 1/f frequency relationship are calculated for the practical cases of $\overline{E_N^2} \propto 1/f^{0.8}$ *to* $\overline{E_N^2} \propto 1/f^{1.1}$ *and it is shown that such errors are not so great as to make the proposed unit impracticable.*

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 466)

The investigation of the noise generated by fixed resistors when a direct current is passed through them, and the method of measuring this noise, has been the subject of a Ministry of Aviation Development Contract. Many of the conclusions reached in this investigation have already been announced[1,2]. A tentative explanation has been put forward as to the origins of this noise, but the present position is that it is not yet fully established. Previous workers[3,4] have illustrated the dependence of this current noise on other electrical parameters, and the accepted relationship has been confirmed. Current noise generated by resistors is of the form:

$$\overline{E_N^2} = \int_{f_1}^{f_2} K\,(V^\alpha/f^\beta)\,.\,df$$

where:

E_N = open-circuit noise voltage at the terminals of the resistor
K = a constant for a given resistor
V = polarizing d.c. voltage across the resistor
f = frequency
α = voltage parameter
β = frequency parameter

The expression shows that the square of the noise voltage is a function of the voltage applied to the resistor, the frequency at which the measurement is made and the passband of the instrument used to make the measurement.

The parameters α and β vary from resistor to resistor and for film resistors the range of α has been found to be from 1·4 to 2·2 and the range of β from 0·6 to 1·1. From this it appears that the relationship between current noise and the parameters which affect it, is so complex that it is not possible to deduce the noise generated by any particular resistor from the knowledge of its resistance value and the conditions of the measurement.

It has been customary to accept the fact that resistor

* *Welwyn Electric Ltd.*

noise can vary materially within any given batch and so specifications have placed an upper limit to the noise which might be produced by a resistor of a given type. The relationship between current noise and applied d.c. voltage has been assumed to be linear so that noise has usually been expressed in microvolts per applied volt (μV/V), with some indication of the pass-band over which the measurement was made. This information is useful if the component is to be used over a pass-band corresponding to that of the instrument which was used to measure the noise. This is not always the case and thus the information is frequently of little value.

Proposals[5,6] have been made for a unit to express resistor current noise which is called conversion gain. This is related to the logarithm of the ratio of the noise power generated by the resistor to the power dissipated in it, the unit being expressed in terms of a single cycle pass-band at 1kc/s. This unit is valuable in making comparisons between types of resistors but suffers from the criticism that it is difficult to deduce from the unit the magnitude of noise generated over a given pass-band at a given centre band frequency.

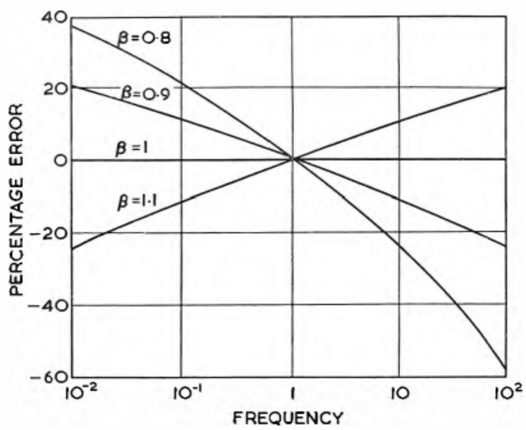

Fig. 1. Errors in noise estimation by extrapolating on the assumption $\overline{E_N^2} \propto 1/f$ *when* $\overline{E_N^2} \propto 1/f^{\beta}$

Examination of noise measurements made on many resistors indicates that the magnitude of parameter β is generally between 0·8 and 1·0 and it is only on resistors of very low noise values that the extreme of 0·6 is found. From this it is deduced that for the bulk of resistors the best design practice would be to establish an upper limit of noise signal and to work on the assumption that the relationship with frequency would involve a variation of β from 0·8 to 1·0.

If the relationship corresponds exactly to $1/f$ it has been shown[7] that the noise developed by a resistor will be constant over any pass-band in which the upper and lower frequency limits have the same ratio. Thus it would, in these circumstances, be convenient to express noise in terms of a decade of frequency.

If α were exactly 2 then it would be possible to express noise over a given pass-band as follows:

$$E_N = KV$$

where E_N = noise voltage

V = polarizing voltage

K = constant

Thus noise μV/V will be constant for a given resistor.

If the latter assumption were made in practice in the case of a resistor where α was significantly less than 2, then an error would be incurred in estimating the noise at a lower level of applied voltage. As, however, the absolute magnitude of the noise would be substantially reduced at the lower voltage the error would amount to a small *overestimation* of a much reduced signal. For practical purposes there can therefore be no objection to expressing in μV/V provided that noise is measured at the maximum working voltage of the resistor.

At a recent Working Group of Sub-Committee 40/1 of the International Electrotechnical Commission held in Washington the British proposal was made that noise should be expressed in terms of μV/V in a decade. In the case of a resistor whose noise power spectrum is governed by an exact $1/f$ relationship the magnitude of noise power quoted in these units remains constant for any decade[7], and is easily computed for any other bandwidth. This unit has the advantage that it is possible to extrapolate the noise voltage to frequencies higher or lower than that at which the noise was measured, to cover cases in which greater or lesser bandwidths are concerned. The practical application of this unit is thus a very simple matter and has advantages over both units previously proposed.

It remains to examine the errors made in the use of this unit arising from the fact that β is not always unity. Curves in Fig. 1 show the errors which will arise for various values of β using the unit on the basis that the frequency relationship is $1/f$. It will be seen that when $\beta = 0{\cdot}8$ there is an error of 37 per cent when extrapolation is carried out at a lower frequency over two decades. As it is proposed that noise will be measured in a band with a centre frequency at 1kc/s, the error in the computation of the noise level at 10c/s will be 37 per cent high. This means that the actual noise observed will be lower than the noise calculated and therefore the error is acceptable as it involves overestimation.

When extrapolating towards higher frequencies over two decades there is an error of -59 per cent, i.e. there will be an underestimation of maximum noise. The error is not negligible, but in the absence of a direct measurement the approximation will give an indication of the magnitude of the noise level.

It must be remembered that in practice bandwidths will frequently be rather less than a decade so that the noise signal will be reduced by an appropriate factor. In addition, thermal agitation noise which is proportional to bandwidth becomes of much greater importance at higher frequencies and will often predominate. For these reasons the errors at higher frequencies are considered to be acceptable.

It is therefore considered that while the unit does not give perfect extrapolation the errors that arise from the assumption of a $1/f$ spectrum are not so large as to render the unit impracticable. It has a considerable advantage over units used heretofore and gives an approximate means by which the engineer can assess the noise generated by fixed resistors at frequencies and at pass-bands substantially different from those which were employed in the instrument used for providing noise data.

REFERENCES

1. Kirby, P. L., Sibilia, G. P. An Investigation of Current Noise in Fixed Resistors. Proc. Symposium on Noise in Fixed Resistors R.R.E. (March 1959).
2. Kirby, P. L. Current Noise in Fixed Cracked-Carbon Resistors. Noise in Electronic Devices (To be published—Chapman and Hall).
3. Campbell, R. H., Chipman, R. A. Noise from Current-Carrying Resistors. *Proc. Inst. Radio Engrs. 37*, 938 (1949).
4. Rollin, B. V., Templeton, L. M. Noise in Semiconductors at Very Low Frequencies. *Proc. Phys. Soc. 66 (399B)* 258 (1953).
5. Conrad, G. T. The N.B.S. Proposed Standard Resistor Noise Test Set. N.B.S. Rep 5400 (Aug. 1957).
6. Conrad, G. T. A Proposed Current-Noise Index for Composition Resistors. *Inst. Radio Engrs. Trans. CP-3*, 14 (Mar. 1956).
7. Bell, D. A. Electrical Noise. p. 228 (Van Nostrand, 1960).

Temperature Compensated Strain Gauges

By R. L. Chandler*, B.Sc., and E. J. Dent†, B.Sc. (Eng.), A.M.I.Mech.E.

Wire-resistance strain gauges are commonly used by the research and development engineer, but they suffer from the limitation that when the temperature varies or is not accurately known an unknown error may be introduced.

This disadvantage has been overcome in the Napier Research Laboratories, where gauges have been developed which compensate for temperature change over wide ranges. These use thermocouples to develop a voltage which compensates for the signal voltage resulting from temperature changes, leaving only the signal due to the applied stress.

So far as is known the Napier self-compensating gauge, which of course is patented, is unique. Under the difficult conditions for which it is designed it provides the only means there is of making accurate strain measurements.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 466)

ONE of the most important components used in contemporary stress analysis is the electrical resistance strain gauge, in which the resistance of an element attached to a stressed surface varies with the surface strain. The fractional change of resistance is directly proportional to the strain in the element, the relationship being expressed as $dR_g/R_g=\lambda e$. The constant of proportionality, or strain sensitivity factor λ, is related to the dimensional changes in the element as well as to certain metallurgical effects due to the hydrostatic component of the stress in the element.

Unfortunately changes in temperature can cause resistive changes which are comparable to the effects of mechanical strain so that when steady, non-alternating stresses are being determined some form of temperature compensation would be essential. The text describes a method for annulling temperature effects which has been devised and developed in the Research Laboratories of D. Napier & Son Limited.

Theoretical Analysis

Temperature Compensation by Older Methods

Temperature compensation with a Wheatstone bridge network has been attempted hitherto by the installation of a second strain gauge similar in all respects to the active gauge and bonded to a similar material. The active and compensating gauges, provided they had been placed close together, would experience theoretically the same changes due to any temperature variations. If the bridge had been balanced initially:

$$R_g/R_d=R_b/R_c$$

but $(R_g+dR_g)/(R_d+dR_d) = R_b/R_c$ at temperature

where R_g, R_b, R_c and R_d represent the resistances of the bridge arms, R_g and R_d being the active and compensating gauges respectively. The ratio is unaltered since dR_a and dR_g are proportionate increases, although in practice the errors due to the difference in temperature of the active and compensating gauge would tend to be rather unpredictable. Moreover, the temperature characteristics of the two gauges cannot easily be made identical.

However, greater accuracy has been obtained in recent years by the use of a special type of strain gauge with an extremely low temperature sensitivity. The gauge grid, in this case, is constructed with two different wire materials in series, the lengths of the materials being proportioned to make the resistance increase due to temperature of one material approximately annulled by a decrease in resistance of the other material. Experimentally it has been found that improved accuracy could be obtained by the use of a compensating gauge of the same construction as the active one. Copper nickel and copper grids have been used in this connexion although the degree of compensation obtained would not satisfy most accuracy requirements, due mainly to the great

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Nomenclature

a	= total length of wire in strain gauge
α	= the apparent temperature coefficient of the active gauge and base
α_1 and α_2	= the temperature coefficient of resistance of lead materials
α'	= the apparent temperature coefficient of the dummy gauge and base
α''	= the apparent temperature coefficient of the active gauge arm
α_r	= temperature coefficient of the strain gauge wire
α_g	= coefficient of linear expansion of the gauge wire
α_b	= coefficient of linear expansion of the base material
b	= total width of wire in strain gauge
β	= the thermocouple constant
e	= strain
e_1	= applied strain
e_2	= apparent strain
i_g	= current in the gauge
I_G	= current through the galvanometer
k	= strain transmission constant
l_1 l_2	= lead resistances
L	= the total lead resistance
λ	= the strain sensitivity factor
R_b, R_c	= resistances of bridge arms
R_t	= resistance of the gauge wire at temperature, t
R_g	= the active gauge resistance
R_d	= the dummy gauge resistance
σ	= Poisson's ratio for the base material
t	= temperature rise of gauge and thermocouple
τ	= temperature
T	= the dummy gauge and cold junction temperature
V_B	= the bridge voltage
V_g	= voltage applied to the gauge
γ	= temperature constant of proportionality

difficulty in obtaining an effective zero temperature sensitivity, especially at slightly elevated temperatures. It would be necessary to consider in advance the material to which the gauge is to be attached as this would influence both the choice of gauge wires and their ratios in the composite grid. In the text a new method is described in which the change in signal voltage due to temperature variation is offset by the voltage from a thermocouple in the gauge, and it is expected that this will supersede the existing methods in view of the greater accuracy which may be obtained and the ease with which the new system may be applied to a variety of investigations, at fairly high temperatures. With this type of gauge a further advantage is its universal application.

Temperature Compensation by the Thermocouple Method

The strain gauge provides a signal by varying the current in the circuit of which it forms part. If the current is held constant, even though the resistance changes, then no signal is transmitted. A thermocouple can give an output which maintains the current constant even when the gauge resistance changes due to change of temperature, and this has been used to cancel signals due to temperature variation without affecting those due to strain.

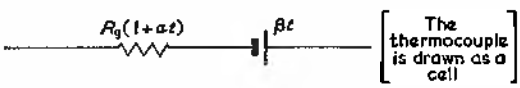


Fig. 1. The gauge and thermocouple junctions

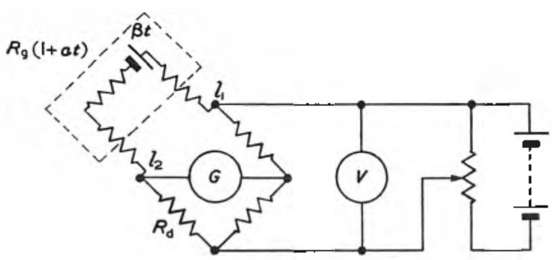


Fig. 2. The bridge circuit with the automatically compensated gauge

If the thermo-electric output of a thermocouple is βt volts, where t is the temperature rise above the cold junction and β the thermo-electric constant applicable to the couple, and the gauge matrix has resistance R_g, with apparent temperature coefficient of resistivity α, the circuit, consisting of gauge and thermocouple junctions can be represented as shown in Fig. 1.

When a voltage V_g is applied to the gauge junctions the current in the gauge is given by:

$$i_g = \frac{V_g + \beta t}{R_g(1+\alpha t)}$$

and if i_g is independent of temperature:

$$di_g/dt = 0$$

i.e. $$\frac{R_g(1+\alpha t)\beta - (V_g+\beta t)R_g\alpha}{R_g^2(1+\alpha t)^2} = 0$$

which requires that:

$$V_g = \beta/\alpha$$

Fulfilment of this condition is a requirement of this method of automatic compensation and, provided that this condition is maintained, temperature variations in the strain gauge are undetectable by the remainder of the Wheatstone bridge circuit.

It will be evident that V_g is the voltage applied to the gauge filament and due allowance must be made for the resistance of the thermocouple leads. Moreover, in practice, it would be more convenient to measure the bridge voltage than the gauge voltage, the correct value being easily ascertained.

The complete bridge circuit containing the automatically compensated gauge would appear as in Fig. 2.

With a lead resistance $l_1 + l_2$ the polarization voltage required for compensation at the active arm becomes:

$$\beta/\alpha\left(1 + \frac{l_1 + l_2}{R_g}\right)$$

and in practice the balancing arm R_d would be arranged to have the same resistance as the active arm:

i.e. $R_d = R_g + l_1 + l_2$

and the bridge voltage would then be

$$V_B = 2\beta/\alpha\left(1 + \frac{l_1 + l_2}{R_g}\right)$$

It is therefore apparent that with a knowledge of the two constants β and α a particular gauge voltage and hence bridge voltage may be applied to the bridge in order that any temperature change in the active arm resistance would not affect the initial balance of the bridge.

The apparent temperature coefficient α would obviously be composed of terms relating to the temperature coefficient of the strain gauge wire, α_r, the coefficient of linear expansion of the gauge wire, α_g, and the coefficient of linear expansion of the base material, α_b. The following expression would then be valid for any combination

$$\alpha = \alpha_r + (\alpha_b - \alpha_g)k\lambda$$

where k is a constant to allow for imperfect transmission of strain from the base to the gauge wire and λ is the strain sensitivity factor for the gauge, whence the resistance of the gauge wire at temperature τ would be

$$R_\tau = R_g + \Delta R$$

(R_g is the gauge wire resistance at room temperature)

or $$R_\tau = R_g[1 + \{\alpha_r + k\lambda(\alpha_b - \alpha_g)\}t]$$

As a result of previous experimental investigations it has been established that the strain sensitivity factors of the materials commonly used in strain gauge manufacture are not subject to change at elevated temperatures, at least up to 400°C. A practical gauge strain sensitivity factor for use in measuring bridge constants may be determined from the constant for a single wire and the dimensions of the strain gauge manufactured from this wire,

i.e. $$\lambda_g = \lambda_w\left(\frac{a - \sigma b}{a + b}\right)$$

where λ_g = Gauge factor
λ_w = Strain sensitivity for a single wire
a = Total length of wire in strain gauge
b = Total width of wire in strain gauge
σ = Poisson's ratio for the base material, this factor being applicable to uni-directional stress.

For accurate strain readings it is obviously an advantage to arrange for low cross-sensitivity. However, pure temperature strains affect both the length and width of the gauge and therefore λ_w would be applicable in the expression for R_r. The value of k has in practice been found to approximate to unity at room temperature and since it is directly related to the strain gauge adhesive shear modulus it is important that adhesives with no marked change in this property at elevated temperatures are selected and a very thin layer is applied under each gauge. The Araldite range, particularly types 15, F and Strain Gauge Cement, and Cellobond J.1531 Resin, display this characteristic, although with some limitations. Refractory adhesives such as 'Brimor' cement and alumina sprayed bond have been shown to be superior in this respect.

The resistance increase

$$R_g\{\alpha_r + k\lambda(\alpha_b - \alpha_g)\}t$$

is balanced effectively by the thermocouple output βt and therefore annulled

i.e. $\quad \beta t = i_g R_g \ \alpha t$

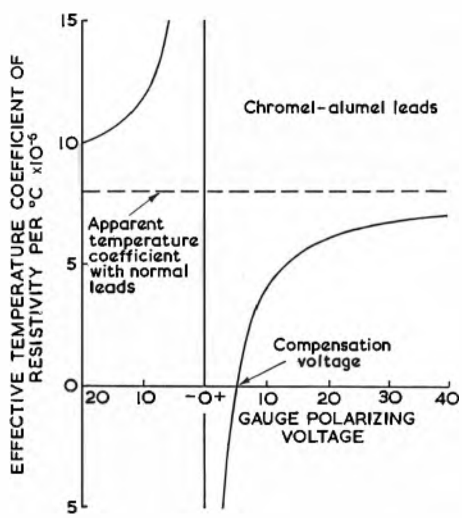


Fig. 3. The relation between the effective temperature coefficient of resistivity and the bridge polarizing voltage for a typical self-compensating strain gauge

In practice the use of thermocouple leads means that the effective temperature coefficient at the active gauge arm is controllable directly by the polarizing voltage, and can be adjusted as desired. This is illustrated in Fig. 3, the curves having been obtained experimentally for an Evanohm gauge.

The expression for the effective coefficient is

$$\alpha_{eff} = \frac{V_g\,\alpha - \beta}{V_g + \beta t} \simeq \alpha - \beta/V_g$$

This expression assumes negligible lead resistance, but a similar equation may be obtained if this is not so, and would form the basis of the theoretical equation of the graph shown.

At zero effective temperature coefficient the slope of the curve for say Nichrome wire would be greater and the corresponding voltage less. Small inaccuracies in voltage setting would then be more critical. One volt for compensation would not be unreasonable for Nichrome.

It is important that the lead resistance should be kept as low as possible by having no unnecessary length of thermocouple connexion. If the true fractional resistance change measured due to applied strain in the surface to which the gauge is affixed is

$$\frac{dR_g}{R_g + L} = \lambda e_2$$

i.e. $\quad dR_g = \lambda e_2(R_g + L)$

it should be $\quad dR_g = R_g \lambda e_1$

where R_g = the strain gauge resistance

L = the total lead resistance

e_1 = applied strain

e_2 = apparent strain

then $e_1 = e_2\,(1 + L/R_g)$ gives the relation between the true and apparent strain which at temperature will be

$$e_1 = e_2'\left\{1 + \frac{L + (l_1\alpha_1 + l_2\alpha_2)t}{R_g}\right\}$$

where α_1 and α_2 are temperature coefficients of resistance of the lead materials.

The bridge sensitivity should be arranged to be of a high order for measurements with the special gauges by the use of a high resistance galvanometer.

The current through the galvanometer in an unbalanced bridge is given by

$$I_G = \frac{E\,(R_2R_4 - R_1R_3)}{R_2(R_1+R_4)(R_G+R_3+R_4)+R_1R_3R_4-R_3R_4^2+R_GR_3(R_1+R_4)}$$

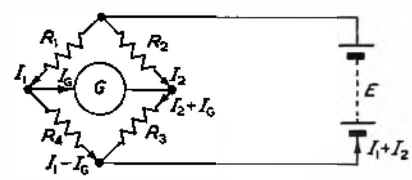


Fig. 4. Nomenclature for bridge system calculations

and the maximum sensitivity when R_1 changes would be

$$dI_G/dR_1 = 0$$

i.e. $\quad R_G = \dfrac{A + B}{2} \qquad$ if $\quad R_1 = R_4 = A$ and $R_2 = R_3 = B$

which is common practice.

For high sensitivity the resistance of the gauge must, it will be seen, be comparable to that of the galvanometer. However, in order to minimize the effect of lead resistance the value of R must be kept high, preferably about 2kΩ. For general purposes the galvanometer resistance should be quite high, although a model with resistance of about 450Ω would give a reasonable sensitivity. The bridge shown in Fig. 2 was designed for use with gauges of the order of 2kΩ, the galvanometer resistance being of comparable value.

The Development of the Thermocouple Method for Temperature Compensation

The Choice of Materials for the Couple

Research on this project initially concerned the selection of suitable thermocouple pairs and strain-gauge wires, the essential requirements being a linear relationship of temperature to e.m.f. and a constant apparent temperature coefficient of resistivity. The expansion coefficients for the common gauge wires and most base materials which are likely to be subjected to examination are constant over the temperature range in which the automatic system can at present be used. While the iron-Constantan and copper-Eureka couples gave a suitable output they do not comply with the linearity condition in the higher temperature ranges and hence only a very limited use was envisaged at this stage. The copper-Eureka couple possessed an additional advantage in being relatively insensitive to temperature changes resistively. However, a chromel-alumel couple, in addition to possessing a constant output with temperature up to at least 1 000°C was found to possess a convenient thermo-electric constant. Having satisfied these requirements, the majority of research was done employing the chromel-alumel couple, which although originally not expected to be used at very high temperatures, owing to other limitations, left some scope for future development.

The temperature coefficients of the strain gauge wires in common use were determined empirically, and it was found that each type of wire possessed inherent limitations to the temperature to which compensation could be maintained. The onset of non-linearity, followed by an inversion, occurred at about 100°C for Eureka wires, at 120°C for Karma wires, and about 430°C for Nichrome.

The bridge polarizing voltage required to compensate this gauge, V_B, over the range 20°C to 430°C is given by

$$V_B = 2\beta/\alpha\ (1+k)$$

where k is a constant equal to the ratio of lead resistance to gauge resistance when thermocouple leads are used.

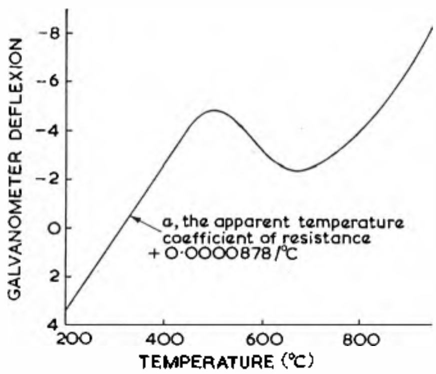


Fig. 5. Bridge unbalance due to temperature with a Nichrome wire resistance strain gauge affixed to a Nimonic base with an aluminium oxide bond

From the experimental results obtained it was noticed that the slope of the resistance/temperature curve in the case of Nichrome wires had approximately the same value from 20°C to 430°C as from 800°C to 1 000°C, and this fact formed the basis of an attempt to obtain compensation in the higher temperature range by the insertion of a known resistance in the active arm, following bridge balancing at 20°C. The results were, however, anomalous possibly due to the inaccuracy with which the required resistance change could be determined. Provided the appropriate adhesive had been employed for the temperature range required it was found that any non-linearity in the resistance/temperature curve of the strain gauge wires was not a function of the adhesive or base material.

At this stage the temperature limitation of the strain gauge wires was accepted, although a search for new resistance materials which would operate in the compensation system is still being conducted. Evanohm wire has been found to have excellent characteristics at least up to 200°C, the limit of the test temperatures with this material to date.

It is suggested that adhesives in the Araldite range and also Cellobond J.1531 Resin commonly used with the resin impregnated matrix strain gauges should not be used where temperature in excess of 200°C are maintained for long periods. It should be mentioned, however, that at temperatures in excess of 100°C Cellobond Resin or a refractory adhesive would be preferred due to the tendency of adhesives in the Araldite range to permit a loss in strain retentivity upon heating from ambient to a temperature above 100°C. The resultant lower strain recording would be unaffected by subsequent cooling although removal of applied strain would result in a residual strain reading of opposite sign. The extent of the errors produced would depend largely upon the thickness of adhesive layer under the strain gauge and the change in mechanical properties of the adhesive within the critical temperature range, and might with experience be resolved. This phenomenon, which has been applied to the stress freezing process of three-dimensional photoelastic model manufacture, is obviously undesirable in the temperature-compensated strain gauge technique.

Refractory adhesives such as sprayed alumina or 'Brimor' cement have been found very satisfactory for higher temperature research. In these cases it is necessary to use a grid or foil type strain gauge, specially manufactured and applied to the material surface by a transfer technique. A typical Nichrome wire strain gauge bonded by the sprayed alumina process to a Nimonic 100 test beam using a Nichrome pre-coat was arranged to compensate satisfactorily up to 430°C although above this temperature the gauge displayed normal performance characteristics even at 1 000°C.

Determination of the Polarization Voltage for Compensation

The determination of polarization voltage on the bridge has presented considerable difficulty, mainly due to its unpredictability for any given combination of gauge, adhesive, base material and thermocouple connexions. This has been found entirely due to the inconsistency of the real temperature coefficient values of the strain gauge wires.

From the practical standpoint the evaluation of V_g for particular combinations has consisted of bridge balancing with thermocouple junctions in a constant temperature atmosphere, initially the same temperature as the measuring equipment and active strain gauge. Subsequent heating of the active gauge produced strain drift measured by galvanometer deflexion for given bridge voltage values. Bridge voltage/effective temperature coefficient curves have been plotted for each type of strain gauge combination and with each themocouple polarity. An example is shown on the graph in Fig. 3, which is typical, and it will be observed that the curves for each polarity are asymptotic to the apparent temperature coefficient of the gauge with ordinary non-compensating leads. One branch of the curve passes across the zero temperature coefficient line and it is at the voltage corresponding to this that compensation is fully obtained, reverse polarity being indicated.

Once the general trend has been empirically established it was only necessary to adjust a voltage control to give zero drift upon attaining an equilibrium temperature condition. The time taken to reach a temperature equilibrium depended largely on the type of specimen under experimental investigation. This procedure of balancing at an elevated temperature by an alteration in bridge voltage is only valid provided the initial resistive balance has been made with zero e.m.f. in the thermocouple normally. In the event of difficulty being experienced in attaining the same temperatures of the gauge and the constant temperature junction, that is with an e.m.f. in the thermocouple it is possible to obtain temperature compensation by careful attention to the balance.

The effect of the initial e.m.f. will be shown as a galvanometer deflexion with zero polarizing voltage. If this deflexion is maintained as a new zero upon the arbitrary polarization of the bridge, by making resistive adjustments, then further temperature changes of the strain gauge will be annulled as in normal compensation upon establishing the compensation voltage setting experimentally. The balance deflexion should be obtained by replacing the supply battery with a resistance equal to its internal resistance.

It will be appreciated that the material to which the temperature-compensated gauge is attached must be completely stress free at the time of the initial balance. Stress change upon heating during the compensation voltage determination will adversely affect the accuracy.

The Form of the Temperature-Compensated Strain Gauge

For the original experimental work the standard strain gauges, particularly the resin-impregnated helically-wound variety, have been found to give a limited degree of compensation. In these cases the thermocouple leads have been welded or soldered on to the nickel attachment strips of the strain gauge, although in order to produce a practical compensating gauge these standard gauges were superseded by specially manufactured ones. The standard gauges were found to produce small spurious e.m.f's, generated no doubt as a result of temperature gradients within these comparatively lengthy intermediate metals formed by the lead tags.

By the use of the special gauges in the ultimate stage of development only one intermediate metal is present, namely the strain-gauge wire. In this case short lengths of flattened thermocouple attachment strips are welded on to the ends of the strain gauge helix or grid, as the case may be, and following the attachment process to the test specimen additional lengths are welded in place. In order to eliminate completely the intermediate metal some research has been done with thermocouple wires wound in strain gauge forms, so that one side lead and the gauge would be of one metal and the other side lead of another, forming a thermocouple pair. Unfortunately the standard thermocouple wires generally have a slightly lower resistivity than the normal strain gauge materials. It was therefore decided not to proceed any further with this particular aspect as it was not essential to normal accuracy requirements, and in any case the special gauges already referred to were found to be satisfactory even when the temperature equilibrium is only approximate.

Performance of Experimental Gauges

Although it was not anticipated that the inclusion of the thermocouple system in a bridge circuit would affect in any way the measurement of applied strain if the lead lengths were kept short, an examination of the performance of both the special and standard gauges has been made at elevated temperatures. A four-point loading, constant-bending-moment rig enabled the measured strain to be compared with the applied strain. A muffle furnace was placed round the loading beams during test. The true strain sensitivity factor, that is making due allowance for the presence of adhesive, was found not to alter appreciably with the introduction of the new system, either at room temperature or at elevated temperatures.

It was evident that an error source would be present due to the heating of that portion of the thermocouple leads near the gauge under test. The magnitude of this has been investigated in detail and in addition two other sources of error in the automatic temperature compensation system, namely the variation in the temperature of the unheated thermocouple junction and the setting of the bridge polarizing voltage. At first it was thought that these three errors presented a very serious problem, but by careful attention to detail the total errors may be reduced to an acceptable or in some cases a negligible level for the majority of strain-gauge applications.

The total strain error can be represented in the form

$$\delta e = \{ (\partial e/\partial \alpha'')\, \delta\alpha'' + (\partial e/\partial V_B)\, \delta V_B + (\partial e/\partial T)\, \delta T \}$$

comprising the three main sources of error, variations in α, V and T,

$$\text{but} \qquad e = 1/\lambda \ \ [\alpha_{eff}\, t + (\alpha'' - \alpha')T] + e_1$$

$$\text{and} \qquad \alpha_{eff} = (\alpha'' - 2\beta/V_B)$$

$$\therefore \delta e = 1/\lambda \{ (t + T)\, \delta\alpha'' + (2\beta/V^2{}_B)\, t\delta V_B + (\alpha'' - \alpha')\, \delta T \}$$

$$\text{or} \qquad \delta e = \delta e_1 + \delta e_2 + \delta e_3$$

Variations in α'' may arise due to variations in lead resistance not being proportional to gauge temperature. The lead environment will determine the extent of this.

The significance of these components is shown as follows

$$\delta e_1 = 1/\lambda \ \ td\alpha'' = 1/\lambda \left\{ \frac{dl_1 + dl_2}{l_1 + l_2 + R_g} \right\}$$

is a term due to thermocouple lead heating

and

$$\delta e_2 = 1/\lambda \left\{ \frac{2\beta \ \ tdV_B}{V^2{}_B} \right\}$$

is a term due to bridge voltage setting

and

$$\delta e_3 = 1/\lambda \ \ (\alpha'' - \alpha') \ \ dT = 1/\lambda \ \ (2\beta/V_B - \alpha') \ \ dT$$

is a term due to variation in temperature of the unheated thermocouple junction.

An analysis of the errors to be expected with different combinations of thermocouple, gauge resistance, lead length and gauge wire material has been made with a view to obtaining the most suitable combinations for particular applications and at the same time minimizing or eliminating the errors by selection where possible. For the majority of applications it has been found convenient to use a room temperature junction in the form of a lagged vessel, provided the duration of test is not more than a few hours. For greater durations it would be necessary to have a temperature-controlled junction either below or above ambient temperature, provided the initial bridge balance has been obtained with the active gauge thermocouple junction at the same temperature as the 'cold' junction. Balancing can otherwise be achieved with due consideration of the galvanometer zero shift.

It was seen that the 'cold' junction need not necessarily be a significant source of error, and this was also found to be the case with the voltage supply. Provided it is possible to obtain the polarizing voltage value which will give compensation by means of a preliminary experiment, or even as part of an actual test by making adjustments, this need not be the cause of undue concern, but it would be necessary to ascertain that no applied strains were present at the time. The reason for this is primarily because each gauge has its own particular compensation voltage due to the inconsistency of temperature coefficients of gauges manufactured from some types of material.

For some wire gauge materials, particularly Nichrome, a prediction may be made with confidence. Karma, and to a lesser extent Evanohm, posses unpredictable temperature coefficients, these being rather sensitive to pre-heat treatments and it would, therefore, be considered essential to obtain a compensation voltage experimentally in these cases, for accuracy. Subjection of the strain gauge system to elevated temperatures will, as explained, enable the compensating polarization voltage to be obtained, although it would be emphasized here that the voltage required may increase as much as 20 per cent to an optimum value with temperature cycling, particularly if temperatures in excess of 300°C are involved. In practice, however, it has been found that Nichrome particularly displays this tendency and that about four cycles will be all that is necessary to obtain optimum voltage value.

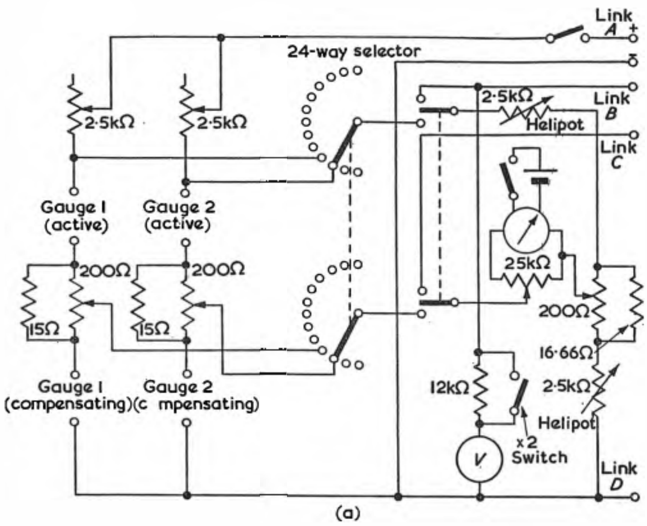
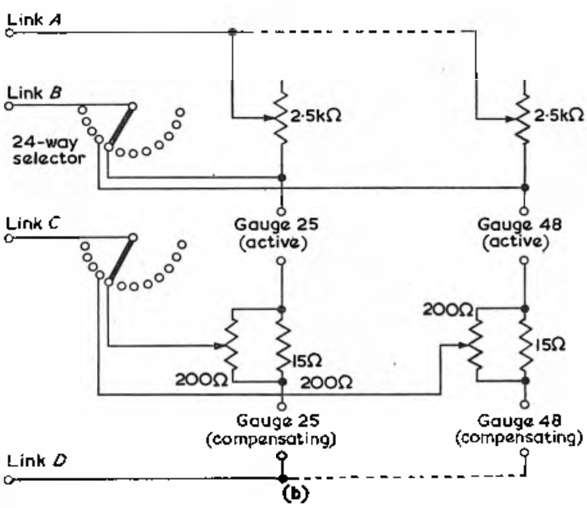


Fig. 6 (a) Strain gauge bridge for temperature compensated strain gauges



(b) Apex unit for temperature compensation strain gauge bridge

Once the compensating polarization voltage has been established for any given combination, an applied stress investigation may be made by adjusting the polarizing bridge voltage to that value, if need be, prior to each strain measurement, and thus eliminating this source of error.

Limitations in the use of a conventional strain gauge multi-channel bridge due to the inconsistency in the polarizing voltage required by individual gauges for compensation has led to the design of a special bridge with 24-channel input. A circuit diagram is shown in Fig. 6, together with a 24-channel apex unit which may be directly coupled.

The bridge and apex are suitable for field stress investigations involving a large number of active gauges, owing to their compactness.

A number of repeatability tests have been made with samples of strain gauges made from one batch of wire, and the only material examined to date which gives consistent temperature coefficients of samples within that batch is Nichrome, and on this basis may be used in stress research at temperature without preliminary experimental work, although a small error may result.

The greatest inherent error to be expected with the compensation system will be seen to be due to the thermocouple wire heating as indicated previously. The effect would be recorded as an apparent strain. In order to minimize the error it is necessary to select a thermocouple pair with low temperature coefficient of resistivity, but still, nevertheless giving the required output, over the temperature range of the test. A chromel-alumel pair partly fulfils this requirement. One of the greatest single factors in reducing this source of error would be to obtain the thermocouple wires in rather a large diameter, an 0·032in diameter wire having been used very conveniently welded directly on to the small section attachment strips of the gauge. Continuation wires of very much larger dimensions may be used to connect to the instrumentation should this be far away.

Compensation voltage determined experimentally will to a certain extent take into account the temperature resistive changes of the thermocouple leads, but as ex-

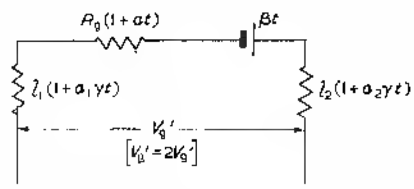


Fig. 7. The active arm of the bridge

plained earlier the changes need not necessarily be proportional to gauge temperature. In practice therefore the error magnitude due to this source may be considerably less than calculation would indicate. The active arm of the bridge may be represented as shown in Fig. 7 if it may reasonably be assumed that the average temperature change of the thermocouple wires is proportional to the active gauge temperature. It is evident that with this assumption, together with a knowledge of the practical bridge compensation voltage the error component δe, which is significant in the case of chromel-alumel couples, would be negligible. The practical bridge voltage, determined by experiment following temperature cycling, would of course have a value given by

$$V_B' = \frac{2\beta\,(R_g + l_1 + l_2)}{R_g\alpha + (l_1\alpha_1 + l_2\alpha_2)\gamma}$$

where α_1 = the temperature coefficient of one thermocouple material

α_2 = the temperature coefficient of the other thermocouple material

γ = temperature constant of proportionality

The compensation system has been examined in conjunction with the N.A.C.A. special circuit for the elimination of errors due to small resistive changes in the strain-gauge leads, or contact resistances in the case of slip rings. A circuit of the type shown in Fig. 8 has been used where the resistive change in the thermocouple wires due to testing has been of the order of one ohm. The N.A.C.A. Technical Note No. 1031 deals fully with the system for single gauges, a number of gauges and gauges of different nominal resistances. The temperature compensation research has so far been confined to one active gauge

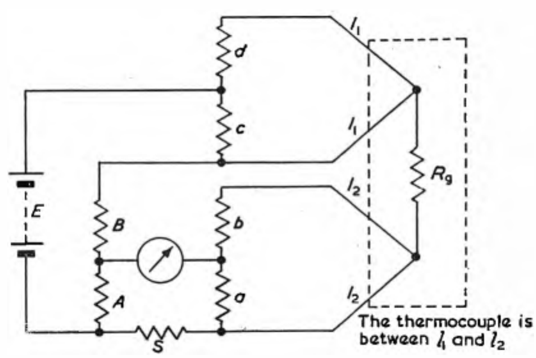


Fig. 8. The elimination of errors due to change in lead resistance

mainly and in one case to two active gauges, although the latter number of gauges has not been tried in conjunction with the N.A.C.A. circuits.

With the circuit of Fig. 8 it is necessary to balance as follows:

$$A/B = a/b \qquad A/S = c/d$$

and the values of R, the active strain gauge resistance, of S, the dummy gauge resistance in the main ratio arms may be for example between 100Ω and 1kΩ. The resistances of the auxiliary ratio arms would be chosen according to the value of A, usually a fixed resistance in the measuring equipment.

l_1 and l_2 are the thermocouple leads to the active strain gauge, and the resistance of these would not normally change more than 1Ω even if a wire of very small section had been in use with a change of 200°C. It would only be necessary to use this circuit in the unusual case when it is not possible to determine the polarizing voltage experimentally, and leads of significant resistance are employed.

Practical Temperature-Compensated Strain Gauges

Characteristics of Gauges of Various Types of Construction

Research has, for reasons already outlined, been performed mainly using chromel-alumel couples, the presence of which appears to have no effect on the effective strain sensitivity of the strain gauge in use. These have been combined with Nichrome, Karma or Evanohm gauges.

Nichrome wire strain gauges have operated successfully up to 420°C with compensation at 0·75 to 1·5V, depending on the base material in use. This type of gauge is particularly recommended for investigation in the higher temperature ranges where Karma and Evanohm would not operate effectively, although it has been found that Nichrome gauges particularly need temperature cycling to obtain an accurate compensation voltage. The bridge sensitivity is rather poor, however, with this gauge material due to the comparatively low polarizing voltage necessary. Above 200°C refractory adhesives were naturally employed.

For more accurate work below 200°C the Evanohm strain gauges have performed excellently. A compensation voltage of about 3·0 using a Dural base gives a very good bridge sensitivity but the actual value required would have to be obtained by a preliminary experiment due to the inconsistent temperature coefficient, as explained earlier. With this type of strain gauge it is possible to read applied strains down to 300lb/in² with a negligible drift during quite rapid heating of the order of 40°C/min. It is possible that the Evanohm gauge would operate at temperatures greater than 200°C and further research will be done to extend the information already available.

The Karma wire gauges will compensate effectively

TABLE 1

Theoretical Temperature Compensation Errors due to the Combined Effect of Power Supply Variation, Change in Temperature of Cold Junction and Change in Lead Resistance caused by Heating

$$\delta e = \delta e_1 + \delta e_2 + \delta e_3$$

Type of thermocouple	Strain gauge cemented to a steel base. Average temperature of thermocouple 100°C above balancing temperature. Cold junction temperature variation 1°C.				Type of strain gauge	
	(A) 21 gauge wire 10ft lengths. Voltage set to 1 per cent accuracy		(B) 21 gauge wire 10ft lengths. Voltage set to 0·1 per cent accuracy			
	Strain δe(%)	Stress (lb/in.²)	Strain δe(%)	Stress (lb/in.²)	Resistance (Ω)	Material
Chromel-Alumel	0·0166 0·1366	4980 40980	0·0162 0·1362	4860 40860	2400 200	Karma
	0·0205 0·1405	6150 42150	0·0169 0·1369	5070 41070	2400 200	Nichrome
	0·0172 0·1370	5160 41100	0·0163 0·1363	4890 40890	2400 200	Eureka
Iron-Constantan	0·0101 0·0759	3030 22800	0·0070 0·0728	2100 21800	2400 200	Nichrome
Copper-Constantan*	0·0055 0·0161	1650 4840	0·0019 0·0125	570 3750	2400 200	Nichrome
Silvo Palladium-Copper	0·0058 0·0196	1725 5880	0·0022 0·0160	645 4800	2400 200	Nichrome
N.A.C.A. Circuit used to eliminate lead resistance effects						
Chromel-Alumel	0·0045 0·0045	1350 1350	0·0009 0·0009	360 360	2400 200	Nichrome

The error values have been based on the most common apparent temperature coefficients for each strain gauge wire material. Evanohm wire would be expected to give errors of the same order as Eureka.

* *This couple is non-linear and would be subject to additional errors*

with a temperature limitation at 100°C but have the advantage in that a high bridge sensitivity results from the polarizing voltage for compensation of between 10 and 38V. The considerable variation in required voltage is brought about by the sensitivity to heat treatment, subjection to temperatures over about 150°C appearing to alter the temperature coefficient permanently.

Examples of the errors to be expected with the compensated gauge due to

(a) Power supply
(b) Cold junction
(c) Lead resistance.

are given in Table 1. It will be realized that the values quoted must be treated as maxima.

Eureka gauges have been included in this table purely for comparison purposes, although their use would not be recommended above 100°C as the temperature coefficient is not constant above this figure. A compensating polarization voltage in the region of 10V would be expected with this type of gauge. An additional couple, namely Silvo Palladium-Copper has been included for interest and certainly suggests that some further research is required with this material. The errors in the table are totals and the combination of the three sources with the conditions stated below.

(a) 21 s.w.g. thermocouple wire with strain gauge attached to steel test piece. A temperature rise average of 100°C in the thermocouple wire and the bridge voltage set to 1 per cent accuracy. The error would, of course, be reduced should the voltage be adjusted in the manner outlined in the text, although the difference would not be appreciable except in the case of Nichrome wire gauges. The thermocouple wire is of 10ft length.
(b) 21 s.w.g. thermocouple wire with strain gauge attached to a steel test piece. A temperature rise of 100°C average in the wire which is of 10ft length. The voltage supply is set to 0·1 per cent accuracy.

Systems involving Evahohm gauges have not been included separately but, depending on the temperature coefficients, would have inherent errors similar to those for Karma wire. It will be noticed that for each type of gauge one particular temperature coefficient value has been assumed, but this is, however, a typical value for the samples already examined in the research.

Lastly the errors to be expected on the assumption that the N.A.C.A. circuit would eliminate the lead effects entirely, which has been shown empirically in at least one case, are shown. With the N.A.C.A. circuit in effect the total errors are very small and would be considered negligible for the majority of applications.

The total errors shown in the table are intended to represent typical examples. They could be reduced by reducing the lead temperature rise and lead length and should be used as a guide in a practical application.

The temperature compensation system may be expected to operate at sub-atmospheric temperatures although no definite research has been conducted to confirm this. However, it has been established that the standard strain gauges exhibit normal characteristics at temperatures down to −40°C with the range of adhesives normally used in the research programme on the compensated strain gauge.

Characteristics of the Gauges Selected for Use

In addition to tests made on simple specimens the new system has been used satisfactorily in practical investigations in which the bridge method of compensation would have introduced appreciable error. The following describes the performance which has been experienced, but particular characteristics could no doubt be further developed for different applications.

Compensation has been regarded as satisfactory when the variation in apparent stress due to temperature variation is less than 500lb/in^2 (on steel). The error due to temperature variation did not exceed this value during a test in which a small assembly was allowed to cool from 200°C at a rate of 40°C/min initially, although in this case the gauges were lagged with an additional coating of adhesive. The error when the normal bridge compensated method was used in the same example was 3 500lb/in^2, even though the temperature was very closely controlled. Had it not been, the error would have been many times greater.

Compensated gauges have been used satisfactorily at temperatures up to 420°C. Below 200°C a gauge of normal construction wound with Evanohm would be used; for higher temperatures a specially constructed gauge using Nichrome.

The stress analyst would inevitably be presented with the problem of measuring strains on rotating members at unknown temperatures and in order to determine the efficacy of the system operating in conjunction with slip rings a rig has been built in which gauges may be strained to known magnitudes by the centrifugal effect on a rotating shaft. Heating may be applied by a muffle furnace and a constant temperature junction is provided to terminate the gauge thermocouple leads.

Conclusions

The type of strain gauge which has been developed enables stress to be measured when the temperature is changing quite rapidly, and when the temperature is not known. This has enabled stress investigations to be made on an assembly cooling after a fitting operation, for instance, and of stress due to load with temperature changing, which could not have been usefully attempted with the usual bridge compensated systems.

Compensation close enough to give a stress reading error due to temperature variation of 500lb/in^2 or less on steel has been arranged at temperatures up to 420°C, and with rates of change of temperature, on cooling from 200°C, of at least 40°C/min.

It is anticipated that the ramifications of the automatic temperature compensated strain gauge technique will be considerably greater than at first envisaged and the authors are by no means complacent with regard to the extent of the experimental verification. It is hoped in the near future to complete the research on the use of the special gauges on rotating members in order that steady stress investigation may be conducted on, for example, turbine rotors. The range of suitable high-temperature materials investigated for use in the strain gauge system will be extended concurrently.

Consideration will be given to circuit modifications, such as direct application of thermocouple e.m.f.'s to the galvanometer, in order to improve overall sensitivity, particularly in the case where materials of high α are used.

Acknowledgments

The authors wish to thank Mr. L. F. Hall, the Chief Research Engineer of D. Napier & Son Ltd. for permission to publish this work and acknowledge the encouragement given by Mr. S. H. Jenkins, the Assistant Manager, Research, and the assistance given by Mr. R. W. Nevin in the experimental investigations.

A Negative Capacitance Pre-Amplifier for Electrophysiological Use

By B. M. Johnstone* and I. D. Pugsley†

A positive feedback circuit, giving negative input capacitance is analysed. The limitations of frequency response are discussed. A transistorized amplifier has been developed to test the theory and indicate other practical limitations.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 466)

ELECTROPHYSIOLOGICAL investigations involving the use of micro-electrodes require an amplifier with frequency response up to 100kc/s while the source impedance may be up to 100MΩ. To give the amplifier a high input resistance with low shunt capacitance, a cathode-follower is normally used, with a guard-ring connected to its cathode. Since an electrometer valve is essential, and since these generally have a low amplification factor, the gain of the cathode-follower may not be high enough, and the guard-ring may not give sufficient reduction of capacitance.

Various methods of overcoming this difficulty by the use of feedback have been described. Ito[1] has described a method of feeding back to the low impedance side of the source, and Solms *et al.*[2] have described a method of feeding back directly to the input grid. The amplifier described here uses feedback via the grid-anode capacitance of the input valve, in order to compensate for stray capacitance and give an effective input capacitance less than zero if required.

Circuit

The complete circuit is shown in Fig. 1 and summarized in the block diagram, Fig. 2. The cathode-follower consists of an ME1400 electrometer with an npn transistor as the cathode load. The transistor is an adjustable constant current generator of high impedance. The ME1400 is operated at 'contact potential' (zero grid current).

The d.c. amplifier consists of one OC45 transistor for impedance transformation, two OC45's as a long-tailed pair and one OC71 as an output transistor enabling the output quiescent voltage to be zero.

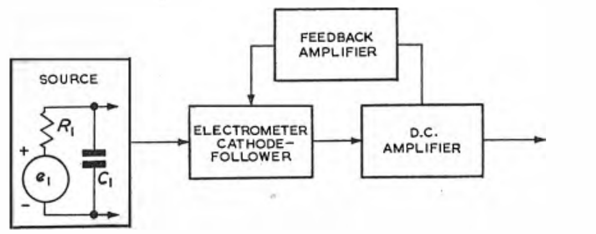


Fig. 2. Block diagram

The analysis does not involve the d.c. amplifier except for the first emitter follower in Fig. 1

The feedback amplifier uses a pair of 2N384 drift transistors for good frequency response. They are operated as a long-tailed pair.

The metering circuit serves three purposes:

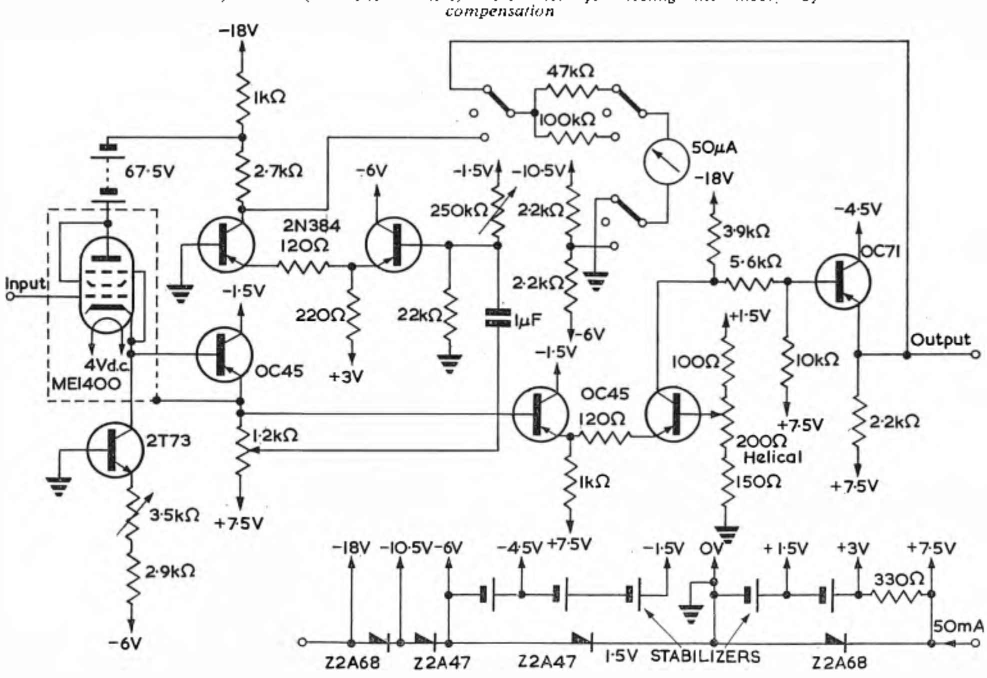
(a) Balance of feedback pair

(b) Balance of d.c. amplifier pair

(c) Monitoring d.c. potentials. In particular, intracellular resting potentials can be measured easily and directly.

Fig. 1. Complete circuit

Modifications (not shown here) were used for testing the theory of compensation



The power supply uses Zener diodes and nickel-cadmium stabilizers. External requirements of 50mA at 25V may be derived from a battery or regulated mains power supply.

Analysis

The block diagram (Fig. 2) may be drawn in the form of an equivalent circuit (Fig. 3). C_4 is the grid to screen-plus-anode capacitance R_3 and C_3 represent a limitation to high frequency

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response. They occur in the cathode-follower, which is drawn as a voltage generator feeding into the R_3-C_3 network. The input impedance of the feedback amplifier (gain A) is assumed to be infinite, and its output impedance is shown as R_4. The gain of the feedback amplifier A is variable, and adjusts the degree of compensation.

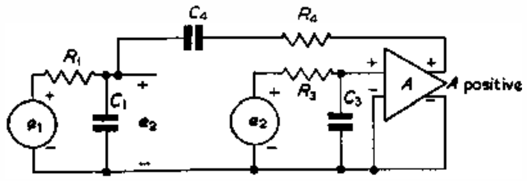


Fig. 3. Primary equivalent circuit

Nominal values are: R_1—10MΩ R_4 1kΩ
C_1—10pF C_4 1pF
e_1—100mV RC_3 1μsec

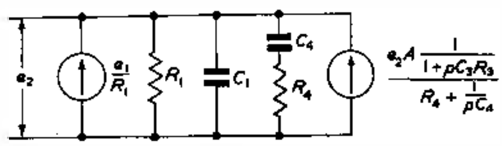


Fig. 4. Secondary equivalent circuit

The coupling capacitor in the circuit diagram (Fig. 1) is not important for the purposes of analysis, because it will only affect low frequencies. The effect is shown in Fig. 6, and will be explained below. The output is taken at the input to amplifier A on Fig. 3 and is equal to

$$\frac{e_2}{1 + pC_3R_3}$$

The circuit may be simplified by using Norton's theorem to that of Fig. 4. Adding current generators and admittances and applying Ohm's law:

$$e_2 = \left((e_1/R_1) + \frac{e_2A\dfrac{1}{1+pC_3R_3}}{R_4 + (1/pC_4)}\right) \times \frac{1}{pC_1+(1/R_1)+1/[R_4+(1/pC_4)]}$$

$$= \left((e_1/R_1) + \frac{e_2ApC_4}{(1+pC_3R_3)(1+pC_4R_4)}\right) \times \frac{R_1(1+pC_4R_4)}{(1+pC_1R_1)(1+pC_4R_4)+pC_4R_1}$$

$$= \frac{e_1(1+pC_3R_3)(1+pC_4R_4)+e_2ApC_4R_1}{(1+pC_3R_3)} \times \frac{1}{(1+pC_1R_1)(1+pC_4R_4)+pC_4R_1}.$$

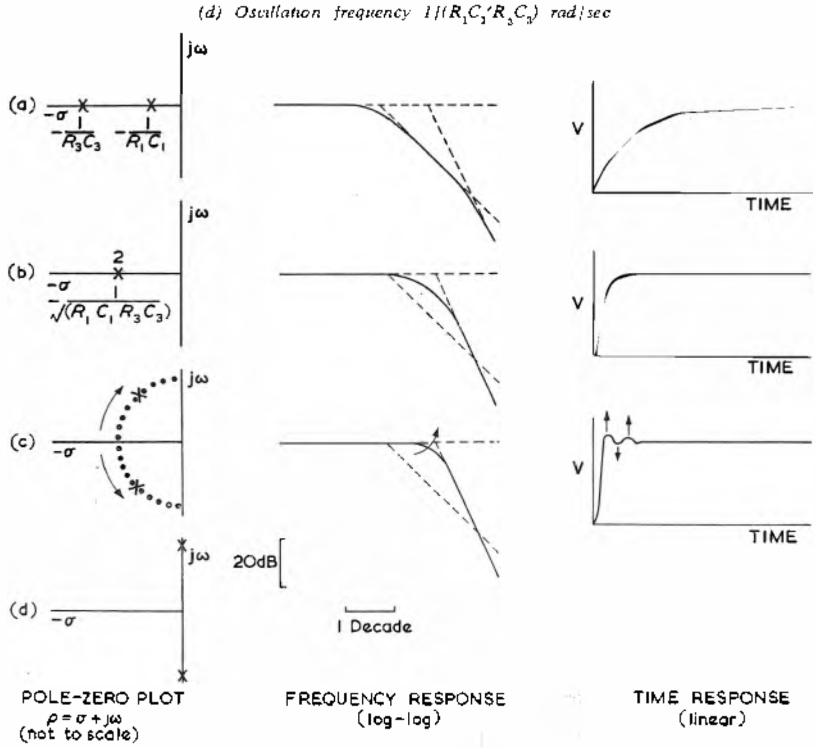
Therefore

$$e_2 = \frac{e_1(1+pC_3R_3)(1+pC_4R_4)}{(1+pC_3R_3)(1+pC_4R_4)(1+pC_1R_1)+pC_4R_1(1+pC_3R_3-A)}.$$

By making the simplification that C_4R_4 is small, and neglecting it (which is permissible, since C_4R_4 is less than 1/1 000 of any other time-constant), the output voltage is equal to:

Fig. 5. Comparison of various degrees of compensation

(a) No compensation
Bandwidth $1/R_1C_1'$ *rad/sec*
Rise-time $2{\cdot}2\ R_1C_1'$ *sec*
(b) Optimum compensation
Bandwidth $0{\cdot}6444/(R_1C_1'R_3C_3)$ *rad/sec*
Rise-time $1{\cdot}5\ (R_1C_1'R_3C_3)$ *sec*
(c) Overcompensation
Radius of circle $1/(R_1C_1'R_3C_3)$ *rad/sec*
(d) Oscillation frequency $1/(R_1C_1'R_3C_3)$ *rad/sec*



$$\frac{e_2}{1+pC_3R_3} = \frac{e_1}{(1+pC_3R_3)(1+pC_1R_1)+pC_4R_1(1+pC_3R_3-A)}$$

$$= \frac{e_1}{(1+pC_3R_3)(1+pC_1R_1) - pC_4R_1A}$$

where $C_1' = C_1 + C_4 \simeq C_1$.

The simplification made above, that C_4R_4 is small, is necessary to avoid the solution of a cubic equation, and it is thus only necessary to solve a quadratic equation in order to find the poles and thus the frequency response.

The transfer function $e_2/e_1\ (1+pC_3R_3)$ has poles p as the roots of the equation:

$$(1+pC_3R_3)(1+pC_1'R_1) = pC_4R_1A$$

These are sometimes real (for example $A = 0$), and sometimes complex conjugate (for example):

$$A = \frac{C_1'R_1 + C_3R_3 - \sqrt{C_1'R_1C_3R_3}}{C_4R_1}$$

For a transfer function which will give no overshoot for a step function input, the poles must both be real. The limiting case here is when they are both equal and:

$$p = -\sqrt{(1/C'_1R_1C_3R_3)}$$

In this case the gain is:

$$A = \frac{C_1'R_1 - 2\sqrt{(C_1'R_1C_3R_3)} + C_3R_3}{C_4R_1}$$

$$= \left[\sqrt{(C_1'/C_4)} - \sqrt{\left(\frac{C_3R_3}{C_4R_1}\right)}\right]^2$$

This is the case of optimum compensation. Fig. 5 shows the position of the poles for different amounts of feedback.

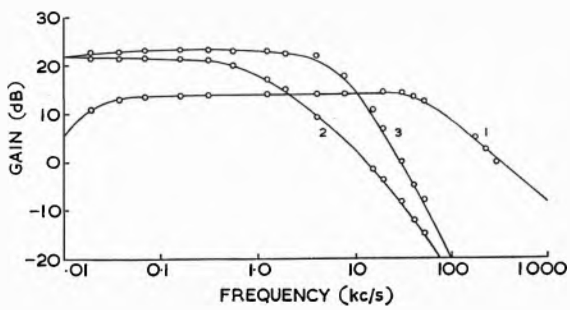


Fig. 6. Response curves of amplifier with frequency response deliberately limited as described under 'Experimental Tests'

(1) Feedback loop response

(2) Main amplifier limited by R_1C_1'

(3) Compensated response curve of main amplifier. Points were obtained experimentally, and curves drawn in accordance with theoretical roll-off shape

With no feedback, the frequency response is limited mainly by R_1C_1 and a normal single time-constant response is obtained. When feedback is applied, the two poles move along the real axis towards each other. They meet at the midpoint (on a logarithmic scale) and when they do, the compensation is considered optimum. There is no overshoot. If more feedback is applied, the poles move on a circle towards the imaginary axis. Ringing increases until the circuit oscillates.

Experimental Tests

In order to make the tests conclusive, steps were taken in order to:

(a) Reduce direct capacitive feed-through across R.

(b) Deliberately limit the feedback amplifier response by increasing the cathode-earth capacitance. This makes the amplifier limited by a single time-constant R_3C_3.

Fig. 6 shows:

(1) The frequency response of the feedback amplifier after limitation.

(2) The frequency response of the main amplifier with a source impedance R_1C_1'.

(3) The frequency response of the main amplifier with the effect of R_1C_1 compensated for by the use of the feedback amplifier. It will be noticed from the curve drawn for optimum compensation that the points indicate slight overcompensation. (See frequency responses, Fig. 5.)

There is another positive feedback effect over and above the feedback through C_4, namely, from anode to cathode. This is present even with the grid earthed. The amount of feedback is determined by the ratio of r_{pa} to R_K, where r_{pa} is the anode resistance of the valve and R_K is the output impedance of the cathode-follower. The effect is to increase the gain of the cathode-follower at all except low frequencies when it is limited by the coupling capacitor (see Fig. 6). This is the only manifestation of the presence of the coupling capacitor.

The measured values of cut-off frequency, compared with theoretical values are:

	Measured	*Theoretical*
(1) Feedback amplifier	70kc/s	(R_3C_3)
(2) No compensation	1kc/s	(R_1C_1')
(3) Optimum compensation	7kc/s	5kc/s

The frequency of oscillation cannot be determined accurately by experiment because of non-linearities in the amplifier, causing clipping and frequency shift, but when the circuit was on the verge of oscillation, it would appear to be about 10kc/s, theory giving 9kc/s.

Conclusion

The general principle that positive feedback decreases bandwidth is applicable here. Therefore, in order to obtain a ten-fold improvement in rise-time, the feedback amplifier must have a rise-time a hundred times as good as the source because its bandwidth will be effectively 'decreased' ten times. The compensated cut-off frequency is 0·64 times the geometric mean of uncompensated cut-off and feedback amplifier cut-off.

General Performance of Amplifier

The following values were obtained without artificial limitations imposed as for testing the theory.

Frequency Response

(1) Source impedance zero:	rise-time 2μsec bandwidth 150kc/s overshoot 5 per cent
(2) Source impedance 50MΩ	(e.g. typical micro-electrode) rise-time 400μsec bandwidth 900c/s no overshoot
(3) As for case (2) with compensation:	rise-time 80μsec bandwidth 4·5kc/s no overshoot

Input impedance

Approximately 10[illegible]Ω at contact potential.

Gain 10.

Noise 100μV peak-to-peak.

Drift 10mV/h, mainly due to filament accumulator.

Operation

No faults have developed during six months, or 500 hours of normal operation. The control of feedback is very smooth and setting up procedure is very simple.

Acknowledgments

The authors wish to thank Dr. D. J. Dewhurst for the original ideas and Dr. Mollie E. Holman and Mr. L. Dally for assistance and encouragement.

REFERENCES

1. Ito, M. The Electrical Activity of Spinal Ganglia Cells investigated with intracellular microelectrodes. *Jap. J. Physiol.* 7 297 (1957).
2. Solms, S. J., Nastuk, W. L., Alexander, J. T. Development of a high fidelity preamplifier for use in the recording of bioelectric potentials with intracellular electrodes. *Rev. Sci. Instrum.* 20 170 (1953).
3. Haapanen, L., Ottoson, D. A frequency compensated input unit for recording with microelectrodes. *Acta Physiol. Scand.* 32 271 (1954).

Limitations in Squaring with Simple Valve Circuits of Wide Bandwidth

By R. M. Huey*, B.Sc., B.E., and J. E. Longfoot†, B.E.

The errors to be expected from squarers using both curvature of a valve transfer characteristic and anti-phased inputs to separate control electrodes, are compared. The former are obtained from a theoretical analysis and the latter from measurement.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 466)

THE applications of square law devices are numerous, the main use being for the accurate measurement of power in complex waveforms. Over a very wide frequency range, it is necessary to have methods of power measurement other than normal peak or average reading devices which can be r.m.s.-calibrated for a specified waveform only. Noise investigations in particular require square law measuring devices, as many noise types can be adequately characterized by their power frequency spectrum. The need thus arises for a simple square law circuit having reasonable accuracy and a wide bandwidth.

Two methods of squaring using valve characteristics are discussed in detail in this article:

(1) By making use of a non-linear transfer characteristic with input applied to one electrode.

(2) By applying input to two electrodes of a suitable multi-grid valve.

The performance of a squarer of the latter type is detailed.

Squaring with Non-Linear Valve Characteristics

The anode current of a triode valve can be written as a finite power series in the grid voltage,

$$i_a = \alpha + \beta e_g + \gamma e_g^2 + \delta e_g^3 + \epsilon e_g^4 + \zeta e_g^5 + \eta e_g^6 + \dots \quad (1)$$

where α, β, etc., are positive real constants. For squaring, only the e_g^2 term is desired. The constants defining the power series depend on

(1) Operating point P on the characteristic.

(2) Dynamic range.

(3) Number of terms chosen to form the power series.

For a reasonably large squared term, the valve should be biased to operate on the knee of the curve (as in Fig. 1) and this immediately places some restriction on the dynamic range.

The greatest difficulty with this type of squarer lies in extracting the squared term from the undesired terms. There are two common methods of achieving this, either by using a matched pair of valves with the two grids fed with equal and opposite voltages, or by using a frequency selective circuit in the anode of the valve.

(a) EXTRACTION OF THE SQUARED TERM BY USING MATCHED TRIODES

Two triodes are used with a common anode load, and the two grids are fed with the same signal but with 180° phase shift between them as shown in Fig. 2.

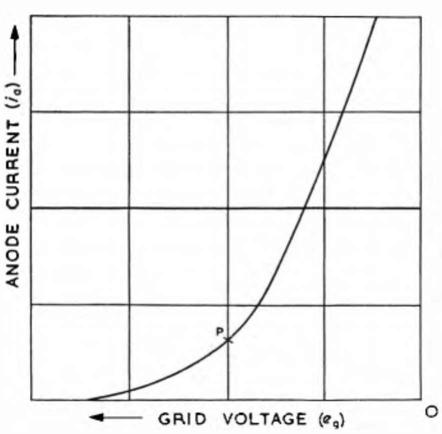


Fig. 1 (left). Valve transfer characteristic

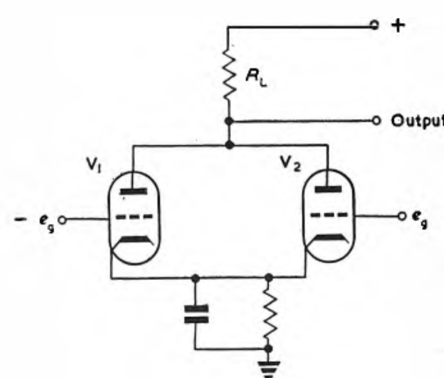


Fig. 2 (right). Matched triode squarer

The anode currents of the two valves will be:

$$i_a = \alpha_1 + \beta_1 e_g + \gamma_1 e_g^2 + \delta_1 e_g^3 + \epsilon_1 e_g^4 + \dots$$

and

$$i_b = \alpha_2 - \beta_2 e_g + \gamma_2 e_g^2 - \delta_2 e_g^3 + \epsilon_2 e_g^4 + \dots$$

Now if the valves are correctly matched,

$$\beta_1 = \beta_2$$
$$\delta_1 = \delta_2 \text{ and so on.}$$

The output voltage V_o will be given by

$$V_o/R_L = i_a + i_b = (\alpha_1 + \alpha_2) + (\gamma_1 + \gamma_2) e_g^2 + (\epsilon_1 + \epsilon_2) e_g^4 \dots \quad (2)$$

If, in the original equation (1) the e_g^4 coefficient and subsequent even power coefficients are small compared to that of the e_g^2 term, then the a.c. component of the output voltage will be predominantly the input voltage (more properly its a.c. component) squared.

Properties

The correctness of squaring is completely dependent on the correct matching (i.e. initial selection) of the valves, and there is no easy control to enable re-balancing the circuit as the valves age.

In general the grid characteristics of pentodes do not show such a marked curvature as triodes, and so triodes

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are to be preferred despite the consequent limitation of high frequency response (due to Miller effect). The circuit of Fig. 2 cannot be direct coupled unless extremely high voltage levels are used, this restriction being caused by the large drift. The desired output signal will be small, as the e_g^2 term is necessarily smaller than the e_g term in equation (1), while the large drift is caused by the drifts in the two valves being additive; the nature of the circuit prevents adequate drift compensation being used.

The disadvantages listed above are considerable.

(b) EXTRACTION OF THE SQUARED TERM BY USING A FREQUENCY SELECTIVE CIRCUIT

This type of squarer can be made useful over only a narrow frequency range. A useful approximate analysis can be made by considering a single frequency signal. Putting $e_g = \cos\theta$, in equation (1)

$$i_a = \alpha + \beta \cos\theta + \gamma \cos^2\theta + \delta \cos^3\theta + \epsilon \cos^4\theta + \zeta \cos^5\theta + \eta \cos^6\theta + \; \ldots .(3)$$

$$\text{Now } \cos^2\theta = \tfrac{1}{2}(\cos 2\theta + 1)$$
$$\cos^3\theta = \tfrac{1}{4}\cos 3\theta + \tfrac{3}{4}\cos\theta$$
$$\cos^4\theta = \tfrac{1}{8}\cos 4\theta + \tfrac{1}{2}\cos 2\theta + \tfrac{3}{8}$$
$$\cos^5\theta = \tfrac{1}{16}\cos 5\theta + \tfrac{5}{16}\cos 3\theta + \tfrac{5}{8}\cos\theta$$
$$\cos^6\theta = \tfrac{1}{32}\cos 6\theta + \tfrac{3}{16}\cos 4\theta + \tfrac{15}{32}\cos 2\theta + \tfrac{5}{16}$$

From this can be seen that a band-pass filter may be arranged to pass the second harmonic term (2θ), and to reject the fundamental (θ) and third harmonic (3θ). As the coefficients ϵ and η may be significant the fourth and sixth power terms contribute to the 2θ output, causing an error.

The above analysis applies for a simple sinusoidal voltage.

Consider now an input of two sinusoids of differing frequency

$$e_g = A \sin\omega t + B \sin(pt+\phi)$$
$$e_g^2 = A^2\sin^2\omega t + B^2\sin^2(pt+\phi) + 2AB\sin\omega t\sin(pt+\phi)$$
$$= \tfrac{1}{2}A^2(1-\cos 2\omega t) + \tfrac{1}{2}B^2\{1-\cos(2pt+2\phi)\}$$
$$+ AB\cos(\omega t - pt - \phi) - AB\cos(\omega t + pt + \phi) \; \ldots . \; (4)$$

yielding desired terms in $2\omega t$, $2pt$, $\{(\omega - p)t - \phi\}$, $\{(\omega + p)t + \phi\}$, and at d.c.

Now to obtain reasonable squaring, the periods of ωt and pt must be close enough together to ensure that the frequencies, $2\omega t$, $2pt$, and $\{(\omega+p)t+\phi\}$ are contained within the pass-band of the filter. It is obvious that an error must be incurred if the $\{(\omega-p)t-\phi\}$ term cannot be contained in the pass-band. In special cases it might be possible to provide separate filters in the output circuit, one a band-pass, to transmit the $2\omega t$, $2pt$, and $\{(\omega+p)t+\phi\}$ terms, and the other a low-pass to give the $\{(\omega-p)t-\phi\}$ and d.c. terms while still rejecting the ωt and pt terms

The error caused by the d.c. components will be ignored as they have the same coefficients as the double frequency terms. The error caused by neglecting $\cos\{(\omega-p)t-\phi\}$ may be expressed as follows:

$$\text{Fractional Error} = \frac{AB\cos\{(\omega-p)t-\phi\}}{\tfrac{1}{2}A^2 - \tfrac{1}{2}A^2\cos 2\omega t + \tfrac{1}{2}B^2 - \tfrac{1}{2}B^2\cos(2pt+2\phi) - AB\cos(\omega t + pt + \phi) + AB\cos(\omega t - pt - \phi)}$$

It is feasible to define a state in which there is always a predominant signal $A\sin\omega t$, with another smaller signal represented by $B\sin(pt+\phi)$.

Let $(A/B) = r$.

$$\text{Fractional Error} = \frac{r\cos\{(\omega-p)t-\phi\}}{\tfrac{1}{2}r^2 - \tfrac{1}{2}r^2\cos 2\omega t + \tfrac{1}{2} - \tfrac{1}{2}\cos(2pt+2\phi) - r\cos\{(\omega+p)t+\phi\} + r\cos\{(\omega-p)t-\phi\}}$$

The instantaneous value above is not a useful measure of the error. It is convenient to evaluate an r.m.s. fractional error Δ (see Table 1).

$$\Delta = \sqrt{\left\{\frac{\tfrac{1}{2}r^2}{\tfrac{1}{4}(r^2+1)^2 + \tfrac{1}{4}\cdot\tfrac{1}{2}r^4 + \tfrac{1}{4}\cdot\tfrac{1}{2} + (r^2/2) + (r^2/2)}\right\}}$$
$$= \sqrt{\left\{\frac{4r^2}{3r^4 + 12r^2 + 3}\right\}}$$

TABLE 1. Error due to rejection of difference frequency term

r	1	2	3	5	10	30
Δ (per cent)	47	40	32	21	11	3

A signal consisting of a fundamental plus harmonics will incur a greater error than a signal whose component frequencies are close. Consider a signal $A\sin\omega t$, with second harmonic $B\sin(2\omega t+\phi)$, then from equation (4).

$$e_g^2 = \tfrac{1}{2}A^2(1-\cos 2\omega t) + \tfrac{1}{2}B^2\{1-\cos(4\omega t+2\phi)\}$$
$$+ AB\cos(\omega t+\phi) - AB\cos(3\omega t+\phi).$$

It can be seen that a filter tuned to $2\omega t$ will reject both the ωt, $3\omega t$ and $4\omega t$ terms, causing a large error. Similar considerations apply for a third harmonic in the input waveform.

Properties

The main drawback is the inaccuracy of squaring all waveforms except a practically pure sinusoid, unless rather special frequency responses are arranged for the output. Using a simple band-pass filter, all the sinusoids present must be close together in frequency, and for reasonable accuracy, one sinusoidal frequency must be substantially larger in amplitude than the others (see Table 1). This device is hence severely frequency restricted. The gain will be about half that in case (a), with the same order of small dynamic range.

The circuit is sometimes used for measuring the power in amplitude modulated signals and can be used as a class-A frequency multiplier with a simple tuned circuit as the filter.

Squaring with Multi-Grid Valves

This system depends on the effect of two grids on the electron stream in a multi-grid valve such as a pentode or convertor valve.

An analysis of the operation of such a device has been given by H. N. Coates[1] and may be summarized as under.

Let x = maximum peak cathode current

p, q = functions relating anode current to bias on grids 1 and 3

a, b = variations in p and q due to signals applied to grids 1 and 3

Then $I_a = (q \pm b)(p \pm a)$

Now the same input signal is applied to grid 1 and

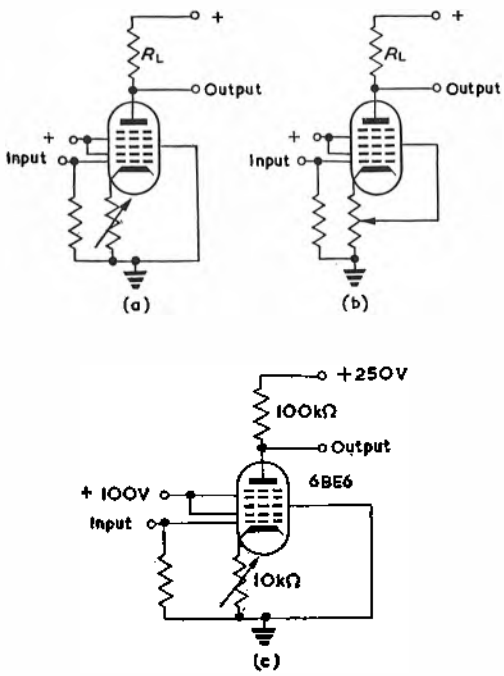


Fig. 3. Multigrid valve squarer
(a) Equalizing by variable cathode load resistor, (b) Equalizing by variable feedback ratio, (c) Circuit values for type (a)

grid 3 but in opposite phase, so

$$I_a = x\,(qp - bp + aq - ab) \qquad \ldots\ldots\ldots (5)$$

and provided that a and b are restricted to swing over a linear part of the transfer characteristics, the desired squared term ab may be extracted by adjusting the bias on the cathode until $+\,aq$ is just cancelled by $-bp$ while pq is a fixed quantity once the bias has been chosen.

In a normal pentode the effect of the control grid is far greater than that of the suppressor so b must be amplified and a reduced. As there is a limit to the amount b can be amplified before encountering cut-off, a must be made very small. This arrangement is most unsatisfactory as the squared term will have a very poor signal-to-noise ratio. There are a few pentodes with larger suppressor grid transconductance, Coates suggesting the Mazda type 6F33. However, convertor valves seem to offer a more effective solution. S. Suzuki[2] has described a rather specialized detector of this type using inputs to two grids of a single type 6SA7, arranged to measure narrow band noise at 10·7Mc/s. However, his scheme depended on an output band-pass filter tuned to double frequency in order to eliminate undesired terms.

A normal convertor valve has oscillator and control grids with anode transconductances of the same order. The circuits of Fig. 3 were found satisfactory using a type 6BE6 valve.

Examination of Fig. 3 (a) and (b) shows that the two grids are effectively fed by signals 180° out of phase (i.e. with respect to cathode), and the variable resistors in the cathode circuit change the bias and or signal voltage to the grids. The effect of the circuit, when correctly equalized, is to make $bp = aq$ in equation (5) and hence

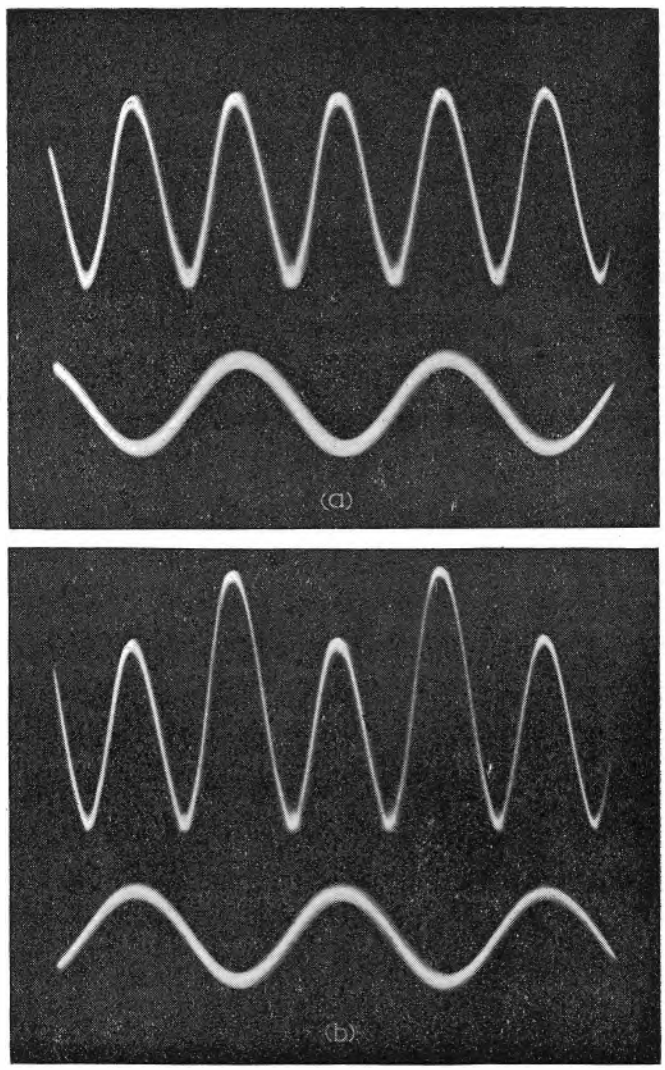
$$I_a = x(qp - ab) \qquad \ldots\ldots\ldots\ldots\ldots (6)$$

A very simple squarer was constructed using a type 6BE6 convertor valve as in Fig. 3(c). It was found that this circuit could be adjusted readily to give a fundamental rejection of better than 40dB for a sinusoidal input signal of fixed amplitude. Input and output waveforms are shown in Fig. 4(a). The difficulty with the circuit was its inability to operate at more than one input voltage level. As soon as the input level was changed (either up or down), the grid to cathode voltage changed causing a consequent change in instantaneous 'bias', and as the gain of each grid is dependent on its grid to cathode instantaneous voltage, the cancellation of unwanted terms is no longer valid. Under conditions of poor cancellation the output voltage from a sinusoidal input has the form shown in Fig. 4(b).

To minimize this effect two valves were used with a common anode resistor, and the grids of each valve fed with a balanced signal. This arrangement is similar to that used in section (a) describing the triode squarer, however, matching is less critical. To avoid the difficulty of matching valves, a balanced circuit was constructed capable of using any two 6BE6 valves, with a common anode load (4kΩ) and individual cathode loads (10kΩ).

The results obtained from this circuit were disappointing, the improvement in the dynamic range (output), being only a few times. However by setting each valve for the correct cancellation of the fundamental terms at opposite ends of the desired input level range, it was possible to obtain improved results (see Fig. 5). With a dynamic range (output) of 2 to 1 a fundamental rejection of 46dB was obtained.

Fig. 4. Response of squarer (Fig. 3(c))
(a) Correctly equalized, (b) Operated outside dynamic range destroying equalization



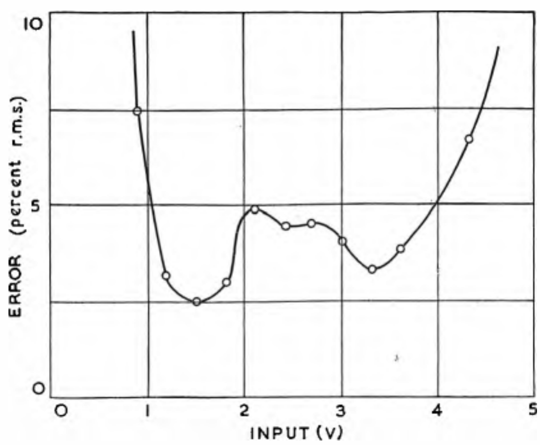


Fig. 5. Ratio of undesired output frequencies (r.m.s. addition) to desired second harmonic (r.m.s. value) as a function of input amplitude (unmatched valves)

Properties

To obtain wide dynamic ranges of good accuracy matched convertor valves should be used, and although the matching need not be as good as for triodes, used as in section (a), it must still be of a high standard. A fairly small dynamic range was achieved with the constructed circuit when no attempt was made to match the valves. This circuit had a much higher gain than either of the triode circuits as the squared term in the output represents a higher proportion of the total output products. This made possible a reduction in the value of the anode load resistor, and a consequent increase in bandwidth, while still retaining a reasonable gain. The amplitude of the output, and the accuracy of squaring were kept constant up to a frequency of 500kc/s, by balancing the stray capacitances, while the low frequency half power point was a few cycles per second. Direct coupling was not feasible, as the total drift of the two valves was high, and any attempt to compensate would upset the squaring action.

This circuit should be capable of an accuracy of 1 per cent with careful matching of the valves. Using unmatched valves, an accuracy of 5 per cent was obtained while still preserving a useful dynamic range (about 16 to 1 in output; see Fig. 5).

An outstanding feature of this type of circuit is its ability to operate as a sine wave frequency doubler without a tuned circuit.

REFERENCES

1. COATES, H. N. A Simple Square Law Circuit with High Frequency Response *Electronic Engng. 29*, 41, 1957.
2. SUZUKI, S. A New Type of Square Law Detector. *Pub. Astronom. Soc. Japan, 7* 121, 1955.

A Single Transistor Radiation Survey Meter

By M. Birnbaum*

A new type of portable radiation survey meter is described, employing a single transistor which provides the high voltage necessary for the G.M. counter and at the same time integrates the pulses to be measured.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 466)

THE modern, portable radiation survey meter has, in addition to the Geiger counter and the supply batteries, two important circuits (see Fig. 1):

(a) The high voltage source that is necessary to feed the Geiger counter.

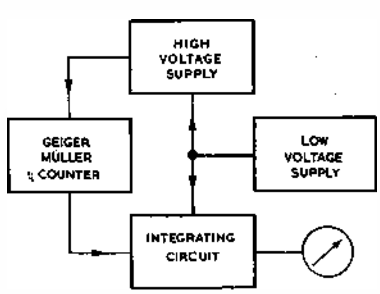


Fig. 1. General arrangement of radiator survey meter

(b) The integrating network for measuring the radioactive doses.

To fulfil these functions two or three transistors are generally used.

In this article a new type of circuit is described, which employs a single transistor both for generating the high voltage and for measuring the counting rate of the pulses given by the counter.

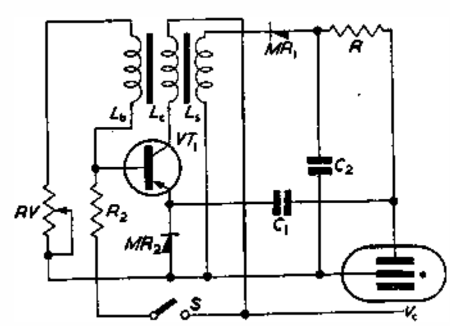


Fig. 2. The meter circuit

The circuit of this meter, shown in Fig. 2, includes a transistor VT_1 in a blocking-oscillator arrangement.

The condition for self-oscillation is fulfilled only if the

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base of the transistor is negative in respect to the emitter, or if a positive voltage pulse appears on the emitter.

When the meter is put into operation, the switch S is closed and the circuit, working as a conventional blocking oscillator provides high voltage pulses which charge the output capacitor through the rectifier diode. The voltage across this capacitor C_2 has the proper value for feeding the Geiger counter.

If the switch S is opened, the transistor works as a blocking generator triggered by the pulses from the Geiger counter, the pattern of the voltage pulse on its collector being shown in Fig. 3.

Apart from the positive voltage pulse (which is nearly rectangular) there also appears a negative voltage pulse, due to the existence of a relatively large inductance in the circuit; this negative pulse can be adjusted to have a peak value several times greater than the battery supply.

The circuit described, utilizes this negative voltage pulse, which appears at each pulse, to feed into the capacitor C_2 a charge so as to maintain the voltage across it nearly constant during the operating period.

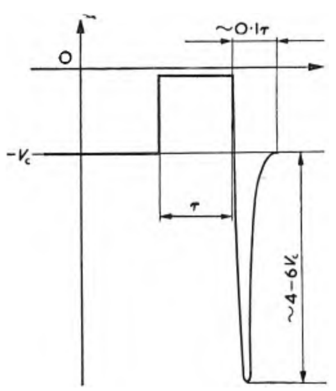

Fig. 3. The Geiger counter waveform

Obviously, there are some difficulties concerning the design of such a circuit:

(a) The output capacitor must have a good quality insulator so as to prevent any leakage. By using a polystyrene capacitor, having a resistance of 10^{12} to $10^{14}\Omega$, a convenient time-constant can be obtained.

(b) The radiation counter must provide on the input resistance of the circuit, voltage pulses of an amplitude greater than that needed to trigger the circuit. For this reason the most suitable is a halogen quenched Geiger counter which, on the other hand, needs a rather low voltage supply (300 to 400V).

The input resistance of this circuit is high; it consists of the emitter resistance of the cut-off transistor (usually 1 to 2MΩ) in parallel with the inverse resistance of the diode MR (being also 1 to 2MΩ).

(c) Care should be taken to ensure that the charge fed into the capacitor C_2 is at least equal to the current through the Geiger counter.

(d) Care should be taken that the negative voltage peak appearing on the collector of the transistor should not exceed the peak voltage admissible for the type of transistor used.

The complete arrangement of the circuit is shown in Fig. 4. The device is provided with a push-button type switch that ensures a correct switching-order; one cannot measure properly unless a certain push-button has been pressed. For instance, to bring the apparatus into operation, one has to press the button labelled 'Start'. In this case, the transistor works as a blocking-oscillator and provides high voltage for the G.M. counter. At the same time the average value of the current measured by the milliammeter may be used to indicate the voltage of the supply batteries.

For actually measuring a radiation dose, one has to press the corresponding button of the switch, enabling the device to measure 1, 3, 10 or 30mr/h at full scale deflexion. By this switching, the base connexion of the transisitor is interrupted, and the transistor stops self-oscillating; at the same time a suitable shunt resistor is connected across the milliammeter to provide the correct sensitivity.

When now a pulse is given by the G.M. counter, it produces a single oscillation of the cut off transistor, and accordingly a collector current pulse.

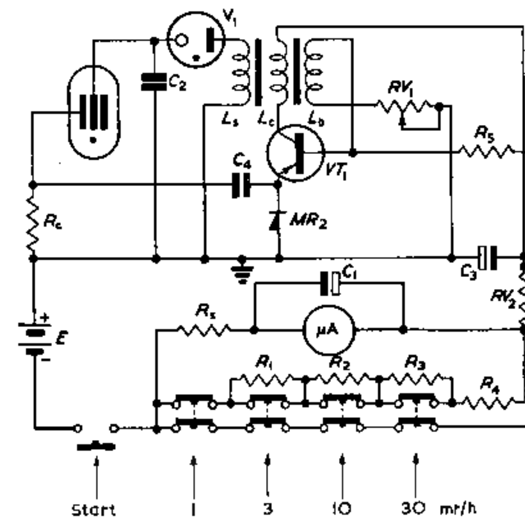

Fig. 4. The complete circuit

These pulses are integrated in the R_xC_1 network and through the shunt resistors R_1 R_2 R_3 R_4 are fed to the milliammeter.

The calibration of the meter can be adjusted with the variable resistor RV by changing the length of the pulse.

By using a Soviet STS-5 Geiger counter giving about 1 200 pulses per minute for 1mr/h, the length of the pulse can be 300 to 500μsec which enables the use of an instrument of 400μA full scale without damaging the transistor VT_1 and without great counting loses.

A CK-1012 cold-cathode type diode has been used to rectify the high voltage pulses. Thus, the meter does not need any power for heaters; the supply delivers energy only if a high radiation dose is being measured.

The meter has been checked a long time to measure the cosmic-ray radiation, and also high radioactive doses. Its errors are smaller than ± 10 per cent.

Conclusions

This meter has certain advantages over the conventional types of radiation survey-meters; it uses only one transistor, and its overall power consumption is extremely low and therefore the life expectance is practically the storage time of the batteries.

Also, it enable the use of a rugged milliammeter while the number of components is small, thus increasing the reliability in operation.

A Graphical Analysis of the Blocking Oscillator

(Part 3)

By D. Chambers*, B.Sc.

(Voir page 326 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 333)

IN this final part, practical aspects are considered and a typical example is worked out to illustrate the proposed method of design.

Output Loading

Output pulses are normally taken across a tertiary winding on the transformer (voltage pulses) or across a resistor in series with the anode or cathode of the valve (current pulses). Transformer coupling allows more flexibility in the choice of output voltage or impedance, and pulses of either polarity are easily obtainable. An overswing voltage is, however, present on the output pulses, and larger trigger pulses are required due to the loading across the transformer. Current pulses, on the other hand, are free from overshoot, and this output arrangement is more suitable for generating pulses across very low impedances (10Ω to 100Ω).

The effect of connecting a load across the transformer is illustrated in Fig. 23 where the proportion of the anode current taken by the load (5·4kΩ) is shown shaded on the paths of operation. (The construction of these paths is described later in the worked example.) It will be evident that as the load current is increased, a greater proportion of the excess current (i_a—i_g) is taken by the load, and the currents available for the shunt capacitance and primary inductance are correspondingly reduced. The equilibrium point is moved to the right and is located where the excess current is equal to the load current (referred to the primary). Loading the transformer therefore increases the rise-time and reduces the amplitude and duration of the voltage pulses.

When the load is connected in series with the anode or cathode of the valve, the equilibrium point occurs at lower values of anode and grid current due to the voltage developed across the load resistor. This is illustrated in Fig. 24 where the shaded paths of operation (for a 1 : 1 turns ratio) apply when there is no series resistor in the circuit, and the unshaded paths show the operation when a 75Ω resistor is connected in series with the anode. The voltage drop across the resistor is added to the transformer voltage drop at each point on the paths. For low values of series resistor (less than 100Ω), the available excess current during the pulse is not seriously reduced, and the rise-time and pulse duration remain almost constant.

Turns Ratio

Although a turns ratio of 1 : 1 between the anode and grid windings is satisfactory for many applications, a more careful choice must be made when optimum operating conditions or a particular pulse shape is required from the circuit. The ratio determines the relative values of voltage and current in the anode and grid circuits during the pulse. The optimum ratio is dependent upon the valve type, the supply voltage and the circuit application. An important factor is the slope of the equilibrium line (joining points of equal anode and grid current) on the valve curves. The steeper the slope, the lower the turns ratio required between the anode and grid windings, and as a general rule, a suitable ratio is the value of $V_a : V_g$ near the operating point on the equilibrium line. For example, a ratio of 1 : 1 is suitable for a 12AU7 and 6J6 while 2 : 1 is better for a 12AT7 or 12BH7.

In some applications, it may be necessary to establish the equilibrium point for more than one value of turns ratio before the optimum ratio can be interpolated. When using high anode supply voltages, the equilibrium point

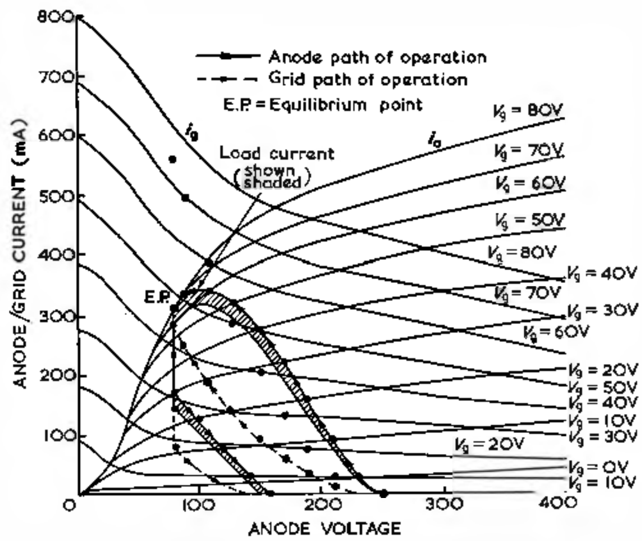

Fig. 23. Positive grid characteristics
Average ½ 12AT7

may occur at excessive values of grid voltage and current, and in these cases, it is better to include a limiting resistor in the grid circuit, than to use a large step-down ratio.

Design Procedure

The following procedure may be employed in the design of most types of blocking oscillator circuit.

(1) Choose the valve type and load arrangement to give the required pulse power into the load.

(2) Determine the turns ratio between the anode, grid and output windings.

(3) Calculate the values of transformer primary inductance and blocking capacitance to give the required pulse width and shape.

(4) Design the transformer.

These factors have been previously discussed, and the method of design will now be illustrated in a worked example.

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Example

A triggered blocking oscillator is required to produce 1μsec rectangular pulses of 20V amplitude and positive or negative polarity into a load of 75Ω. The supply voltage is 250V and the repetition frequency is 1kc/s.

There are two possible ways of producing the pulses across the load; the output can be taken across a tertiary winding, or the load can be connected in series with the anode or cathode circuits of the valve. Both these possibilities will be considered and suitable circuits are shown in Fig. 25.

Output Taken Across Transformer Winding

The output pulse power (5·3W at a duty ratio of 0·001) can be met by using a small general purpose valve, such as a 12AT7, and a suitable circuit is shown in Fig. 25(a). If larger pulses had been required, the same design procedure would apply but a valve with more suitable blocking oscillator characteristics would be chosen.

The turns ratio is not critical in this circuit arrangement and 2 : 1 step-down between anode and grid will be satisfactory.

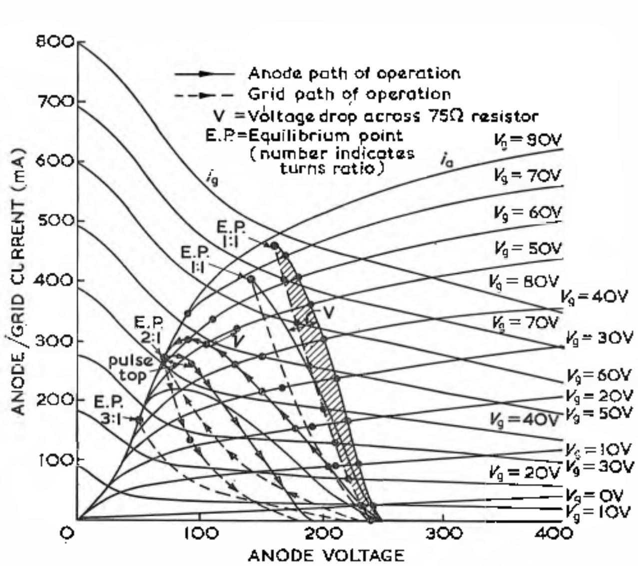


Fig. 24. Positive grid characteristics
Average ½ 12AT7

The anode and grid paths of operation are drawn in Fig. 23, and the proportion of the anode current taken by the load is shown shaded. In constructing the paths, leakage inductance is assumed to be negligible and allowance has been made for the effect of the turns ratio. The anode voltage drop between successive points is therefore twice the corresponding grid voltage rise, and the value of grid current at each point on the grid path is halved. Equilibrium occurs just before the intersection of the anode and grid paths at the point where the anode current is equal to the sum of the grid current and the load current. The operating conditions of the valve at the equilibrium point are $v_a = 80$V, $v_g = 75$V, $i_a = 0{\cdot}31$A, i_g (referred to primary winding) $= 0{\cdot}28$A. The anode voltage drop is 170V so that the turns ratio required to produce a 20V pulse in the output winding is 8·5: 1 and the load current in the primary is then 0·03A. For a flat-topped voltage pulse, the end of the pulse top is on a line drawn perpendicularly downwards from the equilibrium point where the excess current is a maximum. In this case $i_{e(max)} = 0{\cdot}09$A at $v_g = 30$V, and the available magnetizing current is $(0{\cdot}09 - 0{\cdot}03\text{A}) = 0{\cdot}06$A. The voltage across the transformer primary winding is 170V, so, for a 1μsec pulse,

$$L_p = \frac{170 \times 10^{-6}}{0{\cdot}06} \text{ H} = 2{\cdot}8\text{mH}$$

The voltage across C (and also the grid voltage) falls by 45V during the pulse top, and the average value of grid resistance is 161Ω. The required value of C can be found from the expression

$V_t = V[1 - \exp(-t/r_g C)]$ and substituting $V_t = 45$, $V = 75$, $t = 10^{-6}$ and $r_g = 161$, $C = 0{\cdot}0067\mu$F.

In the design of the transformer, a core with high pulse permeability is chosen to give the required pulse inductance with a minimum number of turns. Economic considerations may indicate using the smallest core consistent with an acceptable flux swing. It may be necessary to calculate the number of turns and flux swing for more than one size or type of core before the most suitable one can be selected. Thin Mumetal laminations give the highest values of pulse permeability but saturate rather easily. Ferroxcube is more economical, but additional damping may be required across the windings to prevent re-triggering, and the operating temperature range of this material is limited. Orientated silicon steel cores have lower pulse permeability but can be operated at much higher flux swings and in long pulse applications, the smallest size of transformer is often possible with this core material.

In this example where a 1μsec pulse is required, a suitable core could reasonably consist of a ½in stack of Interservices shape 450 Mumetal laminations, 0·004in thick. The core constants are then, $l = 5{\cdot}06$cm, $A = 0{\cdot}403\text{cm}^2$, $K = 0{\cdot}8$, and average value of μ_p is 750 for a 1μsec pulse. The number of primary turns required to give a pulse inductance of 2·8mH on this core

would be $N = \sqrt{\left(\dfrac{2{\cdot}8 \times 10^{-3} \times 5{\cdot}06 \times 10^9}{4\pi \times 0{\cdot}403 \times 0{\cdot}8 \times 750}\right)} = 68$ turns.

The flux swing is $\dfrac{170 \times 10^{-6} \times 10^8}{68 \times 0{\cdot}8 \times 0{\cdot}403} = 775$ gauss. The corrected number of turns and flux swing would be 70 turns and 755 gauss respectively. (This flux swing is well below the start of core saturation, and it would have been possible to use a smaller core).

Forty gauge enamelled wire would comfortably fill the winding length for the primary, and the grid winding would have 35 turns of 33 gauge. The output winding of 8 turns could be arranged as a pair of 24 gauge wires in one layer to give good coupling. Satisfactory insulation between windings would be obtained using 3 layers of 0·00075in wax impregnated paper, and the effective shunt capacitance between windings would then be approximately 40pF. The leakage inductance of a practical transformer would be between 2μH and 4μH and with this low value of leakage inductance, only the effective shunt capacitance need be considered in determining the time of rise of the pulse. The average excess current during the rise time is 0·11A and the total shunt capacitance, including the valve capacitance, is about 60pF. The approximate time of rise is then

$t_r = \dfrac{60 \times 10^{-12} \times 170}{0{\cdot}11}$ sec $= 0{\cdot}093\mu$sec. Positive or negative pulses are available by simply reversing the connexions to the output winding.

The circuit of Fig. 25(a) was tried, using a blocking capacitor = $0{\cdot}0068\mu F$, a transformer manufactured to the above design, and 20 different valves (5 random samples from 4 different makers). The amplitude of the output pulse, in every case, was between 18V and 22V, the duration of the pulse top between $0{\cdot}75\mu$sec and $1{\cdot}2\mu$sec, the rise-time between $0{\cdot}07\mu$sec and $0{\cdot}10\mu$sec, the fall-time between $0{\cdot}12\mu$sec and $0{\cdot}16\mu$sec and the maximum bowing or slope on the pulse-top was 3V. Fig. 25(b) shows the waveform of a typical output pulse.

low values of anode current are required, it is better to add a series grid resistor than to use a turns ratio greater than 3:1.

During the pulse top, the ideal anode path of operation is a horizontal line beginning at the equilibrium point and moving to the right. (A perfectly flat-topped pulse is not always realizable in practice, and there is often a small droop near the end of the pulse due to the relatively slow start of the trailing edge). As the grid voltage falls, the constant anode current is initially maintained for a small increase in anode voltage. When the grid

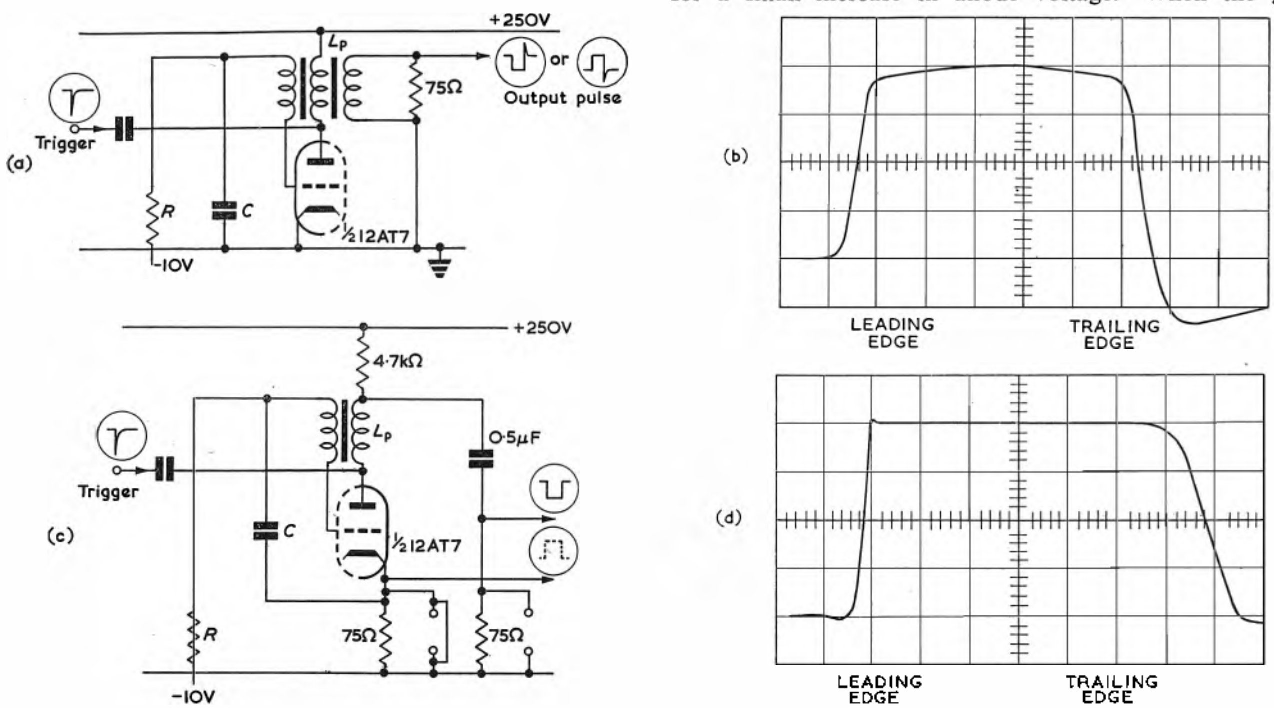


Fig. 25. Circuit to produce a 20V 1μsec pulse into 75Ω load

(a) Output taken across transformer **(b) Output across transformer winding**
(c) Output in series with the anode **(d) Output taken across series resistor**

20V, 1μsec pulse
X axis — 0·2μsec/division
Y axis — 5V/division

Load Connected in Series with the Anode

The same type of valve (12AT7) will be used in the circuit of Fig. 25(c) to enable the two modes of operation to be more easily compared.

The turns ratio in this case is more critical and can be interpolated if the paths of operation are drawn for the ratios 1:1, 2:1 and 3:1 step-down into the grid (Fig. 24). The values of grid current on the paths have been divided by the turns ratio factor, and allowance has also been made for the voltage drop across the 75Ω load resistor in the anode circuit.

When constant current pulses are required, it is advantageous to operate the valve on the diode line (where the anode current curves merge below the knee) since a small change in grid voltage has negligible effect on the anode current. The load current in this example is 0·27A, and occurs on the diode line when the anode voltage is 70V. The turns ratio is chosen so that equilibrium occurs at this point, and in Fig. 24, it will be seen that a turns ratio of 2:1 gives the required operating conditions. If larger or smaller pulses had been specified, a different ratio would have been interpolated from the equilibrium points shown in Fig. 24. For example, 25V pulses would be produced with a ratio of 1·7:1, and a ratio of 2·7:1 would give 15V pulses. When relatively

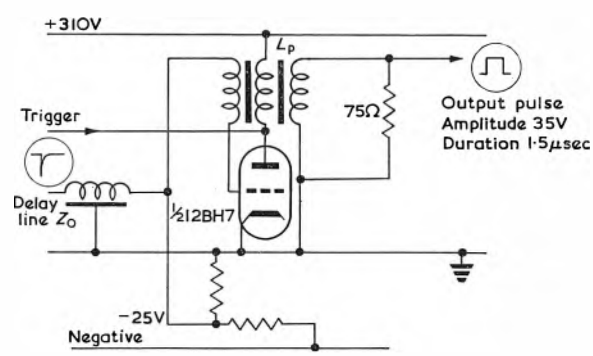


Fig. 26. Delay-line controlled blocking oscillator
When Z_o = 200Ω, L_p = 1·5mH, output turns ratio 5 : 1
When Z_o = 1kΩ, L_p = 2·2mH, output turns ratio 6 : 1

voltage approaches 45V, however, maximum excess current is reached and the anode voltage begins to rise more quickly. The falling grid voltage finally causes less excess current to flow, and the pulse terminates.

In Fig. 24, it will be seen that the grid voltage falls from 70V to 45V during the pulse top, while the anode voltage rises by 22V. The net change in voltage required across the blocking capacitor is therefore 14V, allowing for the 11V fall across the grid winding. The average value of grid current is 0·35A and the value of C to give a 1μsec pulse is $\dfrac{0{\cdot}35 \times 10^{-6}}{14}\text{F} = 0{\cdot}025\mu\text{F}$.

During the pulse top, the anode voltage increases from 70V to 92V and the maximum excess current at the terminating point is 0·13A. Allowing for the voltage drop across the series resistor, the mean voltage across the primary of the transformer is 149V and the required value of L_p is $\dfrac{149 \times 10^{-6}}{0{\cdot}13}\text{H} = 1{\cdot}15\text{mH}$.

The number of primary turns necessary to give this value of inductance on a ¼in stack of Interservices shape 450, 0·004in Mumetal laminations would be approximately

$$N = \sqrt{\left(\frac{1{\cdot}15 \times 10^{-3} \times 5{\cdot}06 \times 10^{9}}{4\pi \times 0{\cdot}403 \times 0{\cdot}8 \times 750}\right)} = 44 \text{ turns, and the flux}$$

swing is $\dfrac{149 \times 10^{-6} \times 10^{8}}{44 \times 0{\cdot}8 \times 0{\cdot}403} \simeq 1\,050$ gauss. The corrected number of turns is 47, giving a flux swing of 980 gauss. Thirty-seven gauge enamelled wire would be suitable for a single layer primary winding, and the grid winding could consist of 23 turns of 28 gauge. The components specified above were used in the circuit of Fig. 25(c) and the following output pulse characteristics were measured for 20 different valve samples: Amplitude 18V to 24V, duration of pulse-top 1·0μsec to 1·4μsec, rise-time 0·04μsec to 0·05μsec, fall-time 0·2μsec to 0·3μsec, bow or slope 0V to 4V. A typical output pulse is shown in Fig. 25(d).

It will be noted that the average pulse-top duration is nearly 20 per cent longer than the designed value (1μsec), and this is probably due to the relatively slow fall-off in current during the first part of the trailing edge, when the excess current remains constant or increases slightly. The rise-time of the pulse is shorter than before due to the greater excess current available to charge the lower shunt capacitance of the two-winding transformer.

Negative pulses are produced by the circuit shown in Fig. 25(c), but positive pulses are obtainable if the shorting link across the cathode resistor is removed, and connected across the resistor in the anode circuit. The blocking capacitor is returned to the cathode, otherwise the pulse amplitude would be too large and the drop in grid current during the pulse would introduce droop on the pulse top.

Other Applications

When the duration of the output pulse is required to be independent of changes in valve and transformer characteristics, delay line control is necessary. A typical circuit is shown in Fig. 26, and the paths of operation of the valve are drawn in Fig. 27 for two values of delay line impedance, 200Ω and 1kΩ. In the construction of the paths it has been assumed that a series resistor, equal to the Z_o of the delay line, replaces the blocking capacitor. A turns ratio of 1:1 is used in this application because of the grid current limiting effect of the delay line. The value of transformer primary inductance is chosen in each case to give a natural pulse duration of approximately twice the terminated duration (double transit time of the line = 1·5μsec).

The condition for successful termination of the pulse is met when the proportion of the reflected pulse in the delay line appearing across the grid forward resistance, is sufficient to cause a decrease in excess current in the valve. For satisfactory circuit operation, the value of delay line impedance should normally lie between 150Ω and 1·5kΩ. Delay lines with values of Z_o lower than 150Ω may not charge sufficiently to cause termination on the reflected pulse, while values greater than 1·5kΩ may cause excessive grid limiting and severely restrict the available output power.

Some applications, such as frequency dividers and television sweep generators, utilize the waveform across the blocking capacitor produced during the charging cycle. Although the circuit is economical for these purposes, its stability is generally inferior to other types of circuit. Variations in valve characteristics, heater voltage and supply voltage have a marked effect on the amplitude of the exponential waveform across the blocking capacitor unless special precautions, such as clamping the negative excursion of the pulse, are taken to improve the stability.

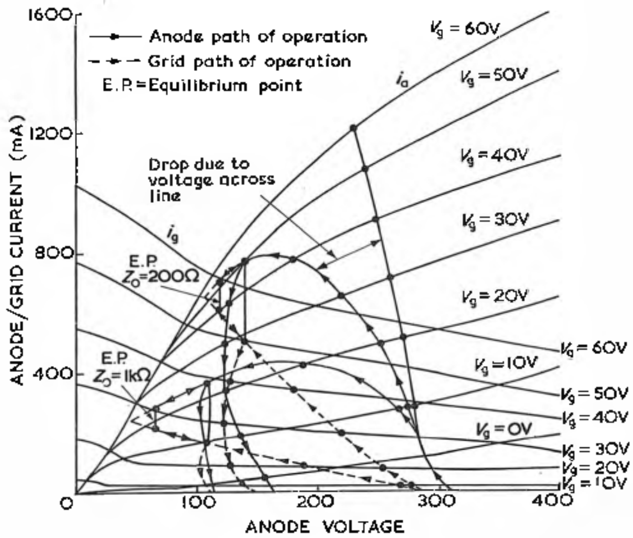


Fig. 27. Positive grid characteristics ½12BH7
Paths of operation for delay line controlled blocking oscillator

If a very large blocking capacitor is used in the circuit, the pulse duration is controlled by the primary inductance of the transformer and the available excess current. The paths of operation during the pulse top retrace the paths during the leading edge, giving relatively high values of excess current. Longer pulses are therefore obtained with this type of circuit, but the pulse shape is not easily controllable.

The design method described in this article is applicable to most blocking oscillator circuits, but there are a few exceptions in which the construction of the paths of operation is not worthwhile. For instance, the values of L_p and C required to generate very short pulses (<0·1μsec) are best found experimentally, bearing in mind the factors that influence the rise- and fall-times (shunt capacitance, leakage inductance and excess current). The transformer primary inductance cannot be determined in the normal manner because the duration of the pulse-top is negligible and the pulse width is usually controlled by the value of blocking capacitor. In these applications, the capacitor charges so rapidly that the normal equilibrium point is never reached. If the value of C is made too small (e.g. less than 50pF) the grid of the valve cannot be driven very positive and the available output power is severely restricted. Maximum coupling and minimum capacitance between windings on the transformer is obtained when a few turns are closely wound on a small closed loop core of Ferroxcube, 0·0006in Mumetal or thin orientated silicon steel.

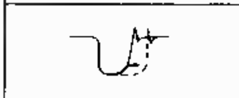
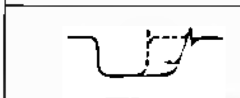
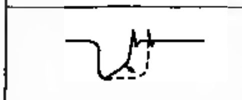
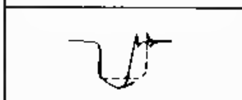
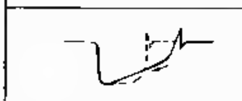
PULSE SHAPE ACROSS TRANSFORMER	CORRECTION REQUIRED
	Increase L_p and C
	Decrease L_p and C
	Increase L_p
	Increase C
	Decrease C
	Decrease L_p

Fig. 28. Correction of pulse shape by changing L_p and C
Full lines show actual shape
Dotted lines show required shape

It may happen that the pulse duration or slope from a blocking oscillator is different from the desired shape and it is useful in these cases to know the necessary corrective measures. If the pulse is too short but approximately the right shape, the values of L_p and C should both be proportionately increased. Similarly, if the voltage pulse has too much droop and is too long, decreasing C will effect a cure. The methods of correcting the duration and slope of voltage pulses are shown in Fig. 28. To correct the shape of current pulses, C and L_p in Fig. 28 should be interchanged. Adjustments to the amplitude of the pulse can be made by changing the turns ratio of the transformer or by varying the value of a resistor in series with the anode, grid or cathode of the valve.

Conclusion

An attempt has been made to give a complete graphical analysis of the operation of the blocking oscillator and to describe a practical procedure for determining component design-centre values. The valve curves and core characteristics shown in this article are representative of a large number of samples measured over a considerable period, but clearly, there can be no guarantee that these characteristics apply to all batches produced by the manufacturers. Special purpose valves, with controlled emission and positive grid characteristics are becoming available, and it is expected that circuit performance with these valves will be far more predictable than is the case with the older valve types.

Acknowledgments

The author would like to thank the Directors of Cossor Radar and Electronics Ltd, for permission to publish this article and for the facilities made available during the investigation.

Television Standards Conversion

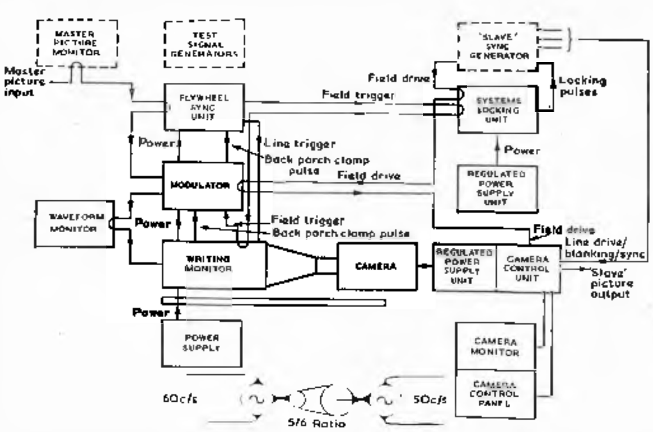
The international exchange of programmes by link or magnetic recordings frequently introduces a need to change standards. The monitor assembly of the Marconi type BD 679 recording channel and the 4½in image orthicon camera type BD 863 together make an ideal combination for this conversion.

By operating the 4½in image orthicon at a point on its characteristic 'below the knee', the storage characteristic is fully exploited to deal with non-synchronous fields. Furthermore the low noise and excellent grey-scale of the camera channel enable the conversion to be made with virtually no deterioration of the picture.

The BBC have been using type BD 863 cameras for many years at their Swingate station where the Eurovision programmes coming from Cassel are converted to 405-lines, and elsewhere Marconi cameras are being used successfully for the standards conversion of video tape recordings.

Where conversion involves non-synchronous fields (525/60 to 625/50 etc.) a correction waveform is applied to the recorded signal to compensate for the decay-time of the phosphor (normally a Willemite type), used for the recording monitor.

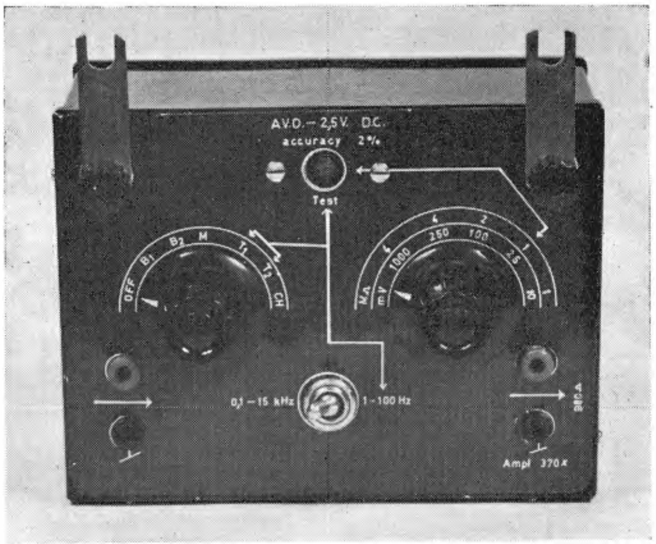
General arrangement of the equipment



The camera equipment

An A.C. Transistor Millivoltmeter with a High Input Impedance

By R. R. Vierhout*



After a preliminary explanation of the operation of feedback systems a description is given of a transistor millivoltmeter with an input impedance of 1 to 6MΩ, a frequency-range of 1c/s to 200kc/s, a full-scale deflexion in the most sensitive range of 10mV with an accuracy of 2 per cent of full scale.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 465)

THE effective input impedance of a transistor-amplifier can be increased by feedback which can be explained as follows.

In Fig. 1(a) and (b) amplifiers with negative feedback are shown. A useful quantity is the transfer impedance W, defined as:

$$W = V_o/i_i$$

In Fig. 1(a) the signal phase-shift from output to input is

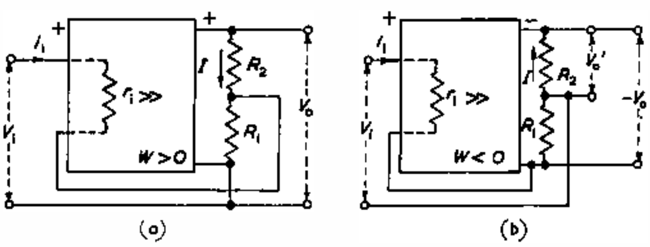


Fig. 1 (a) An amplifier without phase reversal with negative feedback (b) An amplifier with phase reversal with negative feedback

zero, hence W is positive, in Fig. 1(b) the signal phase-shift is π radians with W being negative.

For both circuits the following equations are valid:

$$V_i = i_i(r_i + R_1) + IR_1$$

$$V_o = i_iR_1 + I(R_1 + R_2) = Wi_i$$

Hence:

$$V_i = i_i\left[r_i + R_1 + \frac{R_1}{R_1 + R_2}\cdot(W - R_1)\right] = i_i\left[r_i + \frac{R_1R_2}{R_1 + R_2} + W\frac{R_1}{R_1 + R_2}\right] = i_i\left[r_i + \frac{R_1R_2}{R_1 + R_2} + (W/g)\right]$$

or:

$$V_i/i_i = r_i + \frac{R_1R_2}{R_1 + R_2} + (W/g)$$

if $g = (R_1+R_2)/R_1$, g is practically equal to the voltage amplification of the amplifier with feedback. The effective input impedance of the amplifier is increased by $R_1R_2/(R_1 + R_2) - (W/g)$, which impedance is situated in the output potentiometer R_1, R_2. If $(W/g) \gg r_i + R_1R_2/(R_1+R_2)$ the calculation can be simplified by considering that the voltage across R_1 is approximately equal to:

$$V_o \,.\, R_1/(R_1 + R_2) = V_o/g = W \,.\, i_i/g \simeq V_i$$

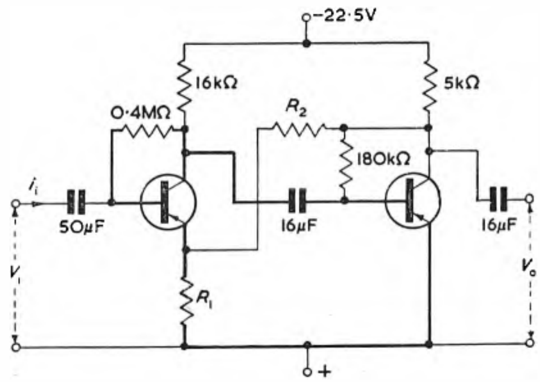


Fig. 2. Practical application of the circuit of Fig. 1(a)

Hence:

$$V_i/i_i \simeq (W/g) \quad \ldots\ldots\ldots\ldots\ldots (1)$$

The practical realization of the type of feedback of Fig. 1(a) is shown in Fig. 2. If the current amplification of the first stage is about 40, the current amplification factor of the second stage is 50 and the output impedance of the last stage is about 3kΩ, the transfer impedance is calculated as 40 × 50 × 3kΩ = 6MΩ. If $g = (R_1+R_2)/R_1 = 6$, the input impedance is increased for $W/g = 6/6 = 1$MΩ, according to formula (1).

Unfortunately parallel to the input impedance W/g (situated in the feedback resistor R_1), there is the collector-emitter-impedance, and the base bias resistor both in series with the very low input impedance of the second stage, indicated by the heavy lines in Fig. 2. This shunt limits the effective input impedance to about 1/3MΩ. If the amplifier introduces a phase-shift of 180° the feedback must be applied as shown in Fig. 1(b).

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The above photograph shows the a.c. transistor millivoltmeter.

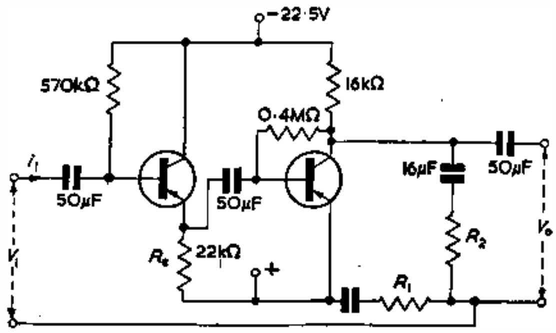


Fig. 3. Practical application of the circuit of Fig. 1(b)

Consider now the type of feedback shown in Fig. 1(b). When using a common emitter connexion an odd number of stages has to be used in order to obtain a phase-shift of 180° between input and output signal. However, feedback over three stages is extremely difficult to obtain on account of the phase lag at the higher frequencies due to the transistor action. With one stage a value of W of 0·5MΩ can hardly be obtained, hence the input impedance remains considerably under 1MΩ (W/g). The solution to this problem is the circuit in Fig. 3. The first stage is an emitter-follower which gives full current amplification without phase-shift. Because the input impedance of the second stage is far lower than the emitter resistance, R_e and the collector emitter impedance of the first transistor parallel to it, practically the full amplified current runs in the base of the second stage. The second stage shifts the phase of the signal 180°. If R_2=2kΩ and R_1=0·6kΩ, g becomes 3. The current amplification of the first stage is 40, the current amplification factor of the second stage is 50 and the output impedance is about 2kΩ, hence W = 40 × 50 × 2kΩ = 4MΩ. The effective input impedance is increased by the value W/g = 4/3MΩ. If R_1 has a value of 5kΩ, W/g appears to be about 6MΩ. Parallel to the resistor R_1, in which is situated the effective impedance W/g, there are no impedances of lower value. Hence this type of circuit is suitable for input stages of transistor millivoltmeters (patent pending).

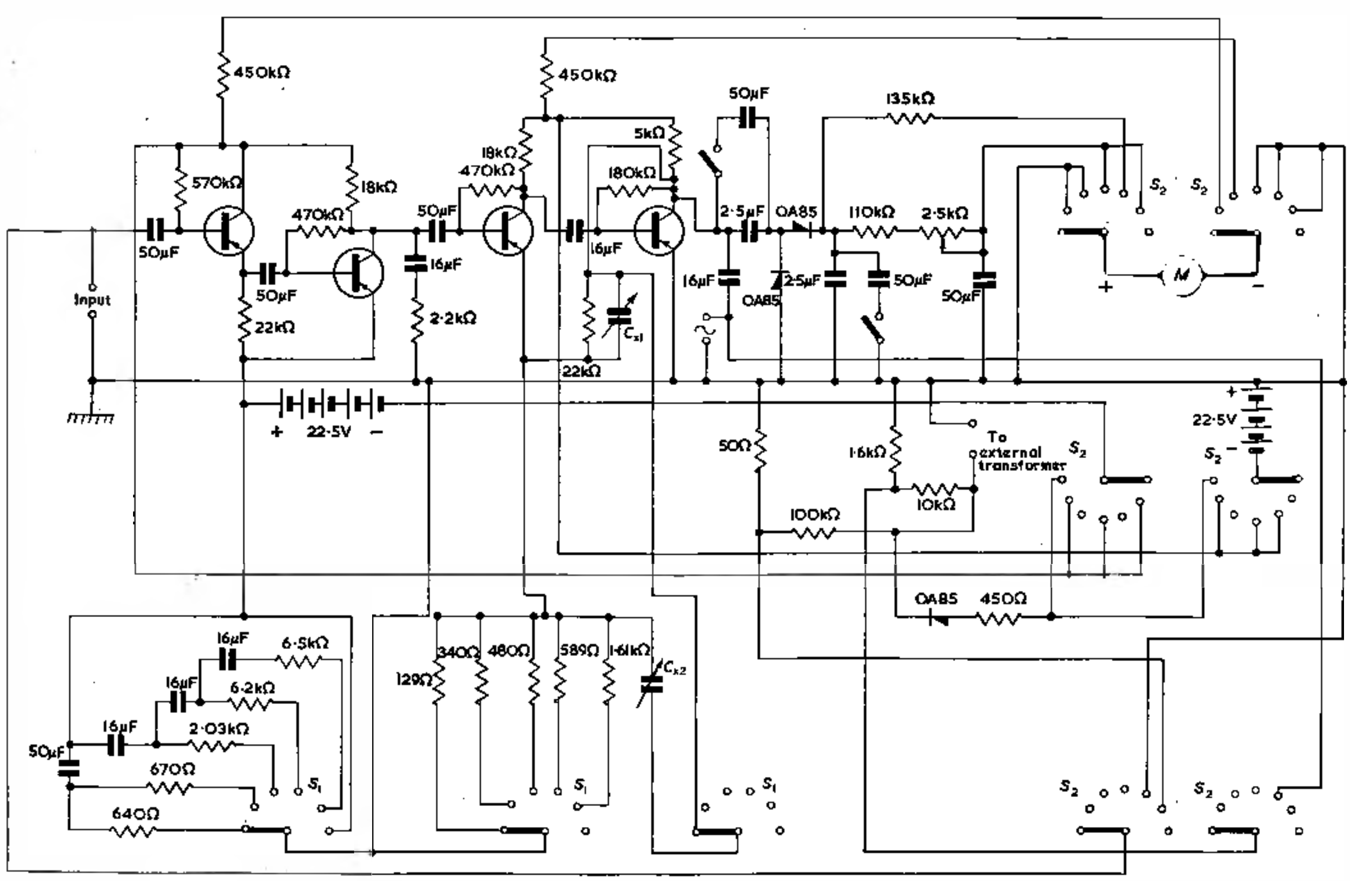
The Circuit of the Transistor Millivoltmeter

The input circuit used is shown in Fig. 3. The next stages are represented by Fig. 2. The batteries for the two circuits must be separated. One hearing-aid 22·5V battery is used for the first two stages, which require a current of 1·5mA, two of these serve the last circuit, which draws 3·5mA. Behind the last stage a rectifier with voltage doubling is connected. This circuit improves the linearity. As the a.c. output signal is not more than 3·7V an ordinary rectifier would show considerable deviation from linearity. A resistance in series with a 50μA full scale meter is connected to the germanium diode rectifier. The d.c. signal at the output of the rectifier is about 10V.

In the circuit of the last two transistors feedback is applied to counteract drift of amplification, due to transistors and the ageing of the batteries.

If the input signal is 10mV, the signal at the output of the first circuit is about 30mV, this being the input of the second circuit. The amplification of this circuit is about 120 and thus the output of it is 3·7V. Hence, because W of the second circuit is of the order of 40 × 50 × 3kΩ = 6MΩ, the input impedance of it is W/g =6 000/123 =

Fig. 4. The details of the complete circuit



50kΩ. This input impedance shunts R_2 of the first circuit and because $R_2 = 2{\cdot}2\text{k}\Omega$ a variation of W of 50 per cent in the second circuit causes only a variation of 2 per cent in the total resistance of R_2 and the above mentioned 50kΩ. Hence the amplification $g = (R_1+R_2)/R_1$ has an accuracy of 2 per cent.

The transistor millivoltmeter can be applied to a type 8 AVO multi-range meter. An extra output permits the use of the instrument as an a.c. amplifier with an amplification of 370.

The complete circuit, which is printed, is shown in Fig. 4.

In the first circuit (Fig. 3) the time-constants of R_2 and R_1 with their respective series capacitors are made equal in order to diminish a peak in the low frequency response. In the millivoltmeter design one switch provides for the ranges of 10, 25, 100, 250, 1 000mV full scale.

By means of the other switch the voltages of the batteries under running conditions are checked in position 1 and 2. Position 3 is the measuring position, positions 4 and 5 are intended for calibration of the amplification of the whole circuit making use of the available mains voltage. If a lead with a built-in transformer is put between the mains and a special input, in position 4 the AVO-meter indicates the mains voltage, because a part of the voltage of the secondary of the transformer is connected to the diode rectifier circuit. In position 5 a smaller part of this secondary voltage is connected to the input of the millivoltmeter. By adjustment of the 25kΩ potentiometer the pointer of the AVO-meter must indicate the same scale division as in position 4. This check is necessary in case of exhausted batteries.

In position 6 of the switch it is possible to raise temporarily the voltage of exhausted batteries with the transformer and lead. It appears that the batteries give a longer life if this charging process is done frequently.

There is a special switch which puts extra large capacitors in the rectifying circuit for frequencies lower than 100c/s. It appears that the frequency range for an accuracy of better than 2 per cent extends from 1c/s to 20kc/s if three OC71 and one OC73 are used in the last stage. If four OC44 are used the range extends from 1c/s to 200kc/s. The noise is negligible for short-circuited input. The input impedance is 1·3MΩ for the 10mV range, increasing by greater feedback in the less sensitive ranges to 6MΩ. The input-capacitance is 74pF on the 10mV range, decreasing to 21pF on the 1V range. This input capacitance is mainly due to negative feedback caused by the capacitance between output and input terminals, which decreases with the amplification factor.

The Advanced Gas-Cooled Reactor

The U.K.E.A.'s prototype Advanced Gas-Cooled Reactor (a.g.r.) at Windscale, Cumberland, is in an advanced state of construction and is due to be commissioned in April next year.

The reactors of the nuclear power stations at present being built for the Electricity Authorities are developments from Calder Hall and while they show considerable advances in size and in their engineering development, their fuel elements differ only in detail. The fuel for all is in the form of a uranium bar of approximately 1in diameter, sealed in a magnesium alloy can; the differences are limited to variations in finning of the can and in methods of supporting the fuel element in the reactor.

This type of fuel element imposes two limitations on reactor operation: the rate at which heat can be produced in the fuel and the temperature to which the carbon dioxide coolant can be heated. These limitations in turn restrict the power which can be generated in a reactor of a given size and the efficiency with which this power can be converted to electricity through the steam cycle.

There are two reasons why the rate of heat production, usually called the 'fuel rating', is limited to a mean of about 2·5MW per ton of fuel, and, the can surface temperature, which controls the carbon dioxide temperature, limited to a nominal figure of 450°C. If either of these figures were to be appreciably increased, the can itself would lose tensile strength as the melting point of the magnox alloy was approached, and at the same time the temperature of the uranium at the centre of the bar would reach 660°C, at which a change takes place in the crystal structure of uranium. This change, called the alpha-beta phase change, results in a sudden expansion of the uranium which would impose severe stresses on the can. These stresses, coinciding with a weakening of the magnox, would result in the rupture of many of the cans, allowing the escape of fission products which it is one of the main purposes of the can to prevent.

Thus, any large improvement in fuel rating or reactor temperature necessitates a complete change in the fuel element and this is the principal new feature of the Advanced Gas-Cooled Reactor. The effects of phase change in the uranium are avoided by using uranium oxide in place of metallic uranium, while the loss of strength in the can is overcome by replacing magnox by a more refractory material. Two materials are being tried out, the first is beryllium, a comparatively novel metal which absorbs very few neutrons, and the other the better known stainless steel. These changes, of course, introduce new problems. Uranium oxide is a much poorer conductor of heat than the metal, and so the rods of fuel have to be made much thinner. This means a greater proportion of canning material to fuel and a less efficient system from the nuclear physics aspect; also, the presence of two oxygen atoms to every uranium atom reduces the proportion of fissile atoms in the fuel. For these reasons, some enrichment of the fuel, by artificially increasing the percentage of uranium 235, is necessary, though this is partly offset by using beryllium in place of magnox for the can, since it captures far fewer neutrons.

However, it is expected that with this new fuel, the rating can be increased to over six times the rating in Calder Hall, which means that over six times the power can be produced in a reactor of the same size. In addition, it will be possible to increase the carbon dioxide temperature by over 200°C, which will raise the efficiency of the steam cycle by something like a third.

Studies indicate that the a.g.r. system should be capable of substantially lower capital costs than magnox stations and have confirmed that it offers the best prospect for the early achievement of competitive nuclear power in the United Kingdom.

Thus the Advanced Gas-Cooled Reactor can be regarded as the successor to the magnox reactor for nuclear power stations coming into operation in the late 1960's. Like them, it will be a source of plutonium for the initial fuelling of commercial fast reactors which are expected to be introduced in the mid 1970's.

The prototype reactor will produce 100MW of heat with a net electrical output of some 28MW.

For the accurate observation of the high temperatures English Electric's Data Processing & Control Systems Division at Kidsgrove is supplying temperature scanning equipment based on their new 'Datapac' transistorized logical elements. The equipment will be used to monitor the vital temperatures including those of the fuel cans, the concrete shielding, the coolant gas and the graphite.

In all 1 250 thermocouples are used to measure temperatures at critical points in the reactors, of these some 250 provide a representative sample of the temperatures of the new fuel cans, 200 are used to record the coolant gas temperatures at the outlets of fuel channels in the graphite moderator block, 250 measure temperatures in the graphite itself while the remainder are distributed to measure temperatures of such important components as the biological shield and other vital structural items.

The equipment scans the thermocouples in sequence, tests each in turn to ensure that the temperature measured lies within specified limits and gives an alarm should the reading be high. The equipment also indicates the exact position of the high reading thermocouple, it records the extent by which the set temperature limit has been exceeded, the time at which the event occurred and automatically provides for record purposes a printed data sheet of all the information obtained.

Complementary Transistor Circuits

By F. Oakes*, A.M.I.E.E., A.M.Brit.I.R.E., and C. Thompson*

The availability of npn as well as pnp transistors has led to the development of complementary techniques which provide a powerful new tool for the circuit designer. A number of circuits useful for pulse and digital applications are described. Several of these have no thermionic valve circuit equivalent, and provide faster operation or greater flexibility than could be obtained with only one single operating polarity.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 466)

ONE of the important differences between thermionic valves and transistors is the availability of two basically different types of transistor. Thermionic valves are restricted to a single polarity only, namely, with anode potentials which are positive with respect to cathode. In the case of transistors, npn types are available with operating polarity analogous to the valve. Their operation is based on conduction by negative charge carriers (electrons), and they operate with collector potentials which are normally positive with respect to emitter.

The alternative pnp type operates normally with collector potentials negative with respect to emitter, operation being based on conduction by positive charge carriers (holes).

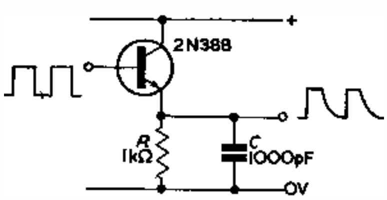


Fig. 1. Conventional emitter-follower circuit using npn transistor

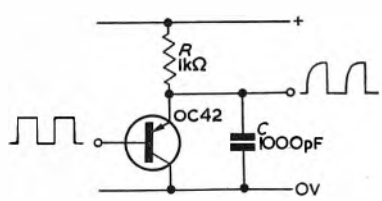


Fig. 2. Conventional emitter-follower circuit using pnp transistor

The availability of two different operating polarities provides new possibilities for the circuit designer, and useful transistor circuits have been developed, for which there are no valve equivalents.

In complementary circuits which have been published so far, the emphasis has generally been placed on ease of bias and coupling arrangements, or on matched symmetrical properties useful in push-pull amplifiers. More recent developments indicate however that more significant advantages can be gained by exploiting the availability of complementary polarity in pulse and digital circuits. Use can be made of this in two distinct ways. One, to make available two alternative complementary versions of a given circuit, the other, to combine complementary transistors in a single circuit. An example of the first is a pulse generator which produces positive-going pulses using an npn transistor, but which can be made to produce negative going pulses by substitution of a pnp transistor and reversal of operating polarities. An example for the second type of circuit is the complementary multivibrator circuit in which one npn and one pnp transistor are used in conjunction.

The principles outlined above are illustrated in the following paragraphs by a number of typical examples.

Emitter-Followers

When an emitter-follower has to feed a capacitive load, difficulties are often encountered in obtaining pulses with fast leading as well as trailing edges. This can be understood by considering the conventional circuit in Fig. 1 which is analogous to the thermionic valve cathode-follower. The positive-going edge of the input signal drives the emitter-follower into conduction. A low output impedance is therefore available, and the capacitor charges rapidly. When the input voltage goes negative, however, the emitter-follower is cut off and the capacitor discharges slowly with a time-constant RC. In a practical case, the emitter-follower output impedance may be of the order of 20 to 100Ω depending on drive source impedance. Hence, with the values shown in Fig. 1 the rise-time will be at least an order of magnitude faster than the fall-time. Reducing the value of R increases the loading on the drive source, and causes a heavy current to flow through the transistor during the whole duration of positive excursion. The circuit as shown is thus only suitable if the positive edge alone needs to be fast. A fast negative going edge can be obtained, with a slow positive going edge, by using a pnp transistor as shown in Fig. 2.

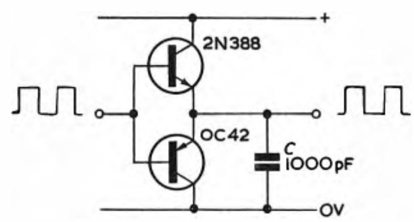


Fig. 3. Complementary emitter-follower

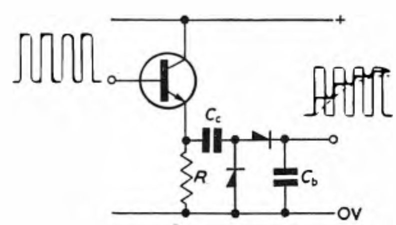


Fig. 4. Diode pump for positive going staircase waveform

If both positive and negative edges have to be fast, the complementary emitter-follower shown in Fig. 3 provides an excellent solution of the problem. During the positive-

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going edge, the npn transistor is conducting, its low output impedance enables the capacitor to be charged rapidly, and a fast edge obtained. During the negative-going edge, the pnp transistor takes over, providing an equally rapid discharge of the capacitor. A further advantage lies in the fact that no current is consumed by the emitter-followers, apart from the drive, load and leakage currents, whereas the conventional emitter-follower requires a current through the emitter series-resistor.

A practical example for the correct choice of an emitter-follower is the diode pump or 'cup-and-bucket' circuit shown in Fig. 4. A positive going staircase waveform with sharp edges can be obtained from the npn emitter-

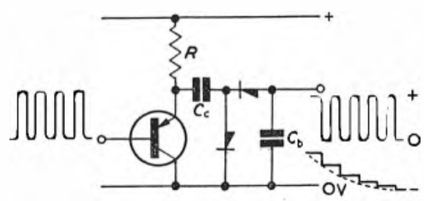


Fig. 5. Diode pump for negative going staircase waveform

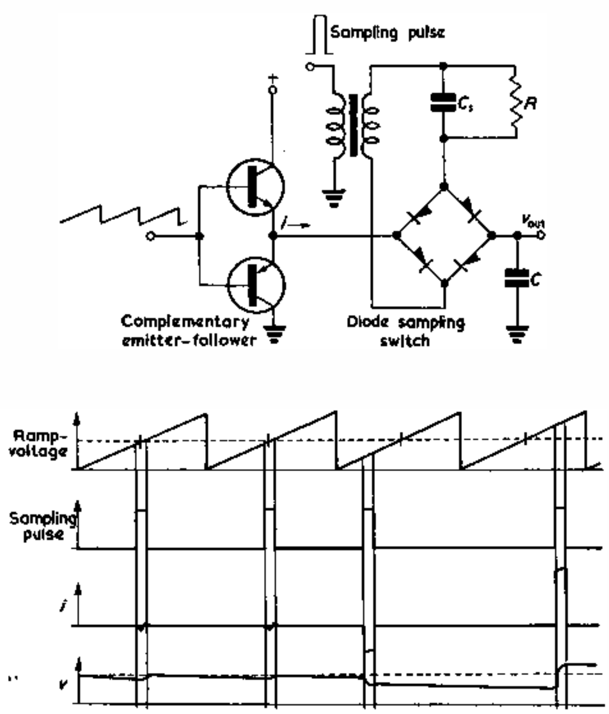


Fig. 6. Complementary emitter-follower driving ramp sampling circuit

follower. For a negative going staircase, the pnp circuit shown in Fig. 5 gives equally good results. The slow edges merely affect the discharge time of the 'cup' capacitor C_C, and a complementary circuit is justified only if fast operation is required.

A practical example for the case of an emitter-follower feeding a capacitive load where fast edges both positive and negative going are demanded, is in connexion with a sampling circuit as shown in Fig. 6. The phase relationship between a recurrent ramp and a sampling pulse is detected here by charging a capacitor to the instantaneous value of the ramp voltage during the period of sampling. Since this capacitor must be charged or discharged rapidly as the phase relationship changes, a substantial current in either direction may be required during the sampling period. A complementary emitter-follower is therefore required to provide the necessary power gain as well as a low output impedance whether driving positive or negative.

Complementary Sampling Circuit

The sampling circuit shown in Fig. 6 can be further improved by replacing the four diodes by a complementary sampling switch which requires considerably less power from the sampling pulse source and provides a very much cleaner output. This circuit is shown in Fig. 7.

The sampling pulse which is applied via an isolating pulse transformer drives the base of the npn transistor positive, and the base of the pnp transistor negative. Both transistors are driven into conduction for the duration of the sampling pulse, during which time they connect the output of the complementary emitter-follower to the storage capacitor C. Rectification in the sampling pulse circuit charges the isolating capacitor C_s. This charge is of such polarity and magnitude that the sampling switch

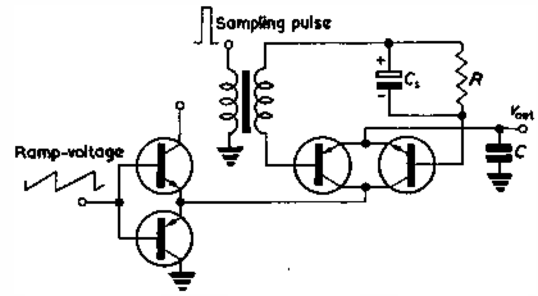


Fig. 7. Ramp-sampler using complementary transistors

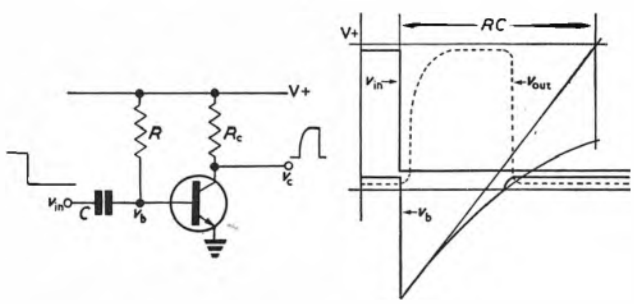


Fig. 8. Circuit for generating pulse with pre-determined duration and fast back edge

transistors are cut off during the standby period. The high resistance R across C_s serves to establish an optimum condition. If no such discharge resistor were provided, and if their cut-off current I_{co} were zero, the sampling transistors would be permanently biased off.

Single Transistor Pulse Generators

A pulse generator is shown in Fig. 8 which will produce a pulse of predetermined length and with a fast back edge, such as would be used to provide a delayed trigger.

Here, a single npn transistor is used to produce a positive pulse. The analogous circuit with a pnp transistor will give a negative going pulse.

The input trigger pulse in the circuit shown is negative going for the npn transistor (it would be positive in the pnp equivalent). This trigger pulse cuts the transistor off and holds the input side of the capacitor near zero potential. If the input voltage step is approximately equal to the h.t. voltage V, the base side of the capacitor falls abruptly to approximately $-V$ and immediately begins to rise exponentially towards the positive rail $+V$, with a time-constant CR.

Soon after the junction of C,R and the base of the transistor traverses zero potential, the base-emitter diode ceases to be cut off, and the junction is caught at a small positive potential. At the same time, the transistor is switched on. It is seen therefore that the collector potential, i.e. the output potential, rises when the input voltage is first applied, and falls back sharply after an interval determined by the input voltage, the rail voltage and the values of C and R. This interval is substantially independent of the transistor parameters—provided that the cut-off base current is sufficiently small to be ignored.

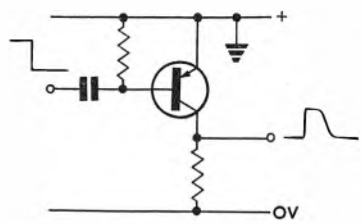


Fig 9. Pulse generator for fast leading edge

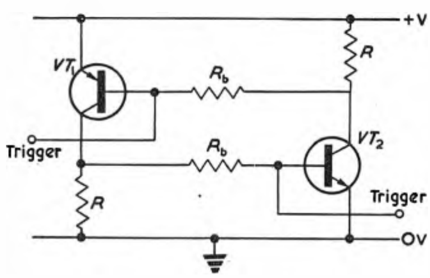


Fig. 10. Basic complementary multivibrator circuit

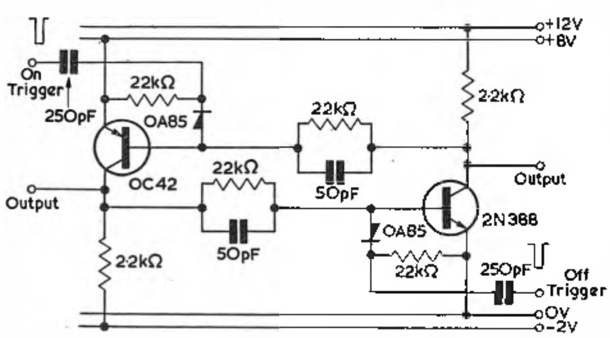


Fig. 11. Complementary bistable multivibrator circuit

total consumption being approximately equal to the on current of one transistor and independent of mark-to-space ratio.

A complementary multivibrator comprising one pnp and one npn transistor is shown in Fig. 10. Two stable states are available; the on-state with both transistors conducting and the off-state with both transistors cut off. This is a fundamental and important difference from the conventional case. The current drawn by this complementary circuit varies substantially over the duty cycle; from twice the value for one transistor to almost zero. Thus, if the

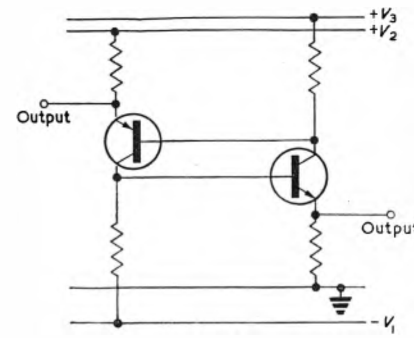


Fig. 12. Complementary direct coupled bistable multivibrator circuit

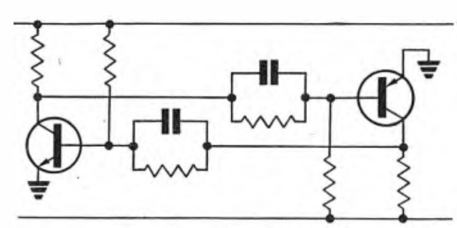


Fig. 13. Complementary bi-stable circuit

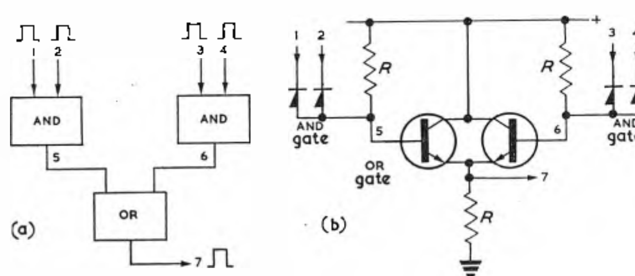


Fig 14. Mixed diode and transistor gating circuit

The circuit of Fig. 9 is very useful where an undelayed inverted pulse with a fast leading edge is required. Unlike the circuit of Fig. 8, however, the pulse duration is ill defined, because the base is driven on by the input voltage, and offers a low input impedance. Thus, the transistor parameters determine the charging rate of C, rendering this difficult to calculate or control. The transistor first works in saturation condition, eventually it goes through the active region when the charging rate is reduced, and a slow, ill defined back edge is produced.

Multivibrators

This family of cross-coupled circuits plays an important part in pulse and digital systems and has therefore received much attention in its valve as well as transistor versions. In the conventional Eccles-Jordan circuit form, two identical transistors are used so that one is in the on-state while the other is in the off-state. Thus, one or the other device is always carrying current, the

mark-to-space ratio is other than unity, a reduction in current consumption can be achieved by arranging for the on-state to be short and the off-state to be long. While this can be a great advantage especially in battery operated equipment, it can give rise to difficulties where power supply regulation is poor, because of the current pulses drawn from the supply rail.

Another important feature is that both transistors switch on simultaneously, thus the low output impedance of each transistor is capable of supplying the heavy drive current required by the other. This ensures a rapid switch on. In the conventional transistorized multivibrator, the transistor which goes into saturation is driven by the transistor which goes into cut-off, and as the switching phase progresses, the latter transistor becomes increasingly unable to provide the drive for the former. Faster switch-on times can therefore be obtained with complementary techniques. The switch-off time of a multivibrator is largely determined by the time-constant of the collector resistor and cross coupling capacitor. Since in the complementary circuit a relatively small capacitor suffices to produce the on edge, the off edge can be made correspondingly faster.

The operation of this circuit can be understood by considering the on-state first. Both transistors are bottomed, each carrying a current V/R in the collector circuit and V/R_b in the base circuit. This state is stable. It can be changed into the off-state by driving the base of the npn transistor negative, so that its collector current is cut off. This will cause its collector to go positive, taking the base of the pnp transistor positive and cutting this transistor off also. Both transistors are now off, this also being a stable state. To return this circuit to the original condition a negative going trigger is applied to the base of the pnp transistor. This will switch the transistor on, its collector potential will therefore rise and, via the cross-coupling, cause the base of the npn transistor to go positive, switching this on also.

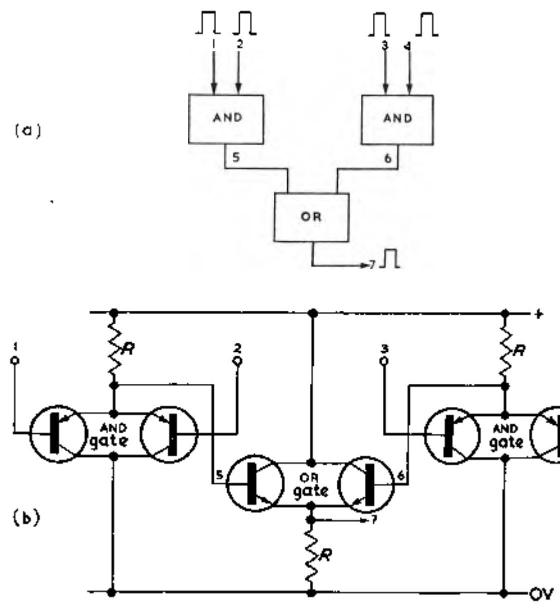


Fig. 15. Transistor gating

Fig. 11 shows a practical version of this circuit with supply rails at 0 and 8V and back-bias rails 2V beyond the supply rails to ensure stability of the off-state. Trigger capacitors and diodes are also shown. The cross-coupling resistors are dimensioned to ensure reasonable saturation conditions. They are bridged by 'speed-up' capacitors in the usual manner. Fig. 12 shows a variant of this circuit with direct cross-couplings. Emitter resistors are included as shown, and heavy loads can be connected across these. In contrast to conventional multivibrators, these loads will not tend to pull the transistors out of saturation.

Another interesting form of complementary multivibrator circuit is shown in Fig. 13. Again it can be seen that in the on-state, both transistors are saturated, in the off-state, both are biased in reverse. A number of variants can be devised of this circuit, e.g. by insertion of emitter loads, or by biasing arrangements for monostable or free-running operation.

Transistor Gating Circuits

PNP and npn transistors can be used in a variety of logical circuits, offering a combination of relatively small driving power requirements, low output impedance and great versatility.

The use of emitter-followers for gating purposes is advantageous where cascaded gates have to be used. Such a case is shown in the block diagram Fig. 14(a). Here, the AND gates are conventional diode gates. The OR gate could also be made a diode gate, with an emitter-follower in the output of each AND gate to provide the required power gain. However, as shown in Fig. 14(b), the need for the diode OR gate is eliminated by arranging the emitter-followers to perform the gating function. For the case shown, i.e. with positive one-state and negative zero-states, npn transistors are required. In a similar manner, the diodes comprising the AND gates can be replaced by pnp transistors, as shown in Fig. 15.

It should be noted that the output impedance of these gates is low in the one-state for an OR gate and in the zero-state for an AND gate. Conversely, the output impedances of OR gates in the zero-state or AND gates in the one-state is high, being that of the load resistor R. Hence, since the replacement of diodes by transistors in these gates enables the load resistors, R, to be reduced in value without increasing the loading on the drive sources, this in turn means that faster operation of the gates can be obtained.

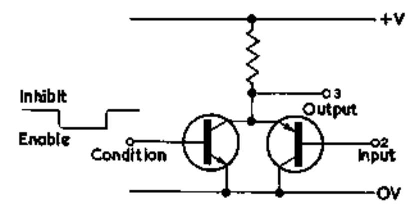


Fig. 16. Complementary gate

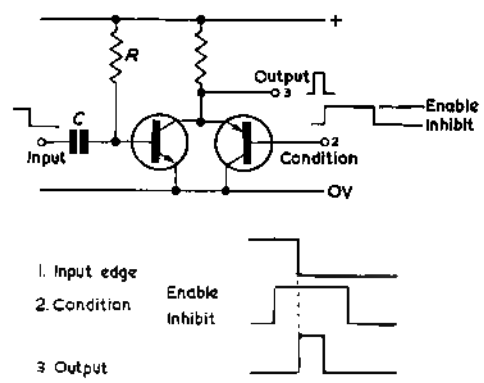


Fig. 17. Variant of complementary gate

For 'inverted' logic, i.e. with positive zero-states and negative one-states, pnp transistors will be required for the OR gates and npn transistors for the AND gates.

Complementary gates can be very useful where a small amplitude signal has to inhibit or enable the passage of a larger signal through the gate. The circuit of Fig. 16 contains a grounded emitter amplifier which can be driven into cut-off or saturation by a small input voltage on its base terminal (1). In the saturated or 'inhibit' condition, the output terminal (3) is permanently held at zero potential. With the npn transistor in the cut-off or 'enable' condition, the potential at the output terminal (3) will rise to that of the input on terminal (2). Thus, the negative condition potential switches the npn transistor off and allows the pnp transistor to function as an emitter-follower between terminals (2) and (3). This circuit is therefore particularly useful for gating analogue voltages or states.

A variant of this circuit is shown in Fig. 17. Here the npn transistor is connected as in Fig. 8, and the functions of terminals (1) and (2) are reversed. An edge applied to input terminal (1) will be inhibited or passed according to the state at condition terminal (2)

To obtain the 'enable' condition terminal (2) is held positive and the emitter-follower is seen to be cut off. In the quiescent state of the gate, i.e. before the edge is applied to input terminal (1), the npn transistor is saturated and output terminal (3) at zero rail potential. When a negative going edge is applied, the npn transistor is cut off and the output goes positive remaining positive for a time dependent on the npn circuit pulse length as described earlier, or on the duration of the 'enable' condition, whichever is the shorter. In order to 'inhibit' the gate, terminal (2) must be held at zero potential. The emitter-follower is then saturated and the output terminal is held at zero potential by the emitter-follower, irrespective of current flow in the npn transistor.

Conclusions

It has been shown that the availability of npn as well as pnp transistors provides a powerful tool for the circuit designer. Not only can single polarity circuits be made available in both versions, but new circuits making use of combinations of both types of transistor, and of which there are no valve equivalents, can be devised. Some of these circuits are symmetrical in their circuit configurations, but only in exceptional cases is it necessary for the pnp and npn units used to have symmetrically matched properties. Transistors of different types and make are therefore quite acceptable in most complementary circuits, and it is expected that the increasing availability of npn transistors will result in a more widespread use of complementary circuit techniques.

Acknowledgment

The authors are indebted to other members of the Transistor Applications Laboratory of Ferguson Radio Corporation on whose experimental work some of the circuit descriptions are based, and to the Directors of this Company for permission to publish this article.

A Simple Circuit for Producing a Voltage Proportional to the Width of a Repetitive Pulse

By H. C. Bertoya*, A.M. Brit. I.R.E.

A circuit is described which converts a repetitive pulse into a direct voltage. The magnitude of the direct voltage is proportional to the width of the pulse. The need for the circuit arose when it was found that the pulse would not drive an integrating motor directly. The circuit is analysed and details of its performance, together with that of the integrator, are given.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 466)

A METHOD of measuring the flow of a liquid has been developed using an ultrasonic beam. The electrical output of the associated measuring system consists of a repetitive pulse of constant height and frequency—in this case, 25V and 800c/s—and the width of the pulse is proportional to the rate of flow.

A moving-coil meter will average the area under the pulse satisfactorily, and may be employed to give a reading directly proportional to pulse width (and hence flow). However, in addition to direct flow indication, it was required to integrate the flow-rate with respect to time using a motor-type integrator. Unfortunately, the inertia of the motor made it appear as a very large capacitance which caused the output stage supplying the motor to overload. To provide a sufficiently low source-impedance would have required a very large available power. The variable width pulse is, therefore, first averaged by an *RC* network giving a direct voltage whose amplitude varies with pulse width; this voltage is used to drive the integrator motor.

The object is accomplished in three stages (Fig. 1). First the variable width pulse is converted into a variable height pulse; then the peak of this pulse is detected; finally the peak voltage drives the integrator motor via a push-pull cathode-follower output stage.

Consider the circuit of Fig. 2(a), where E is the input voltage and e is the output voltage. The input and output waveforms are shown in Figs. 2(b) and 2(c) respectively.

Let $e_n' = K_1 e_n \qquad b_n = K_2 a_{n-1}$

$$e_n + b_n = E \qquad e_n' + a_n = E \qquad e_0 = E$$

Thus one may write:

$$\begin{aligned} e_n &= E - b_n \\ &= E - K_2 a_{n-1} \\ &= E - K_2(E - e_n'_{-1}) \end{aligned}$$

Continuing the substitution for $e_n'_{-1}$ and re-arranging terms:—

$$e_n = E(1-K_2) + E(1-K_2)K_1K_2 + E(1-K_2)K_1^2K_2^2 + e_{n-3}K_1^3K_2^3 \quad \ldots\ldots \quad (1)$$

Now, $e_{n-n} = e_0 = E$ so that equation (1) may be written:—

$$e_n = E(1-K_2)[1 + K_1K_2 + (K_1K_2)^2 + \ldots\ldots + (K_1K_2)^{n-1}] + E(K_1K_2)^n \quad \ldots\ldots \quad (2)$$

The first term in equation (2) is a geometric series in which K_1 and K_2 are less than unity. Therefore as n tends to infinity $E(K_1K_2)^n$ tends to zero, and equation (2) may be rewritten:—

$$\underset{n\to\infty}{e_n} = \frac{E(1 - K_2)}{(1 - K_1K_2)} \quad \ldots\ldots \quad (3)$$

Now consider the input wave shown in Fig. 2(d) and let $K_1 = \exp(-t_1/CR) \quad K_2 = \exp(-t_2/CR)$ and $t_1 + t_2 = \tau$

Equation (3) may now be written:—

$$\underset{n\to\infty}{e_n} = E\left[\frac{1 - \exp(-\tau/CR)\,.\,\exp(t_1/CR)}{1 - \exp(-\tau/CR)}\right] \quad \ldots\ldots \quad (4)$$

* *British Scientific Instrument Research Association.*

Insert limits:—

$$\text{Let } t_1 \to 0 \text{ then } e_n \to E\left[\frac{1 - e^{-\tau/CR}}{1 - e^{-\tau/CR}}\right] \to E\,.\,1$$

i.e. $e_n \to E$ as $t_1 \to 0$

$$\text{Let } t_1 \to \tau \text{ then } e_n \to E\left[\frac{1 - e^{-\tau/CR}\,.\,e^{\tau/CR}}{1 - e^{-\tau/CR}}\right] \to E\,.\,0$$

i.e. $e_n \to 0$ as $t_1 \to \tau$

In a given system $e^{-\tau/CR}$ is a constant, m say, so that equation (4) becomes:—

$$e_n = E\left[\frac{1 - m\,\exp(t_1/CR)}{1 - m}\right]$$

$$\text{and } de_n/dt_1 = \left[\frac{Em}{(1-m)CR}\right]\,.\,[\exp(t_1/CR)]$$

The term $Em/(1-m)CR$ is a constant but the term $\exp(t_1/CR)$ causes a change in slope as t_1 varies from 0 to τ. If CR is made large in comparison with t_1, the non-linearity may be made very small.

The peak of the pulse given by equation (4) is measured by a circuit consisting of a semiconductor rectifier and an RC network. The problem of non-linear operation of the rectifier at small signal levels is avoided in this particular application. Because of considerations relating to the flowmeter system as a whole, the pulse width does not vary over the complete range $t_1 = 0$ to $t_1 = \tau$. A time equivalent to a phase margin of about 10 degrees is retained at either end of the range and so with the pulse described the variation in peak voltage is from approximately 0·7V to 24·3V.

The detected voltage is now passed to a double-triode cathode-follower, with the motor connected between the

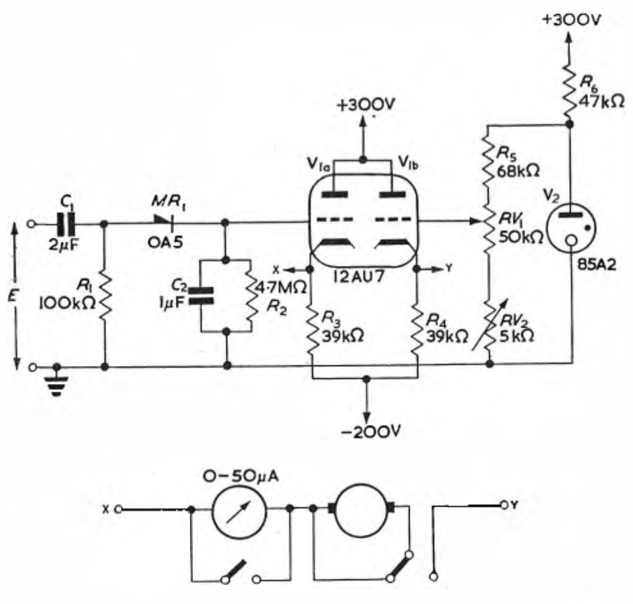


Fig. 1. Pulse-width to d.c. convertor and integrator

cathodes. Backing-off facilities are provided to enable a zero count to be set at zero flow. To avoid the difficulty of observing extremely low count-rates, the stage is balanced with the aid of a series microammeter. The integrator motor is started and stopped by means of a switch. A further pole on the switch short-circuits the motor when switched off in order to prevent inertial overrun.

Performance

It was desired to test the whole integrating circuit rather than merely the pulse-width to d.c. conversion; thus the

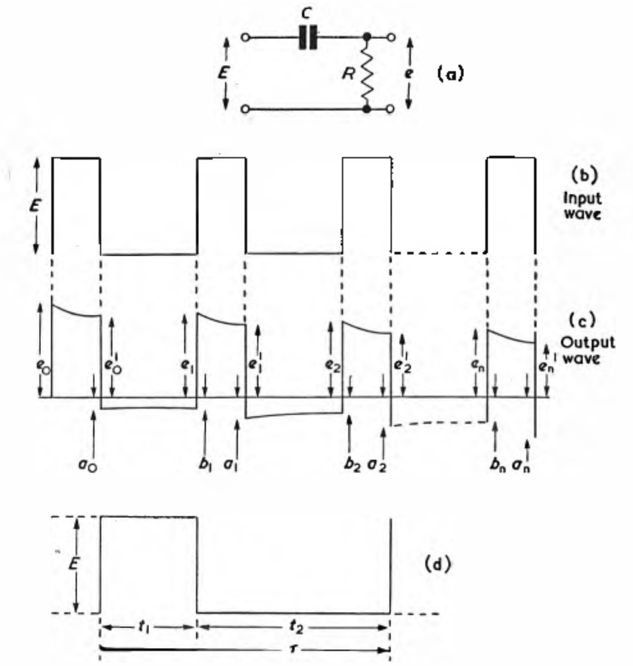


Fig. 2. Width to height conversion circuit and waveforms

change in voltage for change in pulse-width was not measured directly. Instead, the reading of the integrator —after running for an accurately timed 10min—was compared with the reading of a potentiometer and galvanometer which averaged the pulse directly.

The input phase was altered from 10° to 350° in 30 steps; the 'best-fit' straight line and the standard deviation were calculated.

Because of stiction, the straight line will not pass through the origin, i.e. a finite voltage must exist before the motor will rotate. The start of the characteristic will be curved and so the first four points were disregarded. This reduced the range of operation of the integrator, within the limits of accuracy stated below, by 20° from 30° to 350°.

The standard deviation was 0·7, with a maximum integrator reading of 1 475. The error this represents at various input pulse-widths is:—

Width	*error*	
350°	0·05%	(Corresponding to a meter f.s.d.)
180°	0·1%	(Corresponding to ½ f.s.d.)
30°	1·0%	(Corresponding to $1/17^{th}$ f.s.d.).

The largest error in this range, as a percentage of f.s.d., is approximately 0·1 per cent.

It may be inferred that the averaging network itself achieves a comparable or higher degree of accuracy.

Acknowledgment

The author wishes to thank the Director and Council of the British Scientific Instrument Research Association for permission to publish this article.

A Sensitive Transistor Recording-Pen Amplifier*

By K. G. Beauchamp†, C.G.I.A., A.M.Brit.I.R.E.

It is shown that if a recording pen movement is connected in a bridge circuit in which two arms of the bridge consist of transistor elements, then a high sensitivity can result.

With this arrangement a constant current is drawn from the supply source and since this need not exceed the maximum pen driving current then low transistor power dissipation, and hence low temperature drift can result.

(Voir page 461 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 466)

A difficulty that arises when a transistor amplifier is used to supply the direct-current for operation of a two-terminal recording pen is that the low impedance of the pen movement severely limits the gain possible with the balanced type of difference amplifier commonly employed. One method of overcoming this is by lowering circuit impedance values, but leads to a twofold disadvantage of increased current consumption and increased dependence of I_{co} on operating temperature with consequent zero drift error.

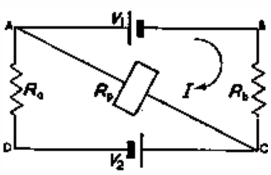


Fig. 1. Basic bridge circuit

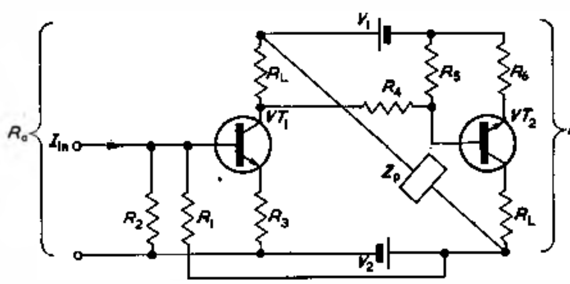


Fig. 2. Bridge circuit using transistor elements to replace resistance arms R_a and R_b

An alternative method is to include the pen movement effectively in series with two transistors in a balanced bridge circuit.

Principle of Operation

This is shown in Fig. 1, where the pen movement, R_p, is connected between terminals A and C of the bridge circuit, consisting of resistors R_a, R_b, and two fixed supply potentials, V_1 and V_2.

In the quiescent state $R_a = R_b = R$, and since $V_1 = V_2 = V$, then a circulating current, I, flows and no potential difference exists across the pen terminals.

Consider now R_a to increase in resistivity by δR while R_b decreases by the same amount.

The change in pen current will be found to be:

$$2\delta i = I \,.\, \frac{2\delta R}{R - 2R_p} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

and if R is made not much greater than $2R_p$ then a high sensitivity can result.

In a practical circuit R_a and R_b are replaced by transistor amplifying stages as shown in Fig. 2 and it is convenient to obtain 'single-ended' input connexions by driving VT_2 from VT_1 to obtain differential action.

One advantage of this bridge circuit is that of low current consumption and the circulating current may be reduced to the value of the peak pen current required.

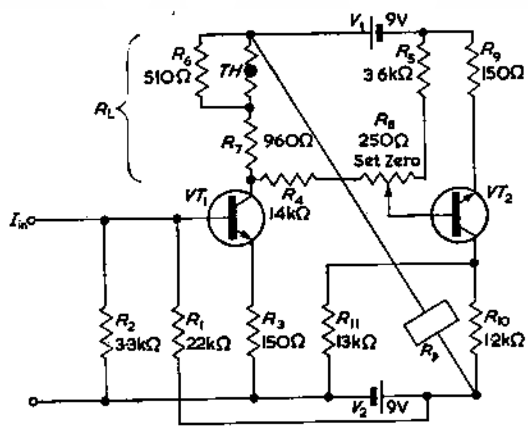


Fig. 3. Complete circuit including temperature drift compensation

Circuit Analysis

Using the base-potentiometer, emitter-regeneration form of gain stabilization, the base input current for VT_1 will be reduced by a fraction:

$$k_1 = \frac{R_2}{R_2 + h_{11(e)}}$$

where $h_{11(e)}$ is the input impedance of VT_1 in the grounded emitter configuration.

Hence the collector current change for an input current i_{in} will be:

$$\delta i_1 = I_{in} \,.\, k_1\beta_1 \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

where β_1 is the current gain for VT_1.

The effective change in this resistive bridge arm R_a, is given approximately as:

$$\delta R_a = (i_1 R_a / I) = (I_{in}/I) \,.\, k_1 R_a \beta_1 \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

* *British Patent Application No. 42102/59 dated 10th. Dec. 1959.*

† *United Kingdom Atomic Energy Authority.*

and that for VT_2 as:

$$\delta R_b = (\delta i_1/I) \cdot \frac{R_L}{R_4 + R_5} \cdot k_2 \cdot R_b \cdot \beta_2 \quad \ldots\ldots (4)$$

Since δR_a must equal δR_b for balanced differential action, and letting $k_1 = k_2 = k$ and $\beta_1 = \beta_2 = \beta$,

$$R_L = \frac{R_4 + R_5}{k\beta} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

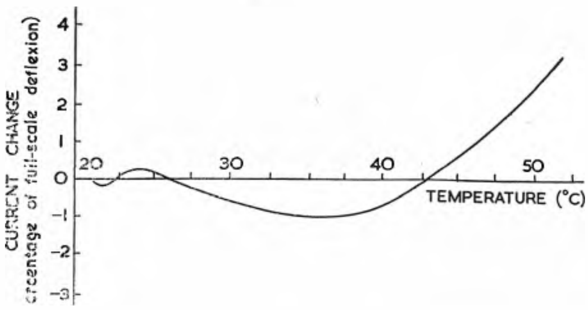


Fig. 4. Performance of the circuit shown in Fig. 3 expressed in terms of percentage change of full-scale deflexion with increase in ambient temperature

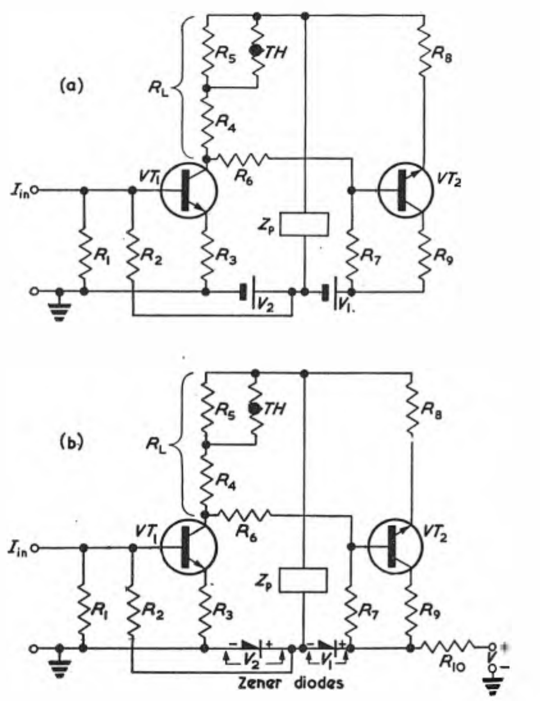


Fig. 5. (a) Modification to Fig. 3 allowing a single tapped supply source to be used

(b) A further modification enabling a two-terminal supply source to be used

Under this condition the amplifier current gain is, from equations (1) and (3):

$$A_i = (2\delta i/I_{in}) = \frac{2k\beta R}{R - 2R_p} \quad \ldots\ldots\ldots\ldots (6)$$

This is the maximum gain available and is reduced by a factor:

$$\frac{1}{1 + (A_i/2) \cdot (R_3/R_L)}$$

due to emitter regeneration.

Temperature Drift Compensation

With the circuit shown in Fig. 2, the I_{co} changes are additive and to compensate for this R_L is made temperature-dependent by inclusion of a Thermistor element in the load circuit. Suitable series and shunt resistors are required to obtain the correct compensating law and good stability may be secured in this way up to 50°C.

It will be appreciated from Fig. 2 that the transistors must be chosen such that the change of I_{co} with temperature must be greater for VT_1 than for VT_2. The exact amount is not important as any reasonable difference can be corrected in the manner described above.

The choice of Thermistor can be made using a trial temperature run using a purely resistive load for VT_1. The positive drift coefficient for the pen current can be equated in terms of load resistance change to secure the same coefficient. A Thermistor is then chosen to have a negative coefficient rather greater than this, since a linear change is not desirable, and the required change of slope obtained by adjustment of series and shunt resistors shown in Fig. 3.

Performance

An amplifier was constructed in accordance with the above design analysis, using silicon transistors. This was required to drive a pen movement having $R_p = 1\text{k}\Omega$ and I f.s.d. = 1mA, from a current source of 50μA maximum.

Allowing $I = 3\text{mA}$ and $V_1 = V_2 = 9\text{V}$, $R = 3\text{k}\Omega$ and for 0·5mA bias chain drain then $R_4 + R_5 = 18\text{k}\Omega$.

$h_{11(e)}$ for the transistors used was 2·1kΩ so that for $R_1 = 3{\cdot}3\text{k}\Omega$, $k = 0{\cdot}65$ and, from equation (6), $A_i = 156$.

Emitter regeneration was arranged to reduce this to the value required and under these conditions $h_{11(e)}$ increases to 5·6kΩ so that inserting a new value for $k_1 = 0{\cdot}38$ in equation (5) $R_L = 1{\cdot}2\text{k}\Omega$.

The complete circuit is given in Fig. 3, where R_L includes the Thermistor compensation network. The measured loop gain was found to exceed the figure calculated by some 15 per cent and necessitated an increase in R_3 to that given in Fig. 3, otherwise circuit static conditions were found to be substantially as calculated.

Variations in pen current for change in ambient temperature are shown in Fig. 4 and found to lie within ±1 per cent between 20° and 45°C.

Long-term changes have been recorded and show a change of less than 1 per cent over a 15 hour period and ±1 per cent over a period of several months.

Input/output linearity of the amplifier was measured to be within 1 per cent over the operating range.

Single-Battery Operation

By interchange of the positions of V_1 and R_a in Fig. 1, a single tapped battery circuit may be arranged as shown in Fig. 5(a).

The circuit design follows the same procedure as that given above.

Alternatively $V_1 + V_2$ may be replaced by Zener reference diodes and a single-battery supply used (Fig. 5(b)). In this latter case the current drawn from V_s should be several times the circulating collector current, I.

Acknowledgements

The author wishes to thank the Director of the Atomic Weapons Research Establishment for permission to publish this article, and to record his thanks to Mr. G. Beaumont of this Establishment who was responsible for the experimental work.

Short News Items

The Royal Society announces that awards of grants by the Paul Instrument Fund Committee have been made as follows:

£2 500, with the probability of further grants totalling up to £3 000, to Dr. J. H. Sanders, university lecturer and demonstrator in physics, Clarendon Laboratory, Oxford, for the construction of an optical maser.

£3 000 to Mr. H. W. Gosling, lecturer in the department of engineering, University College of Swansea, for the construction of an instrument for checking the stability of the standard ampere.

The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Physical Society, the Institute of Physics and the Institution of Electrical Engineers, was set up in 1945 "to receive applications from British subjects who are research workers in Great Britain for grants for the design, construction and maintenance of novel, unusual or much improved types of physical instruments and apparatus for investigations in pure or applied physical science."

The United Kingdom Atomic Energy Authority has issued a new edition of its 'Glossary of Atomic Terms' which has been prepared for journalists, teachers and laymen in general. It lists and explains in simple, concise language some five hundred terms in common use where nuclear matters are handled. The definitions range from 'Absorber' through 'Fission Yield' and 'Proton' to 'Zeta'—Zero Energy Thermonuclear Apparatus: the equipment used at Harwell for research into possible fusion reactions and various aspects of plasma physics. The glossary is available from H.M.S.O. at 3s. 6d.

The Sixth Annual Conference on Magnetism and Magnetic Materials will be held in New York City from 14 to 17 November, 1960, at the New Yorker Hotel. This conference is sponsored jointly by the American Institute of Electrical Engineers and the American Institute of Physics, in co-operation with the Office of Naval Research, the Institute of Radio Engineers and the Metallurgical Society of the American Institute of Mechanical Engineers. Authors should submit titles and abstracts of proposed papers by 26 August to A. M. Clogston or R. C. Fletcher, Programme Chairmen, Bell Telephone Laboratories, Murray Hill, New Jersey. Further conference details can be obtained from the Local Chairman, L. R. Bickford, Jr, IBM Research Centre, Yorktown Heights, New York.

A Reactor Simulator has been specially designed and constructed by General Precision Systems Ltd, Aylesbury, for the Reactor School, Atomic Energy Research Establishment, Harwell. It will be used for instruction in the theory and practice of reactor control. The control desk of the simulator is basically similar to that of a nuclear power station but is capable of registering power levels accurately with the remarkable range of one tenth of a watt to one thousand megawatts so that experience in handling all types of reactors may be obtained. The temperatures of the uranium and graphite are computed and recorded continuously, push button controls are employed throughout and a unique feature is the digital display of the position of fine and coarse control rods.The Reactor School runs standard courses which last for sixteen weeks. The syllabus is designed to give graduate engineering and physics students the necessary background of technical training and information required for the basic theory and operational techniques of reactors.

Welwyn Electrical Laboratories Ltd of Bedlington, Northumberland, have announced that the title of the Company has been changed to Welwyn Electric Ltd.

Bonn University Observatory is shortly to begin an intensive investigation into the temperatures prevailing in interstellar gas. To this end, special amplifying equipment employing travelling wave tubes has been manufactured for the University by Marconi's Wireless Telegraph Co. Ltd of Chelmsford, England. This consists of a dual channel amplifying system incorporating two travelling wave tubes in cascade in each channel. The radio telescope, a parabolic mirror of 83ft diameter mounted on a pyramidal tower about 60ft high, scans the sky picking up the cosmic continuum radiation emanating from galactic and extra-galactic radio sources under observation. The signals in the neighbourhood of the hydrogen line frequency (1 420Mc/s) are amplified by one pair of travelling wave tubes, the other pair being used to amplify reference noise signals from a resistor at a known temperature. The outputs from the two amplifying channels are detected, integrated and compared and the effective cosmic temperature determined. From these, data contour maps are prepared and so accurate has the system proved in initial tests that a discrimination of 0·1°K has been achieved. The research programme at Bonn University Observatory is being carried out under the direction of Professor Becker, who is assisted by Herr Heinz G. Müller.

The British Standard Code of Practice for the reception of sound and television broadcasting (C.P.327.201:1960) which has just been revised, incorporates developments in sound and television broadcasting since 1951, when the code was first published under the title 'Broadcast reception, sound and television by radio'. The revised Code takes account of the introduction of television broadcasting services in Band III, and the introduction of frequency-modulated (f.m.) sound broadcasting in Band II (v.h.f.). Particular attention has been paid to the question of aerial systems for these Bands, and to the question of combining such systems with aerials for Band I television. The existing recommendations in respect of Band I television and medium-wave, long-wave and short-wave broadcasting have also been completely reviewed and amended where necessary to reflect technical developments. Copies of this publication may be obtained from the British Standards Institution, Sales Branch, 2 Park Street, London W.1. Price 12s. 6d. (Postage will be charged extra to non-subscribers.)

MIDAS magnetic data recording equipment, manufactured by Royston Instruments Ltd. has been installed in the SC.1 VTOL aircraft to measure vibration levels in the aircraft. The equipment consists of a power unit mounted in the forward lower equipment bay and a recorder and selector switch mounted in the aft lower equipment bay. There are approximately 30 vibration pick-up points on the aircraft, six points being recorded simultaneously. There is also a speech channel enabling test conditions to be recorded at the same time. A six-way switch is provided which gives five 10sec runs and a 10sec blank per minute.

A Marine Radar Simulator which is being designed and supplied by Ultra Electronics Ltd for the Nautical Department of the Liverpool College of Technology, will be the first installation of its type in a civilian nautical college in this country. The equipment has been developed specially to suit the needs of the College and provides full facilities for training deck officers in the interpretation of radar display information, without the expense and hazards of training under operative conditions at sea. A feature of the Liverpool installation is that it includes two 'own' ship units. Provision is also made for four target ships, true motion presentation and for coastal mapping and special effects. Three display units, Kelvin-Hughes Type 14/12P, are also being supplied.

The Postmaster General has approved, in principle, the second stage of the BBC's plans for extending and improving the coverage of the television service and of the three sound services on v.h.f. by building additional low power satellite stations in various parts of the country. The stations in this second group are to be in the following areas:

Satellite stations for both television and v.h.f. sound.

- Forfar, Angus
- Grantown-on-Spey
- Lewis
- Pitlochry/Aberfeldy
- Shetland
- Skye

Satellite stations for television.

- Caernarvon
- Hastings
- Scarborough
- Swindon

Satellite stations for v.h.f. sound.

- East Lincolnshire
- Enniskillen (for which a television station is being provided under Stage 1)
- Pembroke/Milford Haven (for which a television station is being provided under Stage 1)
- Sheffield (for which a television station is being provided under Stage 1)
- South-West Scotland.

Mr. J. J. Hill, a member of the research staff at the National Physical Laboratory, has accepted an invitation from the USSR Academy of Sciences for a fortnight's lecture tour in the Soviet Union. The subjects of his lectures will be 'Precise audio-frequency power measurements' and 'The design aspects of wide range precision transformers.' Mr. Hill will also visit the Leningrad and Moscow Institute of Electromechanics, the Polytechnics Institute and the Institute of Automation and Remote Control.

An Analogue Computer designed by the Reactor and Computer Division of Miles Electronics Ltd, Shoreham Airport, has recently been delivered to G.F.K.F. (Society for Promotion of Nuclear Research), Dusseldorf, Germany. It will be used for reactor operator training and for testing reactor control gear and computation. The installation simulates the complete reactor closed loop with the exception of heat transfer which is programmed on an associated PACE General Purpose Computer, depending on reactor type simulated and problems under investigation. The installation thus combines special purpose equipment, designed for nuclear applications and general purpose computing units. It incorporates a kinetic unit and a unit simulating six groups of delayed photo-neutrons is included. Both these units incorporate: automatic resealing over 10 decades, real or slow time operation, variable prompt neutron lifetime to suit 'slow' or 'fast' reactors, variable delayed neutron yields, choice of initial conditions over 2 decades, facility for step reactivity inputs, source for use in start up problems. The logarithmic power is recorded continually over 10 decades and reactor period indicated on a period meter. Xenon poisoning is computed in real time or speeded up on two servo integrators.

A new subsidiary company M.C.P. Electronics Ltd has recently been formed under the joint management of Mining and Chemical Products Ltd of Alperton, Middlesex, and the semiconductor division of Compagnie Française Thomson-Houston. Compagnie Française Thomson-Houston are manufacturers of transistors, diodes, and rectifiers on the Continent while Mining and Chemical Products Ltd, specialize in the refining of the rare metals and semi-conducting materials.

The National Institute of Agricultural Engineering held its Open Days on May 11 to 12 at Wrest Park, Silsoe, Beds, when exhibits were on view demonstrating the progress in the mechanization of agriculture. Among the electronic exhibits was an analogue computer used for the study of a tractor suspension system.

Eitel-McCullough, Inc., manufacturer of Eimac electron-power tubes, have formed a new subsidiary at 26 Rue de Mont Blanc, Geneva, Switzerland. The corporation known as Eitel-McCullough, S.A., will be responsible for sales, services, patents and licensing in connexion with the promotion and sale of Eimac products in Europe.

An order worth £250 000 for an Orion Electronic Digital Computer has been placed with Ferranti Ltd by the National Institute for Research in Nuclear Science. The computer will be installed at the Institute's Rutherford High Energy Laboratory at Harwell, Berkshire. Equipped with magnetic tape units and high speed printers, Orion will be used by the Laboratory for analysis and computation of data obtained from NIMROD, the 7GeV Proton Synchroton which is now under construction at Harwell.

Correction

In the article 'A Simple Multi-Dimensional C.R.T. Display Unit' in the June issue we wish to make the following amendments: Page 344, in the third line below equation (2) read Y' for y' and in fifth line $-\sin\theta$. Page 345, caption to Fig. 1 (a) The view along Y axis, and (c) The view of (b) along Z axis. Page 346, left-hand column, in second line from bottom read (equation (6)). Page 347, third paragraph from end, in sixth line read x' for x''.

Correction

In the article entitled 'A Voltage-Modulated Variable Pulse Rate Generator' by E. J. C. Fowell and A. Cowley in the May issue, the phantastron valves V_1 and V_6 in Fig. 7 should be CV4064 type.

PUBLICATIONS RECEIVED

F.B.I. REGISTER OF BRITISH MANUFACTURERS 1960, thirty-second edition, is published by Kelly's Directories Limited, and Iliffe & Sons Limited, price 42s. post free. This register contains lists of the products and services of over 7 500 member firms under more than 5 400 alphabetical headings. In addition to the Classified Buyer's Guide there are seven other sections in the Register, giving addresses of companies and firms, and information about trade associations, proprietary names, trade marks, etc. The French, German and Spanish glossaries give translations of every product term used in the main buyers' guide, each being numbered for easy reference between the English headings and their translations.

FIRST COURSE IN WIRELESS by 'Decibel', fourth edition, has been published by Sir Isaac Pitman & Sons Ltd, price 12s. 6d. In the light of developments in wireless communication, radar, and television in recent years, this book has been completely revised and brought up to date. The minimum of mathematics have been employed, resulting in a book which is free from complications and can be followed by all readers. A chapter has been added in this edition to cover developments in crystal diodes and transistors.

COMMUNICATIONS by C. A. Marshall is published by Phoenix House Ltd, 38 William IV Street, Charing Cross, London W.C.2, price 9s. 6d. This book is mainly written for young people and describes the basic principles of the many different means of radio communication, namely, by telephone, cable, radar, radio, teleprinter, telephoto, television and radio telephone, and reasons are given why one system is chosen instead of another for a particular purpose.

MICROWAVE INSTRUMENTS AND SPECIAL COMPONENTS is a loose leaf catalogue constituting a complete record of the products and activities of Decca Radar Limited in this field. From time to time additional and revised sheets will be issued to keep the catalogue up to date. Decca Radar Limited, Microwave Sales Division, Decca House, 9 Albert Embankment, London S.E.11.

NEWMARKET TRANSISTORS APPLICATION NOTES is a loose-leaf booklet covering the range of Goltop transistors. The complete range includes a.f., r.f., Switching, i.p. (Intermediate Power), Power, n.p (Noodle Power) and v.h.f. Drift. Newmarket Transistors Ltd, Exning Road, Newmarket, Suffolk.

RELAYS AND ELECTRICAL AUTOMATIC CONTROL APPARATUS is a pocket sized catalogue of the range of products manufactured by Londex Ltd, Anerley Works, 207 Anerley Road, London S.E.20.

B. & R. RELAYS' NEW CATALOGUE is designed to provide a comprehensive guide to the range of relays produced by this firm, with technical specifications and notes on the application of the various types. New relay data sheets can easily be added to this loose leaf catalogue as they are produced. B. & R. Relays Limited, Temple Fields, Harlow, Essex.

SONDES PLACE RESEARCH INSTITUTE of Dorking, Surrey, have recently produced a brochure giving an up-to-date description of the facilities offered for sponsored research and development in chemistry, chemical engineering, mechanical engineering and related subjects. It provides a fully integrated range of services covering technical and economic assessment of new projects; laboratory research; pilot plant development; and the design and prototyping of full-scale plant.

METHODS OF DETERMINATION OF RESISTIVITY OF METALLIC ELECTRICAL CONDUCTOR MATERIALS (B.S. 3289:1960). This new British Standard prescribes the methods of determining resistance and resistivity to be used in reference and routine tests on metallic conductor materials. The accuracy of the reference method is as high as can be expected under laboratory conditions, and that of the routine method sufficient for most practical purposes. It is intended that the reference method should be used in any case of dispute. An appendix to the 20-page standard explains the use of various temperature coefficients of resistance and resistivity. British Standards Institution, Sales Branch, 2 Park Street, London W.1. Price 5s. (Postage extra to non-subscribers.)

LETTERS TO THE EDITOR

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

Economy in the Series Stabilizer

Dear Sir,—The dual series section stabilizer described by Collins and Pearce in their article 'Economy in the Series Stabilizer' in your February issue represents one very interesting method of controlling large amounts of power with transistors which are themselves only capable of handling a fraction of the total controlled power.

There is, however, another circuit[1] which has found much application in these laboratories and which has very similar properties to those of the circuit described by Collins and Pearce. This circuit is shown in Fig. 1.

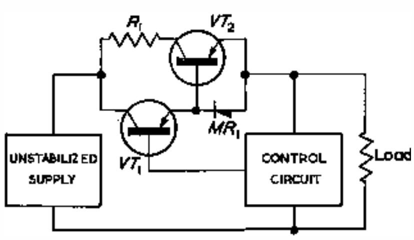


Fig. 1. Basic arrangement of transistor series stabilizer

This circuit is also based on the fact that a resistive shunt will enable a transistor to handle up to four times the voltage for a given current as compared with the conventional series stabilizer. As demonstrated by Collins and Pearce, however, the simple resistive shunt is not by itself suitable if the currents drawn by the load are likely to vary over wide limits. A means is required for removing the shunt resistor from the circuit at low currents and this has been achieved by the provision of a second transistor in series with the shunt resistor.

In the circuit of Fig. 1 MR_1 is a non-linear resistance, in fact, a silicon rectifier used to obtain a base-emitter voltage on VT_2 which is adequate to ensure that VT_2 is firmly bottomed under normal conditions. Under light load conditions the control circuit will tend to cut off transistor VT_1, and the current flowing through it will be insufficient to keep transistor VT_2 bottomed. The effective resistance of transistor VT_2 is thus increased to whatever value may be found necessary to achieve proper regulation.

As in the Collins and Pearce circuit both transistors may be subjected to their maximum dissipation, but the conditions necessary for maximum dissipation in one transistor result in the dissipation of the other transistor tending to zero. Thus one heat sink may be shared by both transistors, but it must be noted that the two collectors are not commoned.

The maximum value of the resistance R_1 is readily determined when the maximum permissible dissipation in the transistors has been fixed by the temperature conditions. For a 50°C heat sink temperature a GET571 transistor will dissipate 16W, and at a maximum load current of 10A the value of the resistor is determined by $R = (4W/I_{max}^2) = 0{\cdot}64\Omega$.

If the voltage drop across MR_1 is ignored, the maximum power in VT_1 will occur at full load current and one half the maximum permissible voltage. Similarly, the maximum power in VT_2 will occur when the voltage drop across the element is at its maximum value and the load current is one half of its maximum value.

With the circuit shown in Fig. 1 the voltage drop across MR_1 will reduce the voltage acting on VT_1, and in consequence VT_1 can never dissipate as much power as VT_2 since the value of the resistance R will then be determined by the conditions for maximum power in VT_2. Moreover the total range of control will be reduced by the amount of the forward voltage drop across MR_1. These disadvantages may be overcome by the use of the circuit[2] shown in Fig. 2 in which a second rectifier MR_2 has been added in series with VT_2. With this circuit the voltage operating on VT_2 will be reduced by the same amount as the voltage on VT_1, and in consequence the total voltage range over which the regulator element will function will be the theoretical maximum of I_{max} R less the bottomed voltage of one transistor. Using transistors of the GET571-573 type, this voltage can be held to a figure of about 0·2V if sufficient base drive is available.

Comparing the stabilizer of Fig. 2 with that of the dual section series stabilizer of Collins and Pearce it would seem that there is little to choose between the two circuits. The circuit of Fig. 2 requires two high current rectifiers, but since

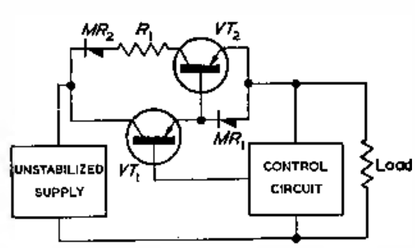


Fig. 2. Basic arrangement of dual series section stabilizer

only the forward characteristics of these rectifiers are used they can be allowed to run at a greater dissipation than would be permissible were they used as true rectifiers. Moreover, since their reverse characteristic is not used they can be the cheapest of all production rectifiers of their type. The dual section stabilizer of Collins and Pearce requires the provision of a stable source of voltage which is capable of supplying relatively large base currents if the shunted transistor is to be bottomed. Alternatively, it may be possible to use a compounded emitter-follower shunting the resistor. These complications may result in a more costly circuit than those shown in Figs. 1 and 2. It should be noted that the circuit of Fig. 2 would only normally be employed when the forward voltage drop across MR_1 becomes an appreciable fraction of the total range of voltage for which the stabilizing element is designed. If the voltage drop across MR_1 is negligible by comparison with the total voltage across the element the simpler and cheaper circuit of Fig. 1 would be preferred.

A number of regulator elements of the types described in this letter have been made. Some of these will permit changes in the unstabilized voltage of up to 14V while carrying any currents up to 15A.

Yours faithfully,

J. W. McPherson,

Applied Electronics Laboratories,
The General Electric Company Ltd,
Stanmore, Middx.

REFERENCES

1. British Patent Application No. 29539/59.
2. British Patent Application No. 36818/59.

The Authors' reply:

Dear Sir,—We are indebted to Mr. McPherson for details of his stabilizing circuit and we make the following general comments.

While we agree that it provides similar advantages to our circuit we feel that the basic approach is one of considering excess current rather than excess voltage. This leads to the possible disadvantage that at full load current the series control element is shunted by R_1 and thus its performance is degraded.

With reference to economy we feel that if one is designing a unit of reasonable performance an extra negative rail will have already been required, for other purposes, from which the reference potential may readily be obtained. A requirement for compounding the buffer section may call for one or two low power germanium transistors which are typically cheaper than one or two high current silicon rectifiers.

In conclusion, the use of either configuration must be determined by design requirements and under certain conditions one circuit may have very definite advantages over the other.

Yours faithfully,

D. J. Collins and J. R. Pearce,

Solartron Research and
Development Ltd,
Dorking, Surrey.

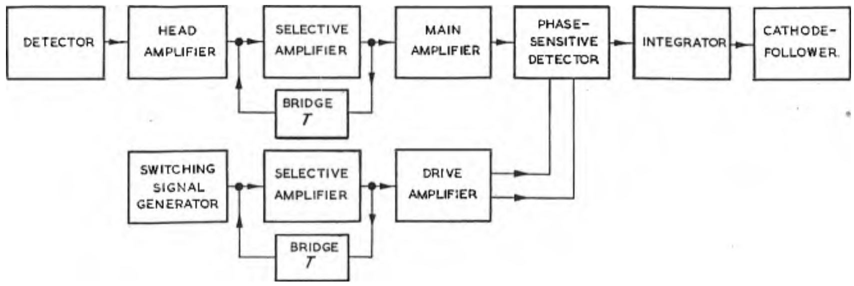


Fig. 1. Phase sensitive detector

Phase Sensitive Detection in the Presence of Noise

DEAR SIR,—With reference to the letters of Good, and Boyd and Thompson in your October 1959 and March 1960 issues, I was concerned a few years ago[1] with a similar problem to that discussed by your correspondents. In this case it was necessary to detect photo-electrically very small amounts of back-scattered 100c/s modulated light of known frequency and phase against a background of full daylight, and the signal was in consequence very small compared to the shot noise produced by daylight on the photocell.

We found that the best results were obtained by the use of an a.c. selective amplifier with cathode-follower output in both signal line and switching signal line to the phase-sensitive detector (Fig. 1). This gave the advantage of compensating for the phase shift produced in the selective amplifier by small changes in frequency (we were using the a.c. mains for operating our light source), and enabled us to obtain much reduced gain variations with frequency.

The total amplifier gain of the system was 120dB at 100c/s with a phase sensitive detector integrator bandwidth of 0·0625c/s.

Yours faithfully,

M. H. WESTBROOK,

Associated Engineering Ltd,
Leamington Spa, Warks.

REFERENCE

1. NICHOLS, J. W., WESTBROOK, M. H. An Automatic Fog Warning Instrument. *Electronic Engng.* *29*, 522 (1957).

The Characteristics and Applications of Zener Diodes

DEAR SIR,—I would like to make a comment about the Appendix to the article by Mr. J. A. Chandler in the February 1960 issue.

Equation (19) should read:

$$I_{1(\max)} - I_{L(\min)} = I_{D(\max)} \qquad -(19)'$$

and (20) should read:

$$I_{1(\min)} - I_{L(\max)} = I_{D(\min)} \qquad -(20)'$$

for the equal and inequalities to be valid, these equations should read:

$$I_{1(\max)} - I_L \leqslant I_{D(\max)} \qquad -(19)$$

and

$$I_{1(\min)} - I_L \geqslant I_{D(\min)} \qquad -(20)$$

Further, the following derivation of the final result of this appendix, which I believe to be more straightforward than that presented in the article, may be of interest.

$$\left.\begin{aligned} I_{1(\max)} &= \frac{V_1(1+k) - V_D}{R_1} \\ \text{and} \\ I_{1(\min)} &= \frac{V_1(1-k) - V_D}{R_1} \end{aligned}\right\}$$

assuming V_D is constant whence from equations (19)′ and (20)′ of this letter,

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

and

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

adding (α) and (β)

$$\frac{2(V_1 - V_D)}{R_1} = I_{D(\max)} + I_{D(\min)} + I_{L(\min)} + I_{L(\max)} \qquad -(\gamma)$$

and since $V_1 > V_D$,

$$k(2V_1/R_1) > k(I_{D(\max)} + I_{D(\min)} + I_{L(\min)} + I_{L(\max)}) \qquad -(\delta)$$

Subtracting (α) and β

$$k(2V_1/R_1) = I_{D(\max)} - I_{D(\min)} + I_{L(\min)} - I_{L(\max)} \qquad -(\epsilon)$$

(δ) and (ϵ) now give

$$I_{D(\max)} - I_{D(\min)} + I_{L(\min)} - I_{L(\max)} > k(I_{D(\max)} + I_{D(\min)} + I_{L(\min)} + I_{L(\max)})$$

i.e.

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

which is inequality (24) of the appendix and equation (3) of the text.

Equation (4) of the text follows directly by dividing $k \times (\gamma)$ by (ϵ)

i.e.,

$$V_D = V_1 \left(1 - k\left(\frac{I_{D(\max)} + I_{D(\min)} + I_{L(\min)} + I_{L(\max)}}{I_{D(\max)} - I_{D(\min)} + I_{L(\max)} - I_{L(\min)}}\right)\right.$$

and it should be noted this is an equality. Equations (5) follow directly from (α) and (β),

i.e.

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

again the absence of the inequalities should be noted.

Yours faithfully,

K. G. NICHOLS,

North Herts Technical College,
Letchworth, Herts.

The Author replies:

DEAR SIR,—A copy of your comment on the Appendix to my article 'The Characteristics and Application of Zener Diodes' in the February issue of ELECTRONIC ENGINEERING has been forwarded to me.

I think the difference in our analyses has arisen from the symbols $I_{L(\max)}$ and $I_{L(\min)}$. In equations (19) and (20) and the analysis following, I have used these to mean maximum/minimum *required* load current, whereas Mr. Nichols has taken them as the maximum/minimum *possible* load current. The possible values are given by the limiting values of my equations (i.e. the equalities) and then these agree with Mr. Nichols' equations.

On page 82 of my article in the paragraph relating to Fig. 11(c), I have used $I_{L(\max)}$ as the required value, but then used the symbol differently in the following paragraph, which may be where the misunderstanding has arisen.

Yours faithfully,

J. A. CHANDLER,

A.E.I. (Rugby) Ltd,
Rugby, Warks.

'Stroboflash'

DEAR SIR,—The article on page 294 of the May issue of ELECTRONIC ENGINEERING uses the word 'Stroboflash' both in the title and in the text.

The word 'Stroboflash' is our own registered trade mark No. 765176, the registration being dated 2 May, 1957, and the registration being in respect of "Flashlight apparatus for measuring the speed of rotating, reciprocating and vibrating bodies and for observing the operation of such bodies"; in other words, apparatus similar to that discussed in this article.

In the circumstances I shall be glad if you will kindly mention in the next issue of ELECTRONIC ENGINEERING that the word was used in error and is, in fact, our registered trade mark.

Yours faithfully,

F. W. DAWE,

Dawe Instruments Limited,
London W.5.

Regulated Power Supplies

DEAR SIR,—Since R. D. Middlebrook[1] draws attention to the feeding of the amplifier valve in a series-control regulator from a point tied to the output terminal through a reference tube as something new, originated by him recently[2] in connexion with transistor circuits, it may be needful to mention that it goes back at least as far as 1943[3]. Its merits have been analysed and emphasized[4], and it is included as a standard method in the books on stabilized power supplies by Patchett and Benson, as well as in more general literature.

Yours faithfully,

M. G. SCROGGIE,

Bromley, Kent.

REFERENCES

1. MIDDLEBROOK, R. D. Regulated Power Supplies (Letters to the Editor). *Electronic Engng.* *32*, 116 (1960).
2. MIDDLEBROOK, R. D. Design of Transistor Regulated Power Supplies. *Proc. Instn. Radio Engrs.* *45*, 1502 (1957).
3. HOGG, F. L. Electronic Voltage Regulators. *Wireless World* *49*, 329 (Fig. 3) (1943).
4. SCROGGIE, M. G. Design of Stabilized Power Units. *Wireless World* *54*, 377 (Fig. 6) (1948).

ELECTRONIC EQUIPMENT

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

(Voir page 457 pour la traduction en français: Deutsche Übersetzung Seite 462)

A PRECISION PULSE SOURCE

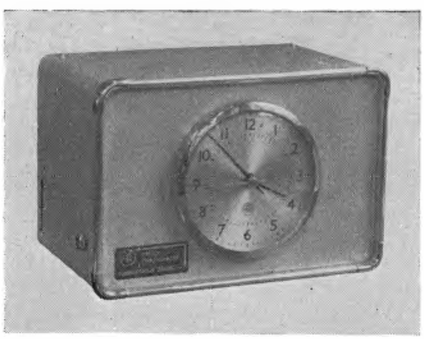
Communication Systems Ltd, Norfolk House, Norfolk Street, London W.C.2

(Illustrated below)

This all-electronic precision pulse source has a time interval accuracy of the order of a few seconds per annum, and is contained within a physically small unit.

The chronometer employs a crystal-controlled oscillator to generate a basic frequency which is fed to a series of binary dividers. The outputs of the binary dividers are fed to separate pulsing relay circuits to produce pulses at the required intervals. The special circuits are completely transistorized and make extensive use of printed circuit panels of the plug-in type.

The standard chronometer for general application produces three square wave



pulses at intervals of 1, 30 and 60 seconds respectively, but other pulse output rates can be provided. For example, for instal lation on board a research ship the chronometer is being arranged to provide outputs of 50p/s for power generator synchronization, and 1kp/s for oscilloscope time-base triggering, in addition to the 1, 30 and 60 second pulses for master time standard use. Similar facilities but based on either sidereal or universal time, at will, are being provided for use at the Jodrell Bank radio telescope. The chronometer can also provide a 20msec marker for use in electrocardiographic study. This can be applied simultaneously and in time phase to a large number of recording instruments.

Frequency stability is achieved by operating the crystal over a particular part of its characteristic. This compensates automatically for fluctuations in the crystal frequency due to normal daily temperature changes. Where the daily temperature variation is excessive, or wide differences in temperature are encountered from day to day, as in airborne or shipborne installations, the chronometer employs a thermally compensated crystal oscillator. This operates under a new principle, requiring neither heaters nor thermostats and, in consequence, consuming no extra power. It tolerates a greater fluctuation of temperature for a given time accuracy target. Alternatively an appreciably better time stability would be possible in more constant ambient temperatures.

The accuracy of the chronometer is not affected by vibration or movement. The chronometer is therefore particularly suitable for mobile installations such as military and outside broadcasting units, trains, ships and aircraft. The standard models operate on 24V d.c. but chronometers for mobile use can be produced to operate on other voltages.

EE 23 751 *for further details*

KLYSTRON POWER UNIT

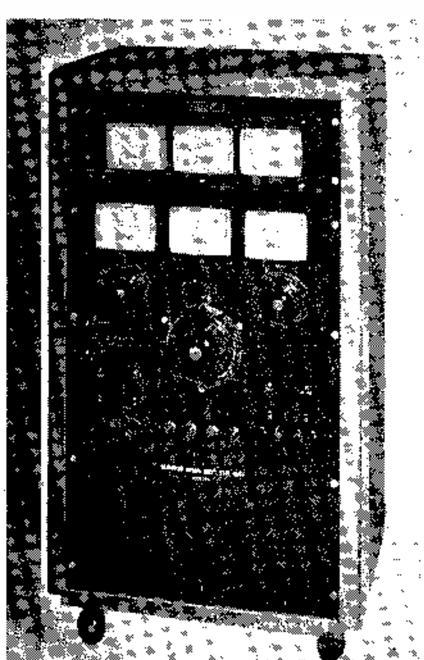
Elliott Brothers (London) Ltd, Valve Division, Borehamwood, Hertfordshire

(Illustrated below)

This unit has been specifically designed for the operation of high power floating drift tube klystrons such as the Elliott types 8FK1, 8TFK2 and 12FK1. A specially designed high speed overload circuit is provided to protect the klystron in the event of internal spark-over.

The equipment provides all the power supplies required for the operation of high voltage klystrons. The circuits are housed in a castor-mounted console on the front panel of which are all controls, monitoring meters and output sockets.

The heater supply can be varied continuously between 0 and 10V d.c. and will provide currents up to 5A. Both output voltage and current are displayed on meters mounted on the front panel and the supply is insulated to work up to 7kV above earth potential.



The focusing electrode supply can be varied continuously between 0 and 1kV d.c. and will provide currents up to 10mA. The output voltage is displayed on a meter mounted on the front panel. This supply is also insulated to withstand up to 7kV.

A composite collector and resonator supply is provided which can be varied continuously between 0 and 5kV d.c. at up to 150mA. Meters are provided to monitor both resonator and collector currents.

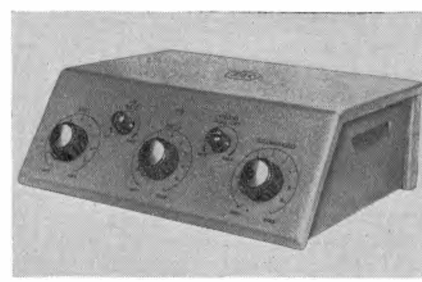
EE 23 752 *for further details*

TRANSISTOR AMPLIFIER

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

(Illustrated below)

This transistor amplifier, type BCS2429, is transportable and provides 15W of



audio power, allowing up to three large projector horn loudspeakers or between ten and twenty smaller cabinet speakers to be operated at good volume with high reproduction quality. It operates from any 12-14V battery capable of supplying 2·5A peak. The average consumption is 1A.

The use of G.E.C. transistors has enabled new construction and layout methods to be used. The main frame of the amplifier consists of a light weight webbed metal casting which forms the heat sink for the power transistors and carries the main wound components. This construction gives high resistance to mechanical shock or damage. The sloping control panel accommodates the three fader connected volume controls, the pre-set tone controls and an indicator lamp, and the unit is finished in silver blue and dark grey hammertone.

FREQUENCY RESPONSE

Microphone stages: level within ±1·5dB from 80c/s to 15kc/s. (Microphone 1 input with bass pre-set at maximum.) Gram/phono stage corrected to give substantially level response from 80c/s to 15kc/s (phono roll-off control at minimum) when used with average crystal pick-up.

POWER OUTPUT

R.M.S. sine wave measurement into a 15Ω dummy load connected to the 15Ω output terminals at 1kc/s.

With 14V d.c. supply, 18W with 5 per cent total harmonic distortion.
With 12V d.c. supply, 14W with 5 per cent total harmonic distortion.
With 14V d.c. supply, 19W with 10 per cent total harmonic distortion.
With 12V d.c. supply, 15W with 10 per cent total harmonic distortion.

INPUT SENSITIVITY

To provide full output:
Microphone 1 or microphone 2 inputs: 500μV. Gram input: 100mV.
Signal-to-Noise Ratios.
All controls at minimum, -60dB.
Gram. volume control at maximum, -60dB.
Microphone 1 or microphone 2 volume controls at maximum, -50dB.

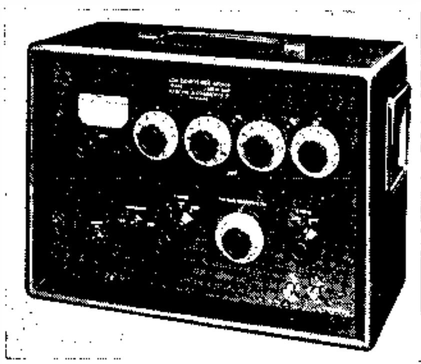
EE 23 753 *for further details*

LOW-CAPACITANCE BRIDGE

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated below)

This low-capacitance bridge, type TF 1342, will measure capacitances from 1 111pF down to 0·002pF with 0·2 per



cent accuracy by means of a three-terminal transformer ratio-arm bridge. Its exceptional discrimination and stability make it suitable for such applications as the measurement of the temperature coefficient of capacitors or changes in valve interelectrode capacitance.

The bridge measures the capacitance between any two terminals of a three-terminal network and is virtually unaffected by the impedance between either of these terminals and the third point. As a result, connexion to the component under test can be made via long leads without affecting the measurement accuracy. This feature allows *in situ* measurement of remote or wired-in components without the need to disconnect associated circuits, and is particularly valuable where it is important to include inherent shunt and stray capacitances which cannot be 'disconnected' for conventional measurement.

The resistive and reactive components of the unknown capacitor are separately balanced out in a transformer ratio-arm bridge, the unknown and standard impedances being connected in opposite sides of the bridge.

Decade switching and read-out are provided; automatic decimal point indication ensures that the read-out is direct reading. Effective shunt resistance of the component under test is balanced by a standard resistor fed via three switched transformer tappings and a continuously variable potentiometer. The total range is from 1MΩ to at least 1 000MΩ.

The bridge is complete with internal 1kc/s oscillator and detector. The detector gain can be switched to provide high or low sensitivity and, to facilitate balancing, automatic gain control is used to give a progressive increase in sensitivity as the balance point is approached.

EE 23 754 *for further details*

EVENT RECORDER

A.E.I. Radio and Electronic Components Division, 155 Charing Cross Road, London W.C.2

(Illustrated below)

An efficient and reliable record of the date and time at which electrical equipment is switched on and off, and of the period of time for which it operates, is provided by this instrument.

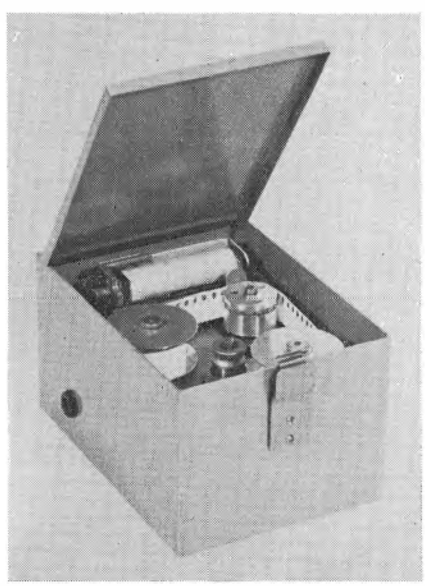
Because its drive operates independently of any supply it can also be used to record the time during which a machine is shut down through mains failure.

It incorporates a 15-day clockwork movement, with an accuracy of ± 5min throughout the 15-day cycle. This movement drives through a sprocket a recording chart at a speed of 1in/h. The chart, 31ft long and ⅝in wide, is copper sensitive Chromograph paper, perforated on one edge for driving purposes. It is calibrated according to the service required by the user.

To ensure ease of paper flow from one spool to the other, a guide device is incorporated. The spools are also designed to facilitate quick replacement of the recording chart.

The stylus is of durable copper, and is mounted on the armature of a solenoid actuated when the supply to equipment under examination is switched on. When the solenoid is operated the stylus is brought into contact with the chart.

The solenoid fitted operates at 200V d.c., 5mA, but solenoids for any supply from 50 to 250V inclusive, a.c. or d.c. can be readily supplied. Low voltage



windings can be made to special order.

An additional solenoid and stylus can be incorporated to convert the instrument to a twin channel recorder. If required, extra channels and speeds other than the standard 1in/h can be supplied.

A sapphire stylus is available and the unit can also be supplied fitted with a synchronous motor.

The recorder is available as a separate unit, in a hammer grey finish metal case with detachable, spring-loaded, self-latching lid (for which a lock is available), or without its case so that it can be incorporated in other equipment.

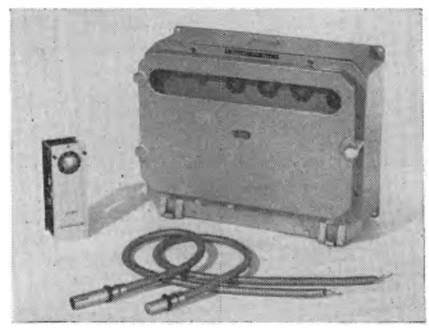
EE 23 755 *for further details*

INDUSTRIAL COUNTER

Airmec Ltd, High Wycombe, Bucks

(Illustrated below)

A new counter, type N256, designed specifically for industrial use, is now in production. Employing a versatile system of plug-in units, the equipment is fully transistorized and is housed in a robust, dust-proof case. Both counting and indication are performed by Dekatron



tubes, the maximum counting speed being well in excess of that required in normal industrial applications.

Plug-in units are available for straight counting, batch counting, input paralysis (to prevent spurious counts due to irregular shaped objects), batch signalling, and pre-batch warning. The main frame accommodates up to six plug-in units, each unit dealing with one digit or special facility; hence straight counting may be performed up to 999 999 and batch counting up to 99 999 or 9 999 if pre-batch warning is required, although these maxima are reduced by one digit if input paralysis is used. The required batch number is set on the front of each unit by means of a rotary switch.

A photocell and light source can be supplied for use with the counter, but a variety of other inputs can be accommodated, such as a closing contact, alternating voltages, etc. Dummy panels are fitted in empty unit positions for the sake of appearance.

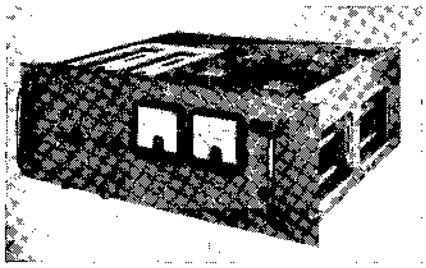
EE 23 756 *for further details*

COMPUTER REFERENCE POWER SUPPLY

Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

(Illustrated next page)

The Type AS756.2 computer reference power supply—the precision reference



voltage source in the 'Space' computer —is now generally available as an independent instrument. Two independent positive and negative outputs are provided, each capable of being loaded up to 200mA.

This reference power supply comprises two high-stability power supply sub-units with drift-corrected feedback control, using a Weston cadmium-mercury cell as a basic reference. The feedback control operates by comparing the output of the Weston cell against a fraction of the negative supply, any error voltage being amplified and applied as a correcting signal to the stabilizing control valve of the negative supply sub-unit.

The positive output is then stabilized by reference to the 'corrected' negative supply, via a summing network and a second drift-corrected amplifier.

Brief Specification:

- Outputs: +100V and −100V, 200mA on each supply.
- Accuracy and Stability: −100V better than 0·1 per cent long term. +100V can be set to within 1mV of negative supply and will remain within 5mV.
- Stabilization Ratio: Better than 1 000:1 on each supply.
- D.C. Source Impedance: Less than 0·05Ω on each supply.
- Ripple and Noise: Less than 15mV peak-to-peak.

EE 23 757 *for further details*

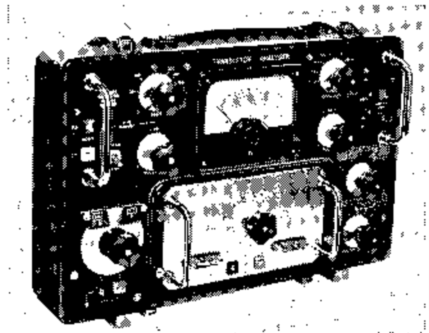
TRANSISTOR ANALYSER

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

(Illustrated below)

This analyser (type CT446) is a compact portable and battery operated tester for checking transistors, npn or pnp, of the small signal and medium power types.

The tester is suitable for taking I_b/I_c characteristics with a choice of V_c, the transistor being in the grounded emitter configuration. β values can be measured at any predetermined point on the I_b/I_c characteristic and I_{co}' is directly indicated in microamperes.



Provision has been made for the measurement of transistor noise by comparing the peak noise over the approximate range 100c/s to 10kc/s with an equivalent input signal calibrated in decibels from the internal 1kc/s oscillator.

The internal monitoring meter, 1kc/s oscillator and high gain amplifier are available at sockets on the front panel for external use.

Collector voltages 1·5V to 10·5V in five steps—automatic polarity reversal for npn and pnp.

Collector current up to 1A. Base current 0 to 1mA and 0 to 40mA. Current gain (β) 0 to 25 and 0 to 250. Noise, two ranges calibrated 1 to 20dB and 1 to 40dB.

EE 23 758 *for further details*

DIRECT-WRITING U.V. GALVANOMETER RECORDER

S.E. Laboratories Ltd, Feltham, Middlesex

(Illustrated below)

A new range of direct-writing u.v. galvanometer recorders is announced.

Models are available with chart widths up to 12in and capable of recording up to 50 channels simultaneously. A range of ultra sensitive galvanometers is available to cover frequencies from 0-100 to 0-5 000c/s. The low frequency ranges are suitable for direct connexion to thermocouples, strain gauge transducers, and other low voltage and low current signals without the use of any intermediate amplifiers. There is a choice of speed from 0·05 to 80in/sec in 20 steps. A particular feature of these recorders is the very long life of the u.v. lamp and rapid replacement from the front of the instrument in case of failure.

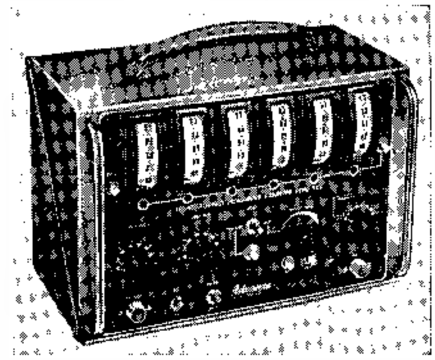
EE 23 759 *for further details*

TRANSISTORIZED COUNTER

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustrated above right)

The Advance 1Mc/s transistorized counter type TC.1 is an addition to the



new range of high quality laboratory-grade instruments now being produced by this company.

The internal frequency standard is derived from an oven controlled 1Mc/s crystal oscillator, providing an accuracy on mains operation of ±1 part in 10^6 at 25°C, and ±5 parts in 10^6 over the range 0°C to 40°C.

Facilities are provided for direct measurement of frequency up to 1Mc/s, for measurement in a range of time units of the duration of 1 to 10 periods of the incoming waveform, for the counting of regular or random electrical pulses and for the measurement of time from 1μsec to 2 777 hours (10^7 seconds). Timing pulses, from 10^{-1} to 10^{-6}p/s in decade steps, are available as an output from the counter.

The frequency measuring period is 0·1, 1·0 or 10 seconds and the display time is from 1 to 5 seconds or hold, with manual or automatic repetition of count.

An easy-to-read six figure display is provided in the form of in-line decade type meters with vertical calibration. Self-checking facilities are incorporated for testing both counter chains and provision is made for the connexion of an external recorder. A valuable extension of the TC.1 facilities is obtained by the use of the v.l.f. convertor type CA.1, which extends the low-frequency range of the TC.1 from 10c/s virtually down to d.c. (10^{-7}c/s).

EE 23 760 *for further details*

GENERAL PURPOSE COMMUNICATION RECEIVER

Marconi's Wireless Telegraph Co. Ltd, Marconi House, Chelmsford, Essex

(Illustrated below)

The type HR120 is a double superheterodyne receiver employing a switched crystal first oscillator having 1Mc/s increments. The first i.f. amplifier is continuously variable from 1 to 2Mc/s and is ganged to the second oscillator which

covers 1·1 to 2·1Mc/s, thus giving a second i.f. of 100kc/s.

Four decade tuning controls are provided, each surmounted by a dial clearly calibrated in Mc/s ×10, Mc/s ×1, kc/s ×100 and kc/s ×1 giving a very high order of frequency setting accuracy.

Switch-selection of bandwidth is provided among the controls on the front panel, all of which are grouped for maximum ease of operation.

Preferred valve types are used throughout, and have American and 'reliable' CV equivalents.

Brief Specification:

- Services: S.S.B. and d.s.b. telephony, c.w. on/off telegraphy.
- Frequency range: 2·1-30Mc/s continuous coverage.
- Noise factor: 8dB at 30Mc/s.
- Selectivity: Four pass-bands are selected by a switch, giving 1·5, 3, 6, and 12kc/s bandwidth at −3dB.
- Image protection: 50-75dB dependent upon received frequency.
- Sensitivity: 1·5μV in series with 75Ω over the whole range for 20dB signal-to-noise ratio for c.w. (1 500 c/s pass-band).
- Oscillator stability: Crystal first oscillator 2 parts in 10^6/°C. Second oscillator 10 parts in 10^6/°C.
- Frequency setting accuracy: ±200c/s after correction at 10kc/s check points.
- Automatic gain control: Not more than 10dB change in output for 100dB increase in input.
- Aerial: Suitable for connexion to 75Ω coaxial cable.
- L.F. filter: An a.f. filter tuned to 1kc/s and having a pass-band of 100c/s can be switched into circuit when required.
- Outputs: 1W in 2·5Ω for loudspeaker. 1mW for headphones. 6mW in 600Ω line.
- Power supplies: 100-120 or 200-250V, 45-65c/s single-phase a.c.
- Power consumption: 120W.

EE 23 761 *for further details*

LOW FORCE UNITOR

Ferranti Ltd, Ferry Road, Edinburgh

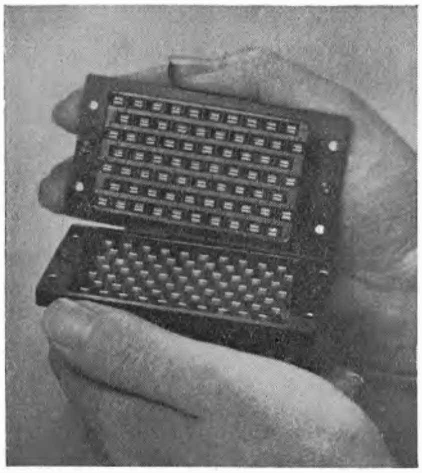
(Illustrated above right)

Now commercially available is a new low force Unitor which has been specially designed to satisfy the need for a reliable multi-pole plug and socket with a low insertion and withdrawal force.

An unusually low nominal insertion force of 4oz per pole in the Unitor is due to the unique construction of the socket clips, the jaws of which are actuated by independent springs.

An important feature of the design is the fact that the total insertion force is in direct proportion to the number of poles. The withdrawal force is approximately two-thirds of the insertion force and there is considerable tolerance to misalignment.

At present, 35 pole, 70 pole and 91 pole types of the Unitor with silver plated contacts are in production. All types are polarized.



The 70 pole type (illustrated) has successfully passed Type Approval tests as laid down in D.E.F.5321 and has received quality approval from R.C.S.C. (Humidity Classification H1; Temperature Category 40/100). It is anticipated that the other types, which are of similar construction, will be type approved in due course.

Main specification details of all silver plated types are: Maximum working voltage 500V; current rating 5A; contact resistance 2mΩ. Overall dimensions of the 70 pole type are 1·9in wide by 3·52in in length.

These Unitors are also being produced with hard gold plated contacts which are especially suitable for use in polluted atmospheres and offer superior contact performance where extremely low voltages are encountered.

EE 23 762 *for further details*

A SWEEP RESOLVER

The Plessey Co. Ltd, Ilford, Essex

(Illustrated below)

A new size 23 sweep resolver—the type 113D8H—for use in radar sweep circuits has been designed by Ketay Ltd.

The component has a flat frequency

response extending into the 100kc/s range, peaking at 500kc/s. In many applications the output of the 113D8H may be connected directly to the deflexion coils of the cathode-ray tube in the p.p.i. display system. A compensation or feedback winding is included on the stator. Voltage range is 0-30V. The operating temperature range of the new sweep resolver is from −55°C to 75°C and its weight is 28oz.

EE 23 763 *for further details*

PRECISION POTENTIOMETERS

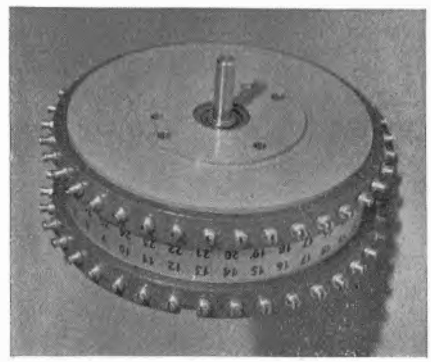
Miles Electronics Ltd, Shoreham Airport, Sussex

(Illustrated below)

A new division of Miles Electronics Ltd has been set up to produce potentiometers and other components.

Two types of precision potentiometers are available at the present time and these are the first and largest of a complete range to go into production.

The type MCD 30/FG, illustrated, is a linear potentiometer with two end connexions and 33 intermediate taps fitted as standard.



This potentiometer which has an external diameter of approximately 3in has been designed as a function generator and by shunting sections of the winding with suitable fixed resistors a very wide range of non-linear outputs can be obtained. A further advantage is that the function can readily be modified.

Twin rows of turret-type terminals are incorporated to facilitate the fitting of the fixed resistors and the terminals are identified in both clockwise and anticlockwise directions.

Multiple precious metal contacts are used for the wiper pick-off assembly and stainless steel shafts and non-corrodible ballraces are fitted as standard. Up to six units can be ganged together on a single shaft. Linear accuracies of ±0·5 per cent, ±0·25 per cent and ±0·2 per cent can be supplied.

The MCD 30/CT is a linear potentiometer of 3in diameter and the general construction is similar to the MCD 30/FG. A centre-tap and twin rows of terminals are fitted as standard and a large number of taps at any angular position can be provided. Up to six units can be ganged together on a single shaft. Linear accuracies of ±0·5 per cent, ±0·25 per cent, ±0·20 per cent and ±0·15 per cent can be supplied.

EE 23 764 *for further details*

BOOK REVIEWS

Progress in Semiconductors, Volume 4

Edited by A. F. Gibson, R. E. Burgess, F. A. Kröger. 292 pp. 121 Figs. Royal 8vo. Heywood & Company Ltd. 1960. Price 63s.

THIS is the fourth of an annual series of volumes in which invited articles by specialists in the semiconductor field are published. The aim of the series has been to present information in the form of reviews of topics of current interest to the semiconductor worker. Many of the articles published in the series to date have also contained accounts of original work by the various authors.

The first paper by Dr. H. Krömer is concerned with negative effective masses in semiconductors. Most of the paper is given over to a generalized discussion of energy contours in crystalline solids, and the concept of effective mass of charge carriers is treated in a concise and clear manner, with particularly useful diagrams of negative effective mass directions in germanium. The paper concludes with a discussion of Krömer's ideas for the negative mass amplifier. Although there are many difficulties in the way of practical achievement of this aim, the attractions of a solid state amplifier or generator working up to frequencies of the order of 10^{12}c/s are obvious.

This paper is followed by an account of oxidation processes on germanium surfaces by Dr. M. Green. The subject is treated mainly from the point of view of the surface chemistry of oxide formation; perhaps a more complete account of the surface energy states associated with oxide layers would have been welcomed by some readers.

Two papers on high field effects in semiconductors are included, one by Professor J. Yamashita on avalanche multiplication, and one by Dr. A. Chynoweth on internal field emission. At a time when these effects are exciting considerable interest in the semiconductor field, these papers are most welcome.

Two other related papers are by Dr. W. Bardsley, in which an excellent review of the electrical effects of dislocations in semiconducting crystals is given, and by Dr. Volger on dielectric properties of imperfect crystals. The latter paper is concerned mainly with dielectric loss mechanisms in ionic solids, although some information is included on covalent crystals.

The paper by Professor A. F. Ioffe and Dr. A. R. Regel, on non-crystalline semiconductors, is a most interesting introduction to a little understood and extremely difficult branch of semiconductor physics. The remaining contribution by Dr. D. Sautter on noise in semiconductors provides a much needed review of the subject, collecting together much information which to date has been available only from various widely scattered sources.

This latest volume continues the high standards set in preceding issues, but there is a slightly greater range of topics. One feels that in a series of this kind, where the articles are of a highly specialized nature, the topics should not cover too wide a range, and the connexion between the subject matter of the articles should not be too tenuous.

A. A. SHEPHERD

Mathematics for Telecommunication Engineers

By S. J. Cotton. 239 pp. 91 Figs. Royal 8vo. Chapman & Hall Ltd. 1959. Price 37s. 6d.

THE title of this book suggests an extremely narrow, specialized interest and conjures up an endless vista of similar texts for dynamicists, power engineers, structural engineers, metallurgists, chemical engineers, mining engineers, etc. Is there a genuine need for such restricted texts? Perhaps there will be a big enough demand for them as the growing number of engineers and technologists swells into a flood!

Despite the title however, the treatments given throughout this book are fairly general and conventional; in fact it reads far more like a book written by a mathematician for mathematicians. Only in the numerous examples and applications has the author drawn on the field of telecommunications. Apart from the examples, the mathematical content and treatment would serve a wider demand.

Mr. Cotton is a Senior Lecturer in the Electrical Engineering Department of the Nottingham and District Technical College. He previously obtained in 1940 a first-class honours degree (London external) as a student under Professor H. Cotton in the University College of Nottingham (now the University of Nottingham); they are not related.

Mr. Cotton is clearly an engineer who is at heart a mathematician interested in telecommunications. He has had a long experience in teaching H.N.C., external degree and City and Guilds students. The book he claims is to bridge the gap between H.N.C. mathematics and the mathematics of the journals of the professional engineering institutions. He says it is particularly suitable for the later years of the C. and G. Telecommunications Courses. He has made a careful selection of matter and the book fairly adequately covers the mathematics required by I.E.E. Part III subjects, and by engineering degrees. There are twelve chapters dealing in turn with (i) revision of calculus, (ii) determinants, (iii) expansions and series, (iv) trigonometric and hyperboloc functions, (v) complex functions, (vi) definite integrals, partial differentials, (vii) differential equations, some network theory (viii) complex variable, (ix) harmonic analysis; Fourier and Laplace transforms (alas not p-multiplied), (x) vector analysis (Maxwell's equations), (xi) special functions (Bessel, Gamma), (xii) probability and statistics.

Each chapter is well illustrated with line diagrams (91 in all) and occasionally worked examples are included in the text. A good selection of exercises is placed at the end of each chapter and at the end of the book; there are altogether 357 exercises and the answers are given. The serious student will appreciate this provision.

Notable omissions from the text are (i) an introduction to non-linear theory (how can one treat the theory of oscillation without it?), (ii) two-dimensional field plotting, (iii) an introduction to electronic computation (for this is having a profound effect on mathematical requirements), (iv) statistical theory should have been taken deeper, e.g. correlation functions.

The book will undoubtedly appeal to, and well satisfy, the telecommunications students to whom it is addressed.

J. E. PARTON

Air Traffic Control

Edited by C. D. Colchester. 100 pp. 19 Figs. Medium 8vo. Marconi's Wireless Telegraph Co. Ltd, Chelmsford. 1960. Price 17s. 6d.

THIS timely little book originated as a study report on future requirements for air traffic control by a team of Marconi engineers. There are seven chapters by different authors. In his introduction Dr. E. Eastwood treats the control of air traffic as a servo-system problem regarding the flight progress strips as the objects under control, with the aircraft attempting to conform to the flight plan and the controllers undertaking error extraction and correction using ground-air-ground communications, ground radar and data displays as control linkages.

Chapter II 'Air Traffic Control Development and its Present State' describes briefly the evolution of procedural control; and lists the existing radio navigational aids, summarizes the rules of the air and gives a very sketchy account of present practice. Faults in the present system are the disproportionate relation between the clearance space and the size of the aircraft, the inaccuracies and delays in position reporting and the large amount of time taken for the controller to build up a mental picture of the traffic situation. The subsequent 70 pages describe some of the equipment being produced and under development for improving the system. Chapter III deals with fixed and aeromobile communications including a proposed automatic data link for all routine messages, an interim palliative on the North Atlantic being the existing radio teleprinter service. Chapter IV is a brief survey of present-day radio navigational aids with a discussion of future possibilities for airborne doppler supported by a gyro heading reference and supplemented by a fix-producing aid or an astro fix. There is a page on automatic landing which is a little out of place as regards a.t.c. Radar is given a dozen pages in Chapter V and is mostly 50cm MTI Surveillance Terminal Area Radar with photographs of installations and displays. For continuous radar coverage along the airways 10cm radars will be added at certain sites where gaps in the 50cm cover appear unavoidable, although they cannot give such good all-weather performance.

A most useful Chapter VI 'Using the Information', is by R. P. Shipway who shows how a modern electronic data handling system can ease the burden of air traffic control although the responsibility and decision making remain with the controller.

Altogether the book is a gallant attempt to cover a tremendous field of technological advance and focus attention towards the solution of a difficult problem. Written mainly for the non-technical reader it steers a reasonably comfortable course between the Skylla of elaborate detail and the Charybdis of over-simplification.

L. J. BRAYBROOK

Encyclopaedic Dictionary of Electronics and Nuclear Engineering

By R. I. Sarbacher. 1417 pp. 1424 Figs. Crown 4to. Sir Isaac Pitman & Sons Ltd. 1960. Price 160s.

THIS encyclopaedic dictionary, a comprehensive reference work in electronics and nuclear engineering, is designed for the scientist, technician or student working in these related fields. The problem of communication is especially acute in those technical projects where a team of specialists representing a wide variety of engineering fields is working together. The author has endeavoured to supply an aid to such communication in electronics, nucleonics and nuclear engineering by presenting a readily usable dictionary of the specialized vocabulary which is beyond the scope of general dictionaries of the English language. Extensive cross-referencing has been provided in order that related entries may be studied and compared or contrasted. Equations, tables, design equations, statistical data, etc., are presented where relevant.

Kempe's Engineers Year Book 1960, Volumes 1 and 2

3 080 pp. Crown 8vo. 65th Edition. Morgan Brothers (Publishers) Ltd. 1960. Price 87s. 6d. per set in case

THIS edition has been revised to cover the recent advances in science and practice of engineering. There are more than 120 pages of new tables and text and the information is arranged and broadly indexed for rapid reference. Completely new sections deal with Electronic Engineering and Nuclear Power (replacing the chapter Atomic Power). There are also numerous alterations and additions due to the issue of new and revised British Standards.

The Theory of Optimum Noise Immunity

By V. A. Kotel'nikov. 140 pp. 15 Figs. Crown 4to. McGraw-Hill Publishing Co. Ltd, Inc., New York. 1960. Price 58s.

THIS book is a translation from the Russian of the author's doctoral dissertation presented in 1947 and published in 1956, and covers original work in the field of information theory, some of which antedates corresponding work carried out in this country and the U.S.A.

The work is concerned with the steps which can be taken at the receiver to increase the intelligibility of signals in the presence of 'white' noise. The term used to express this increase is 'noise immunity'; it is a generic term, the actual meaning depending on the communication system and the type of signal.

A short first part is devoted to the mathematical basis of the probability theory used later. Part 2 is concerned with the reception of telegraph signals, either amplitude or frequency keyed. The case of many levels is covered, including that of 32 frequencies, one for each symbol transmitted. Part 3 considers signals giving the instantaneous values of a parameter at separated time intervals, as in telemetry. The use of amplitude, pulse and frequency modulation are separately covered, and the effects on them of weak and strong noise. In Part 4 the corresponding cases for the transmission of waveforms as modulation are considered.

The subject is, by its nature, one that must be treated mathematically, but every attempt is made to give a physical picture of the subjects, including a discussion at the end of each chapter on the geometrical meaning of the results in n-dimensional space.

The translation has been excellently done, and the book forms a most worthwhile addition to the literature in its field.

W. H. ALDOUS

Principles of Electricity and Magnetism

By Y. Rocard. 767 pp. 697 Figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1960. Price 70s.

THE well-known French text *Electricité* was developed from a series of lectures given by Professor Rocard at the Sorbonne. This first English edition contains a considerable number of additions which have also been incorporated in the second French edition, published recently.

Commencing with electrostatics, Part 1 covers the groundwork clearly and concisely, taking in important recent developments. Part 2 deals similarly with magnetics. Parts 3, 4 and 5 deal respectively with electrokinetics, electrodynamics and alternating currents, while Part 6 is devoted to an account of the propagation of electromagnetic waves along lines and through space. Part 7 deals with the free electron and electronics and the final section is devoted to a general survey of systems of units.

Programming Programme for the BESM Computer

By A. P. Ershov. 158 pp. 21 Figs. Medium 8vo. Pergamon Press Ltd, New York, London. 1960. Price 60s.

THIS book is mainly a description of the inner workings of a scientific Autocode for the BESM, an electronic computer built by the Russian Academy of Sciences, but a brief programmer's view of the BESM is also given. It is a fast machine with a rather small store, but magnetic tape is also available.

Although the Autocode described is advanced, since it aims to construct a programme which is almost as efficient as one obtained by direct machine language coding, it does not appear to be easy to use due to the complicated way in which a problem must be presented. On the whole the Autocode is more likely to be useful to experienced programmers than to the novice. As almost the whole of the book is devoted to the methods of one particular Autocode scheme, it is likely to interest only those specialists who have to plan and programme such schemes.

Several typographical mistakes were noted and the translation does not always lead to good English. The book is reproduced from typewritten sheets with most of the mathematical symbols inserted by hand and others set up correctly in type. This has led to a very poor presentation, hardly justifiable in a book of this price.

M. J. MARCOTTY

Exploding Wires

Edited by W. G. Chace and H. K. Moore. 373 pp. 190 Figs. Medium 8vo. Plenum Press Inc., New York. Chapman & Hall Ltd. London. 1960. Price 76s.

THIS book has been published primarily to record the papers of the Conference on the Exploding Wire Phenomenon conducted by the Air Force Cambridge Research Centre in Boston, Massachusetts, on 2 and 3 April 1959. A second purpose of the book is to serve as a monograph on exploding wires.

This work will be of help to all workers in the fields of high speed photography, shock wave and thermonuclear research and also to students, technologists and scientists concerned with the fluid physics field.

Radio and Electronics, Volumes 1 and 2

Edited by J. H. Reyner. 1023 pp. 1268 Figs. Crown 4to. Sir Isaac Pitman & Sons Ltd. 1960. Price 105s. per set.

THIS two-volume work is designed to provide the student, the craftsman and the young professional engineer with a knowledge of the fundamental background to the use of radio and electronic techniques.

Volume 1 deals with the basic principles of electrical engineering, circuit design, electronic components, thermionic valves, crystal diodes and transistors, cathode-ray tubes and their associated circuits, the propagation and modulation of electric waves and very high frequency phenomena and techniques. The treatment is largely non-mathematical, but there is a comprehensive section on engineering mathematics.

Volume 2 is devoted mainly to the practical application of the principles dealt with in the first volume, but additional theoretical information is included where necessary. Among other things it includes a survey of radio transmission and reception, the special problems and requirements of marine and aeronautical equipment including navigational aids and ancillary devices, sound recording and reproducing equipment including disk, film and tape techniques. The final two sections describe the more important applications of electronics to medicine, including diathermy, biological measurements and X-rays, and to industry in the fields of measurement, counting and control, etc.

Encyclopaedia on Cathode-Ray Oscilloscopes and Their Uses

By J. F. Rider and S. D. Uslan. 1274 pp. 1039 Figs. Demy 4to. 2nd Edition. John F. Rider Publisher Inc., New York. Chapman & Hall Ltd. London. 1959. Price 210s.

THIS second edition has been greatly expanded to include all new types of oscilloscopes and their applications. It starts with cathode-ray tube construction and theory, then carries one through an analysis of modern oscilloscope circuits, commercial scope types and maintenance, to a detailed treatment of how the oscilloscope is operated for "all" applications. New sections on pulse measurements and square wave testing have been added.

Spezialröhren-Eigenschaften und Anwendungen. (Special Valves—Properties and Applications)

By F. Cubasch. 439 pp. 319 Figs. Demy 8vo. Verlag für Radio-Foto-Kinotechnik GmbH., Berlin-Borsignwalde. 1960. Price

THE 'special' valves dealt with in this book are those not ordinarily used in radio and television equipment, but travelling field valves like magnetrons and klystrons are not included. The book contains detailed information and illustrations of selected valves manufactured by Valvo GmbH. Hamburg with data sheets and characteristic curves, although the explanations and fundamentals of calculation are quite general and valid also for the products of other firms.

A brief introduction is followed by three main sections dealing with high vacuum valves, gas-filled valves and photocells. The first of these main sections comprises in its first subsection valves for amplification and measuring purposes, amongst them the so-called colour series—blue for aeroplane and ship equipment, green for electronic computers, yellow for communication services and red for industrial control equipment—valves for the u.s.w. range and special purpose valves, e.g. decadic counters and synchronizing valves. A second subsection deals with transmitting and power amplification valves. Here ample space is devoted to the calculation of transmitter operations. Cathode-ray tubes are treated in the third subsection. Hints given for taking photographs of oscillograms will be welcome.

The second main section deals with gas-filled valves comprising gas-filled rectifiers, thyratrons, ignitrons, stabilizing tubes, Geiger-Müller counters and relay tubes. Several examples are given for the calculation of l.v. gas-filled rectifier equipment while much less space is devoted to h.v. rectifiers mainly used for the operation of transmitter valves and power amplifiers. In the subsection dealing with thyratrons, special attention is paid to the methods for adjusting ignition and extinction times and to protective devices. The various connexions used for igniting ignitrons are described and special measures are given for preventing damage to be done to ignitron pins. Thermostatic control of the flow of cooling water is recommended. The application of ignitrons with single-phase and three-phase connexions for welding purposes and the operation of electric furnaces are discussed in detail and some numerical examples are given. Stabilizing valves based on glow discharge for voltage stabilization and voltage comparison are then explained, also with several numerical examples for their application. The brief subsections on Geiger-Müller counters and on relay valves are perhaps the weakest parts of the book as the treatment is too concise for ready comprehension. The book ends with a section on photocells and photomultipliers.

A special feature of the book is the recommendations for the operation of every group of valves dealt with. These contain a wealth of useful information for the designer and the user. The text is written in a fluent and readable style. A few misprints could be noticed, e.g. the root sign is apparently omitted in the equations 102 and 103 on page 179.

A comparative table of equivalent valves, a rather brief bibliography and a name and subject matter index complete this well produced volume.

R. NEUMANN

Classification and Indexing in Science

By B. C. Vickery. 225 pp. Demy 8vo. 2nd Edition. Academic Press Inc., New York. Butterworths Scientific Publications, London. 1959. Price 30s.

THE first edition of this book was written primarily as a study of method, but proved to be of help to those concerned with constructing systems of information retrieval and was even adopted as a teaching manual. This new edition has been revised and enlarged in order to serve these purposes better.

Industrial Electronics and Control

By G. Kloeffler. 533 pp. 452 Figs. Medium 8vo. 2nd Edition. John Wiley & Sons Inc., New York. Chapman & Hall Ltd. London. 1960. Price 80s.

THIS edition differs from its predecessor and from all previous texts in that it approaches the electronic theory of rectification, amplification, and oscillation through solid state theory rather than by way of the vacuum and gaseous tubes. In addition, the author examines such recent developments as solid-state thyratrons, cryotrons, and cold-cathode vacuum tubes.

Two new chapters on semiconductors have been included at the outset. The material on servomechanisms has been completely rewritten and information has been added on magnetic amplifiers, computers, and electronic measurement.

Value Engineering 1959

165 pp. 68 Figs. Medium 8vo. Interscience Publishers Inc., New York. 1960. Price $6.00.

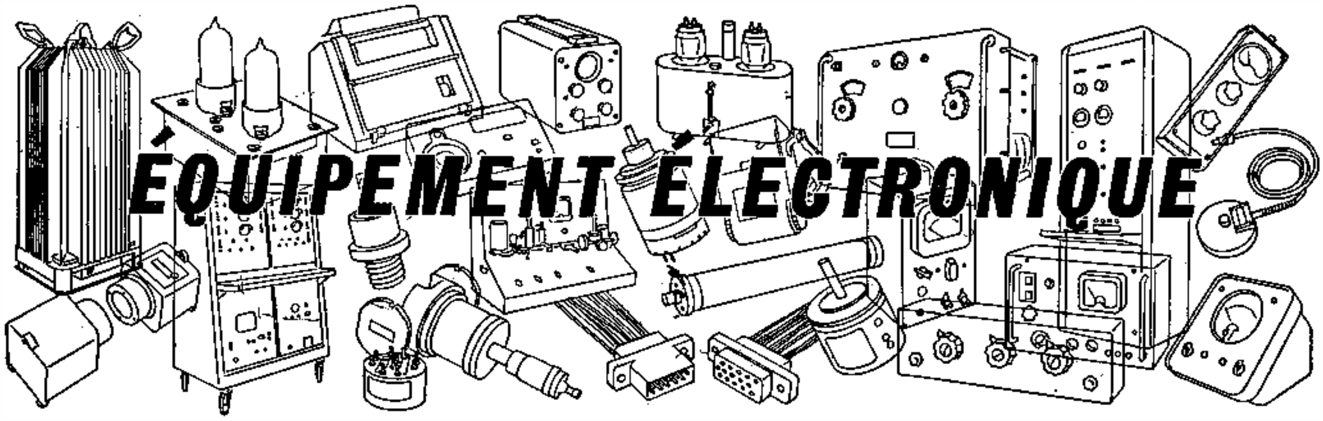
THIS book contains all the technical papers presented at the Electronic Industries Association Conference on Value Engineering held at the University of Pennsylvania on 6 and 7 October 1959. This was the first comprehensive industry conference on this subject. The papers cover basic concepts and philosophies, techniques, application, and relationship with other work areas. Customer, manufacturer, and vendor viewpoints are represented.

A Simple Approach to Electronic Computers

By E. H. W. Hersee. 101 pp. 27 Figs. Crown 8vo. Blackie & Son Ltd, Glasgow. 1959. Price 12s. 6d.

THIS book presents in some detail, and as simply as possible, a few of the basic working principles of both digital and analogue machines with the idea that anyone with a clear understanding of these elementary and fundamental ideas will begin to appreciate the essential simplicity of computers.

(Pages 457 to 466 French and German section)



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 450 à 453*

UNE SOURCE D'IMPULSIONS DE PRÉCISION

(*Illustration à la page 450*)

Communication Systems Ltd, Norfolk House, Norfolk Street, London W.C.2

Cette source d'impulsions—ou chronomètre—entièrement électronique, est à intervalle de temps d'une précision de l'ordre de quelques secondes par an et elle est contenue dans un petit boîtier.

Le chronomètre emploie un oscillateur piloté par quartz, produisant une fréquence de base qui alimente une série de diviseurs binaires. Les sorties de ces derniers alimentent des circuits de relais d'impulsions à part, afin de produire des impulsions aux intervalles voulus. Les circuits spéciaux sont complètement transistorisés et font un usage étendu de panneaux à circuits imprimés du type à prises.

Le chronomètre standard pour les applications générales produit trois impulsions d'ondes carrées à des intervalles de 1,30 et 60 secondes respectivement, mais d'autres taux de sortie d'impulsions peuvent être assurés. Par exemple, en vue de son installation à bord d'un navire de recherches, le Chronomètre a été étudié pour produire des sorties de 50 impulsions par seconde pour la synchronisation avec le générateur électrique, et 1000 impulsions par seconde pour le déclenchement de la base de temps oscilloscopique, en plus des impulsions de 1,30 et 60 secondes, à l'usage de l'étalon de temps principal. Les mêmes caractéristiques, mais basés cette fois sur le temps sidéral ou universel, au choix, seront assurées à l'intention du radio-télescope de Jodrell Bank. Le Chronomètre peut également être muni d'un marqueur de 20 msec pour l'étude électrocardiographique. Ce dernier peut être appliqué simultanément et en phase de temps à un grand nombre d'instruments d'enregistrement.

La stabilité de fréquence est realisée en pilotant le quartz au-dessus d'une certaine partie de sa caractéristique. Ceci compense automatiquement les fluctuations dans la fréquence piézoélectrique, dues aux changements de température normaux et quotidiens. Lorsque la variation quotidienne de température est excessive ou que l'on trouve de fortes différences de température d'un jour à l'autre, comme cela se produit dans les installations d'avion ou de bateaux, le Chronomètre emploie un oscillateur à cristal à compensation thermique. Cet oscillateur fonctionne selon un nouveau principe, n'exigeant ni réchauffeurs ni thermostats et, par conséquent, ne consommant pas de courant supplémentaire. Il tolère une plus grande fluctuation de température pour un objectif donné de précision de temps, ou bien une stabilité de temps nettement meilleure serait possible dans des températures ambiantes plus constantes.

La précision du chronomètre n'est pas affectée par les vibrations ou les mouvements. Le Chronomètre convient donc particulièrement aux installations mobiles telles que les unités militaires ou de radiodiffusion en extérieurs, les trains, les bateaux et les avions. Les modèles standard fonctionnent sur courant continu de 24 Volts, mais des Chronomètres pour usage mobile peuvent être fournis sur commande pour fonctionner sur d'autres tensions.

EE 23 751 *pour plus amples renseignements*

BLOC D'ALIMENTATION À KLYSTRON

(*Illustration à la page 450*)

Elliott Brothers (London) Ltd, Valve Division, Borehamwood, Hertfordshire

Cet appareil a été spécialement étudié pour les klystrons à tube de glissement flottant de grande puissance, tels que les klystrons Elliott type 8FKI, 8TFK2 et 12FK1. Un disjoncteur à maxima, à action rapide et d'une conceptionspéciale, protège le klystron en cas d'arc intérieur.

L'équipement fournit toutes les alimentations nécessaires à la mise en action de klystrons à haute tension. Les circuits sont logés dans une console montée sur étrier, dont le panneau avant porte toutes les commandes, compteurs de contrôle et prises de sortie.

L'alimentation de chauffage est à variation continue entre 0 et 10V c.c. et fournit des courants allant jusqu'à 5A. Tant la tension de sortie que le courant sont affichés sur des cadrans du panneau avant et l'alimentation est isolée pour fonctionner jusqu'à 7000V au-dessus du potentiel de terre.

On peut faire varier continûment de 0 à 1000V c.c. l'alimentation de l'électrode de focalisation, qui peut fournir des courants allant jusqu'à 10mA. La tension de sortie est affichée sur un cadran monté sur le panneau avant. Cette alimentation est également isolée pour résister jusqu'à 7000V.

Une alimentation mixte de collecteur et résonateur peut être variée continûment entre 0 et 5000V c.c. et jusqu'à 150mA. Tant le courant du résonateur que celui du collecteur sont contrôlés par des enregistreurs.

EE 23 752 *pour plus amples renseignements*

AMPLIFICATEUR À TRANSISTORS

(*Illustration à la page 450*)

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

Cet amplificateur à transistors Type BCS 2429, est portatif et fournit une puissance basse fréquence de 15W, permettant d'actionner, à un bon niveau sonore et avec une reproduction de haute qualité, jusqu'à trois grands haut-parleurs à pavillon de projection ou dix à vingt petits haut-parleurs à coffret. Il fonctionne sur n'importe quelle pile de 12 a 14V capable de fournir un courant de pointe de 2,5A, la consommation moyenne étant de 1A.

L'emploi des transistors G.E.C. a permis l'utilisation de nouvelles méthodes de fabrication et de disposition. La charpente principale de l'amplificateur se compose d'un moulage en coquille métallique nervurée de poids léger, formant la source froide des transistors de puissance et portant les principaux enroulements. Cette construction assure une haute résistance aux chocs ou à l'endommagement mécanique. Sur le panneau de commande en pente se trouvent les trois commandes de puissance et d'affaiblissement sonores, les correcteurs de tonalité préréglés et une lampe de signalisation. Le coffret est de couleur bleue argent et gris fer foncé.

RÉPONSE DE FRÉQUENCE

Etages microphoniques: niveau en deçà de ±dB, de 80 Hz à 15 Hz (Microphone 1: entrée avec basses fréquences préréglées au maximum).

Etage gramophonique corrigé pour

donner largement une réponse de niveau de 80 Hz à 15 kHz (commande de circuit éliminateur au minimum) lorsqu'il est employé avec un pick-up piézoélectrique moyen. Sortie de puissance.

Mesure d'ondes sinusoïdales efficaces dans une charge fictive de 15 ohms, reliée aux bornes de sortie de 15 ohms à 1000 Hz.

Avec une alimentation en continu de 14V: 18V avec 5% de t.d.h.

Avec une alimentation en continu de 12V: 14W avec 5% de t.d.h.

Avec une alimentation en continu de 14V: 19W avec 10% de t.d.h.

Avec une alimentation en continu de 12V: 15W avec 10% de t.d.h.

SENSIBILITE D'ENTRÉE

Pour fournir une sortie complète:

Entrées du microphone 1 ou du microphone 2: 500μV.

Entrée du gramophone: 100mV.

Rapport signal/bruit toutes les commandes étant au minimum: 60dB.

Commande de niveau sonore du gramophone au maximum: 60dB. Commande de niveau sonore du microphone 1 ou du microphone 2 au maximum: 50dB.

EE 23 753 *pour plus amples renseignements*

PONT À FAIBLE CAPACITÉ

(Illustration à la page 451)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Ce pont à faible capacité, Type TF 1342, peut mesurer des capacités de 1111 F jusqu'à 0,002$\mu\mu$F, avec un degré de précision de 0,2% au moyen d'un pont à bras de proportion de transformation à trois bornes. Sa discrimination exceptionnelle et sa stabilité le rendent approprié pour des applications telles que la mesure du coefficient de température de condensateurs ou des changements dans la capacité interélectrode de tubes.

Le pont mesure la capacité entre n'importe quelles deux bornes d'un réseau à trois bornes et il n'est virtuellement pas affecté par l'impédance entre l'une ou l'autre de ces bornes et le troisième point. Par conséquent, la connexion à la pièce soumise à l'essai peut être effectuée par de longs conducteurs sans affecter la précision de la mesure. Cet avantage permet la mesure sur place de pièces éloignées ou bobinées, sans qu'il soit nécessaire de débrancher les circuits connexes, et elle est particulièrement utile lorsqu'il est important d'inclure des capacités à parasites et dérivation inhérents qu'on ne peut 'débrancher' pour les mesures conventionnelles.

Les composantes réactives et résistives du condensateur inconnu sont équilibrées séparément dans un pont à bras de proportion de transformation, les impédances normales et inconnues étant reliées sur les côtés opposés du pont.

Le pont offre aussi la commutation de décades et la lecture. L'indication par point décimal automatique assure une lecture directe. La résistance de shunt de la composante soumise à l'essai est equilibrée par une résistance classique, alimentée à travers trois prises de transformateur commutées et un potentiomètre à variation continue. La gamme totale s'étend de 1 mégohm à au moins 1000 mégohms.

Le pont est complet avec oscillateur intérieur de 1000 Hz et détecteur. Le gain de détecteur peut être commuté pour fournir une haute ou une basse sensibilité. Pour faciliter l'équilibrage, on emploie une commande de gain automatique qui donne une augmentation progressive de sensibilité à mesure qu'on approche le point d'équilibre.

EE 23 754 *pour plus amples renseignements*

ENREGISTREUR D'ÉVÉNEMENTS

(Illustration à la page 451)

A.E.I. Radio and Electronic Components Division, 155 Charing Cross Road, London W.C.2

Un enregistrement efficace et sûr de la date et du moment où un appareillage électrique est mis en circuit ou hors circuit, ainsi que de sa durée de fonctionnement, est fourni par cet instrument.

Vu qu'il est entraîné indépendamment de toute alimentation, il peut également être utilisé pour enregistrer la durée d'arrêt d'une machine par suite d'une panne de secteur.

Il comprend un mouvement d'horlogerie d'une précision de ±5 minutes pendant tout le cycle de 15 jours. Ce mouvement entraîne, au moyen d'un galet, une bande enregistreuse, à la vitesse de 2,5 cm par heure. La bande a 9m de long et 1,5cm de large, et elle est en papier chromographique sensible au cuivre, perforé sur un des côtés pour l'entraînement. Elle est étalonnée en fonction des besoins de l'utilisateur.

Afin d'assurer l'avance régulière du papier d'une bobine à l'autre, un dispositif de guidage a été incorporé. Les bobines mêmes ont été conçues de manière à faciliter le remplacement rapide de la bande enregistreuse.

Le style est en cuivre durable et il est monté sur l'armature d'une solénoïde mise en action lorsque l'alimentation du matériel soumis à l'examen est mise en circuit. Lorsque la solénoïde est actionnée, le style est mis en contact avec la bande.

La solénoïde dont est muni l'enregistreur fonctionne à 200V cc 5 mA, mais des solénoïdes pour n'importe quelle alimentation, de 50 à 250V inclusivement, c.a. ou c.c., peuvent être fournies sans difficulté. Des enroulements à basse tension peuvent être faits sur commande spéciale.

Une solénoïde et un style supplémentaires peuvent être incorporés afin de transformer l'instrument en un enregistreur à deux canaux. Des canaux supplémentaires et des vitesses autres que la vitesse normale de 2,5 cm/heure peuvent être prévus.

L'instrument est doté d'un style à pointe de saphir et il peut également être muni d'un moteur synchrone. Il est fourni comme un élément à part soit dans un coffret gris fer, avec couvercle détachable à ressort et autoverrouillage (une serrure étant fournie à cet effet), soit sans coffret de façon à ce qu'on puisse l'incorporer à d'autres installations.

EE 23 755 *pour plus amples renseignements*

COMPTEUR INDUSTRIEL

(Illustration à la page 451)

Airmec Ltd, High Wycombe, Bucks

Un nouveau compteur, type N256, spécifiquement conçu pour l'usage industriel, est actuellement en production. Utilisant un système des plus adaptables d'éléments à fiches, l'équipement est entièrement transistorisé et logé dans un robuste coffret protégé contre la poussière. Tant le comptage que l'indication se font par tubes Dékatron, la vitesse de comptage maxima étant bien au-delà de celle requise pour les applications industrielles normales.

Les éléments interchangeables sont fournis pour le comptage direct, le comptage par lots, la paralysie d'entrée (afin d'empêcher de faux comptages dûs à des objets de forme irrégulière), la signalisation de lots et l'avertissement avant les lots. Le cadre principal englobe jusqu'à six éléments interchangeables, chaque élément traitant un seul chiffre ou caractère spécial. Le comptage direct peut donc s'effectuer jusqu'à 999 999 et le comptage par lots jusqu'à 99 999 si l'avertissement "prélots" est nécessaire, quoique ces maxima soient réduites par un chiffre si l'on emploie la paralysie d'entrée. Le chiffre de lots voulu est fixé à l'avant de chaque élément au moyen d'un commutateur rotatif.

Une source de lumière et de cellule photoélectrique peut être fournie à l'usage du compteur, mais une variété d'autres entrées peuvent être prévues, telles que le contact de fermeture, les tensions alternantes, etc. Des panneaux fictifs sont fixés dans les positions vides des éléments pour l'esthétique.

EE 23 756 *pour plus amples renseignements*

ALIMENTATION DE RÉFÉRENCE DE CALCULATEUR

(Illustration à la page 452)

Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

Ce bloc d'alimentation de référence pour calculateur électronique, Type AS756.2, constituant la source de tension étalon de précision du calculateur SPACE, est maintenant fourni comme instrument indépendant. Il comporte deux sorties indépendantes, positive et négative, chacune d'elles pouvant être chargée jusqu'à 200mA.

Il se compose de deux sous-blocs d'alimentation à haute stabilité avec commande de réaction à dérive corrigée, utilisant une cellule de cadmium-mercure Weston comme référence de base. La commande de réaction fonctionne en comparant la sortie de la cellule Weston avec une fraction de l'alimentation négative, toute erreur de tension étant amplifiée et appliquée sous forme de signal correcteur à la lampe de commande stabilisatrice du sous-bloc d'alimentation négative.

La sortie positive est ensuite stabilisée en se référant à l'alimentation négative "corrigée," par le truchement d'un réseau intégrateur et d'un second amplificateur à dérive corrigée.

Brèves caractéristiques.

Sorties:
+100V et −100V, 200mA sur chaque alimentation.

Précision et stabilité:—
100V supérieurs à 0,1% à long terms. +100V peuvent être réglés à 1mV près d'alimentation négative et resteront en deçà de 5mV.

Rapport de stabilisation:
Supérieur à 1000/1 sur chaque alimentation.

Impédance de source c.c.:
Inférieure à 0,05 ohm sur chaque alimentation.

Ondulation et bruit:
Inférieurs à 15mV de crête à crête.

EE 23 757 *pour plus amples renseignements*

ANALYSEUR DE TRANSISTORS

(Illustration à la page 452)

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

Cet analyseur Type CT446 est un appareil portatif compact, fonctionnant sur pile, pour l'essai des transistors n-p-n ou p-n-p des types à faible signal et puissance moyenne.

Il est étudié pour la prise des caractéristiques I_b/I_c avec un choix de Vc, le transistor se trouvant dans la configuration d'émetteur mis à la terre. Les valeurs B peuvent être mesurées à n'importe quel point prédéterminé de la caractéristique I_b/I_c et le I'co est indiqué directement en μA.

On peut également mesurer le bruit de transistor en comparant le bruit de pointe dans la gamme approximative de 100Hz à 10 kHz, avec un signal d'entrée équivalent étalonné en décibels à partir de l'oscillateur intérieur de 1000 Hz.

.Des raccords sur le panneau avant permettent d'affecter le contrôleur intérieur, l'oscillateur de 1000 Hz et l'amplificateur à grand gain à l'usage extérieur.

Les tensions au collecteur sont de 1,5V à 10,5V en cinq plots, et l'inversion de polarité se fait automatiquement pour n-p-n et p-n-p.

Le courant au collecteur va jusqu'à 1A. Courant de base: 0 à 1mA et 0 à 40 mA. Gain de courant (B): 0 à 25 et 0 à 250. Niveau de bruit pour deux gammes étalonnées: 1 à 20 dB et 1 à 40 dB.

EE 23 758 *pour plus amples renseignements*

ENREGISTREUR GALVANOMÉTRIQUE À LAMPE ULTRA-VIOLETTE À INSCRIPTION DIRECTE

(Illustration à la page 452)

S.E. Laboratories Ltd, Feltham, Middlesex

Cette société vient d'annoncer la réalisation d'une nouvelle gamme d'enregistreurs galvanométriques u.v. Les modèles livrables comportent des largeurs de bande d'un maxima de 30 cm. pouvant enregistrer jusqu'à 50 canaux simultanément.

Il existe une série de galvanomètres ultra-sensibles dont les gammes de fréquences vont de 0-100 à 0-5000 Hz. Les gammes de basses fréquences peuvent être reliées directement à des thermocouples, à des transducteurs d'extensomètres, ainsi qu'à des signaux de basse tension et bas courant, sans qu'il soit nécessaire d'employer des amplificateurs intermédiaires. Un choix de vitesses est offert de 0,05″ a 80″/sec en 20 plots. Un des atouts de cet enregistreur est la très longue durée de la lampe ultra-violette et la possibilité de la remplacer rapidement, par l'avant de l'instrument, en cas de panne.

EE 23 759 *pour plus amples renseignements*

COMPTEUR TRANSISTORISÉ

(Illustration à la page 452)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

Le compteur transistorisé de 1 MHz, Advance Type T.C.I., vient d'être ajouté à la nouvelle gamme d'instruments de laboratoire de haute qualité que produit actuellement la susdite société.

L'étalon de fréquence intérieure provient d'un oscillateur à cristal de 1 MHz contrôlé au four, fournissant une précision de fonctionement sur secteur de ±1 partie dans 10^8 à 25°C, et 5 parties dans 10^6 dans la gamme de 0°C a 40°C.

Des dispositifs sont prévus pour la mesure directe de fréquences jusqu'à 1MHz; pour la mesure, dans une gamme d'unités de temps de 1 à 10 périodes, de la forme d'onde d'entrée, pour le comptage d'impulsions électriques régulières ou aléatoires et pour la mesure de temps de 1 μsec à 2777 heures (10^7 secondes). Des impulsions de minutage, de 10^{-1} à 10^{-6} impulsions par seconde, par degrés de décades, sont assurées en tant que sortie du compteur.

La période de mesure de fréquence est de 0,1 1,0 ou 10 secondes et le type d'affichage est de 1 à 5 secondes ou fixe, avec répétition manuelle ou automatique du comptage.

Une indication à six chiffres d'une lecture facile est présentée sous une forme analogue à celle des appareils de mesure enligne du type à décades, avec étalonnage vertical. Des dispositifs d'autovérification sont incorporés pour essayer les deux chaînes de comptage et un enregistreur extérieur peut être relié au besoin. Les avantages qu'offre le T.C.I. peuvent être étendus avec profit par l'emploi du convertisseur fréquence très basse type C.A.I. qui prolonge la gamme de basse fréquences du T.C.I. de 10 Hz virtuellement jusqu'au courant direct (10^{-7} Hz).

EE 23 760 *pour plus amples renseignements*

RÉCEPTEUR UNIVERSEL DE COMMUNICATIONS

(Illustration à la page 452)

Marconi's Wireless Telegraph Co. Ltd, Marconi House, Chelmsford, Essex

La récepteur type HR120 est un appareil à double changement de fréquence, utilisant deux oscillateurs à cristal de commutation dont le premier est à accroissements de 1 MHz. Le premier des deux amplificateurs f.m. est à variation continue de 1 à 2 MHz et il est jumelé au second oscillateur dont la gamme va de 1,1 a 2,1 MHz, donnant ainsi une seconde fréquence moyenne de 100 kHz.

Le récepteur est doté de quatre commandes d'accord de décades, dont chacune est surmontée d'un cadran clairement étalonné en MHz × 10, MHz × 1, kHz × 100 et kHz × 1, donnant une précision de réglage de fréquence très élevée.

Le choix par commutateur de la largeur de bande se fait par les commandes du panneau avant, qui sont toutes groupées afin de faciliter au maximum le fonctionnement.

Les types de lampes les plus en faveur sont employées dans l'ensemble du récepteur et elles ont des équivalents Americains et "CV" de sécurité.

BRÈVES CARACTÉRISTIQUES

Applications:
Téléphonie à deux bandes latérales et à bande latérale unique. Télégraphie à ondes entretenues par intervalles.

Gamme de fréquences:
2,1 à 30 MHz couverture continue.

Niveau de bruit:
8dB à 30 MHz.

Sélectivité:
Quatre bandes passantes choisies par commutateur, donnant une largeur de bande de 1,5 3,6 et 12 kHz à −3dB.

Protection d'image:
50-75 dB, selon la fréquence reçue.

Sensibilité:
1,5μV en série avec 75 ohms dans toute la gamme pour un rapport signal/bruit de 20dB pour les ondes entretenues (1500Hz bande passante.)

Stabilité des oscillateurs:
Premier oscillateur à cristal: 2 parties dans 10^6/0°C. Second oscillateur; 10 parties dans 10^6/0°C.

Précision du réglage de fréquence:
±200 Hz après correction aux points de contrôle de 10kHz.

Antifading:
Pas plus de 10dB de changement de sortie pour une augmentation d'entrée de 100dB.

Antenne:
Pouvant être reliée à un cable coaxial de 75 ohms.

Filtre b.f.
Un filtre b.f. accordé a 1 kHz et ayant une bande passante de 100Hz peut être mis en circuit si nécessaire.

Sorties:
1W dans 2,5 ohms pour le haut-parleur. 1mW pour les casques. 6mW dans une ligne de 600 ohms.

Alimentations:
100-120 ou 200-250V, 45-65 Hz c.a. monophasé.

Consommation:
120W.

EE 23 761 *pour plus amples renseignements*

UNITOR À BASSE FORCE

(Illustration à la page 453)

Ferranti Ltd, Ferry Road, Edinburgh

Le nouvel Unitor à basse force, qui peut être obtenu maintenant sur le marché, a été spécialement conçu pour répondre au besoin d'une prise de courant à plusieurs pôles avec une force réduite d'insertion et de retrait.

La force d'insertion nominale exceptionnellement basse, de 113 grammes par pôle, est due à la construction particulière des étriers à douilles, dont les mâchoires sont actionnées par des ressorts indépendants.

La conception de l'Unitor se distingue particulièrement par le fait que la force d'insertion totale est en proportion directe du nombre de pôles. La force de retrait est à peu près des deux tiers de la force d'insertion et il y a une tolérance considérable de mauvais alignement.

La production actuelle de l'Unitor comporte des types à 35 pôles, à 70 pôles et à 91 pôles, avec contacts argentés. Tous ces types sont polarisés.

Le modèle à 70 pôles (de notre gravure) a subi avec succès les épreuves d'homologation prescrites par le D.E.F. 5321 et sa qualité a été acceptée par le Comité de Normalisation des Pièces de Radio (classification d'humidité: H1; catégorie de température: 40/100). On s'attend à ce que les autres types, qui sont de construction similaire, soient bientôt aussi homologués.

Les détails principaux de tous les types argentés sont comme suit: tension de service maximum: 500V; indice de courant: 5A; résistance de contact: 2 mégohms. Les dimensions totales du type à pôles sont: 1,9 pouces de largeur × 3,52 pouces de longeur.

Ces Unitors sont également produits avec des contacts dorés, particulièrement indiqués pour l'utilisation dans des atmosphères polluées et garantissant un excellent fonctionnement de contact dans les cas de tensions extrêmement basses.

EE 23 762 *pour plus amples renseignements*

UN SÉPARATEUR DE BALAYAGE

(Illustration à la page 453)

The Plessey Co. Ltd, Ilford, Essex

Un nouveau séparateur de balayage format 23, type 113D8H, destiné aux circuits de balayage de radar vient d'être mis au point par la société Ketay Ltd.

Cette pièce détachée a une réponse de fréquence plate s'étendant à la gamme de 100kHz, avec crête à 500kHz. Dans de nombreuses applications, la sortie du séparateur peut être reliée directement aux bobines de déviation du tube cathodique du système indicateur de position de radar. Un enroulement à réaction ou compensation est inclus dans le stator. La gamme de tensions va de 0 à 30V. La gamme de températures de fonctionnement du nouveau séparateur s'étend de −55°C à 75°C, et son poids est de 870 grammes.

EE 23 763 *pour plus amples renseignements*

POTENTIOMÈTRES DE PRÉCISION

(Illustration à la page 453)

Miles Electronics Ltd, Shoreham Airport, Sussex

La société Miles Electronics Ltd., a établi une nouvelle division pour la production de potentiomètres et d'autres pièces détachées.

Deux types de potentiomètres de précision sont réalisés actuellement et ce sont les premiers et les plus grands d'une série complète devant être produite sous peu.

Le type MCD 30/FC, qu'on voit sur la gravure, est un potentiomètre linéaire avec deux connexions d'extrémité et 33 prises intermédiaires standard.

Le diamètre extérieur de ce potentiomètre est d'environ 76,2 mm, et il a été conçu comme générateur de fonctions. En shuntant des sections de l'enroulement avec des résistances fixes convenables, on peut obtenir une gamme très étendue de sorties non-linéaires. Un avantage supplémentaire est que la fonction peut être aisément modifiée.

De doubles rangées de bornes du type à tourelle sont incorporées afin de faciliter le montage des résistances fixes, et les bornes sont identifiées aussi bien dans le sens des aiguilles d'une montre que dans le sens inverse.

Des contacts multiples en métal précieux sont utilisés pour l'assemblage sélectif frotteur, et des axes en acier inoxydable, ainsi que des cages à billes résistant à la corrosion, constituent des pièces standard. Jusqu'à six unités peuvent être groupées sur un seul axe. Des précisions linéaires de ±0,5% ±0,25% et 0,2% peuvent être assurées.

Le potentiomètre linéaire MCD 30/CT a 7,62 cm de diamètre et il est d'une construction générale semblable à celle du MCD 30/FG. Il est normalement doté d'une prise centrale et de doubles rangées de bornes. Un grand nombre de prises, à n'importe quelle position angulaire, peuvent être prévues. Jusqu'à six unités peuvent être réunies sur un seul axe. Des précisions linéaires de ±0,5% ±0,25% ±0,20% et ±0,15% peuvent être obtenues.

EE 23 764 *pour plus amples renseignements*

Résumés des Principaux Articles

Un indicateur pour simulateur de vol par A. G. Barnes et W. F. Coulshed

Résumé de l'article aux pages 402 à 407

La société English Electric Aviation Ltd. emploie un simulateur de vol pour évaluer les caractéristiques d'avions en cours d'étude et de réalisation. Une image du sol est présentée au pilote sur un tube cathodique dans le poste de pilotage simulé. Cet indicateur, dont l'image est basée sur celle du type à "contact analogique", a été mis au point à Warton en veu de répondre aux besoins spécifiques du simulateur de vol. Cet article décrit les limites de construction et les méthodes d'électronique employées pour produire l'image, donnant en même temps un bref aperçu de l'emploi de l'indicateur.

Le rôle des ondes de charge d'espace dans les dispositifs modernes pour hyperfréquences par G. D. Sims et I. M. Stephenson

Résumé de l'article aux pages 408 à 412

Cet article constitue une introduction à l'étude des propriétés des ondes de charge d'espace et au rôle qu'elles jouent dans le fonctionnement de certains dispositifs électroniques modernes.

La première partie traite des caractéristiques fondamentales de la propagation des ondes d'espace de charge et montre qu'elles peuvent être considérées en fonction d'une analogie de ligne de transmission. La deuxième partie analyse les modifications que subissent les ondes de charge d'espace par suite de différentes conditions de lancement ou de différentes symétries de faisceaux. Enfin, la troisième partie définit le rôle des ondes simples d'espace de charge, des ondes ioniques et des ondes de cyclotrons dans les oscillateurs à contre-ondes, les tubes à ondes progressives, les amplificateurs paramétriques et les plasmas simples.

Unités pour parasites de courant par P. L. Kirby et R. H. W. Burkett

Résumé de l'article aux pages 412 à 413

Les auteurs traitent des rapports entre la puissance des parasites de courant, la tension appliquée à une résistance et la bande de fréquences de mesure. Ils montrent que si le spectre de la puissance des parasites est 1/f, une unité pratique pour exprimer les parasites sera µV/V dans une décade, et ils proposent l'adoption de cette norme. Après avoir calculé les erreurs résultant de l'hypothèse d'un rapport fréquence 1/f pour les cas pratiques de $E_N^2 \propto 1/f.8$ *à* $E_N^2 \propto 1/f1.1$, *ils démontrent que de telles erreurs ne sont pas d'importance suffisante pour rendre l'unité envisagée impraticable.*

Extensomètres à compensation thermique par R. L. Chandler et E. J. Dent

Résumé de l'article aux pages 414 à 421

Les extensomètres à fil résistant sont communément utilisés par l'ingénieur de recherches et d'exploitation. Leur efficacité est, cependant, limitée par le fait qu'une erreur indéterminée peut s'introduire lorsque la température varie ou qu'elle n'est pas connue avec précision.

Ce désavantage a été surmonté par les Laboratoires de Recherches Napier, qui ont mis au point des extensomètres à compensation thermique dans une gamme étendue de températures. Ils sont dotés de thermocouples afin de produire une tension qui compense la tension de signal dûe aux changements de température, ne laissant que le signal causé par la contrainte appliquée.

Autant qu'on le sache, l'extensomètre à auto-compensation—qui est, evidemment, breveté—est seul en son genre. Dans les conditions difficiles d'utilisation pour lesquelles il a été prévu, il assure le seul moyen disponible d'effectuer des mesures de contrainte précises.

Un préamplificateur à capacité négative pour l'électrophysiologie par B. M. Johnstone et Ian D. Pugsley

Résumé de l'article aux pages 422 à 424

Les auteurs analysent un circuit à réaction donnant une capacité d'entrée négative. Ils évaluent, ensuite, les limites de réponse de fréquence. Un amplificateur transistorisé a été créé afin de pouvoir vérifier la théorie et indiquer d'autres limites pratiques.

Les limites de quadrature dans les circuits à tube simples d'une grande largeur de bande par R. M. Huey et J. E. Longfoot

Résumé de l'article aux pages 425 à 428

Les auteurs comparent les erreurs auxquelles il faut s'attendre de la part des circuits de quadrature utilisant tant la courbure d'une caractéristique de transfert de tube que les entrées antiphasées pour séparer des électrodes de commande. La première s'obtient par l'analyse théorique et les autres par la mesure.

Un contrôleur de radiation à transistor par M. Birnbaum

Résumé de l'article aux pages 428 à 429

L'auteur décrit un nouveau type de contrôleur de radiation, employant un seul transistor qui fournit la haute tension nécessaire au compteur de Geiger et qui intègre en même temps les impulsions devant être mesurées.

Un millivoltmètre transistorisé pour courant alternatif à grande impédance d'entrée par R. R. Vierhout

Résumé de l'article aux pages 435 à 437

Après une explication préliminaire du fonctionnement des systèmes à réaction, l'auteur décrit un millivoltmètre à transistor avec une impédance d'entrée de 1 à 60MΩ, une gamme de fréquences de 1 Hz à 200 kHz, une déviation totale de 10 mV dans la gamme la plus sensible avec une précision de 2% de l'échelle totale.

Circuits à transistors complémentaires par F. Oakes et C. Thompson

Résumé de l'article aux pages 438 à 442

L'existence de transistors n-p-n et p-n-p a conduit à l'élaboration de techniques complémentaires qui apportent aux constructeurs de circuits un nouvel outil puissant. Un nombre de circuits servant aux applications numériques et d'impulsions sont décrits dans cet article. Beaucoup de ceux-ci n'ont pas d'équivalents de circuits à tubes électroniques et sont d'un fonctionnement plus rapide et d'une plus grande souplesse qu'on ne pourrait obtenir avec une polarité unique de fonctionnement.

Un circuit simple pour produire une tension proportionnelle à la largeur d'une impulsion à répétition par H. C. Bertoya

Résumé de l'article aux pages 442 à 443

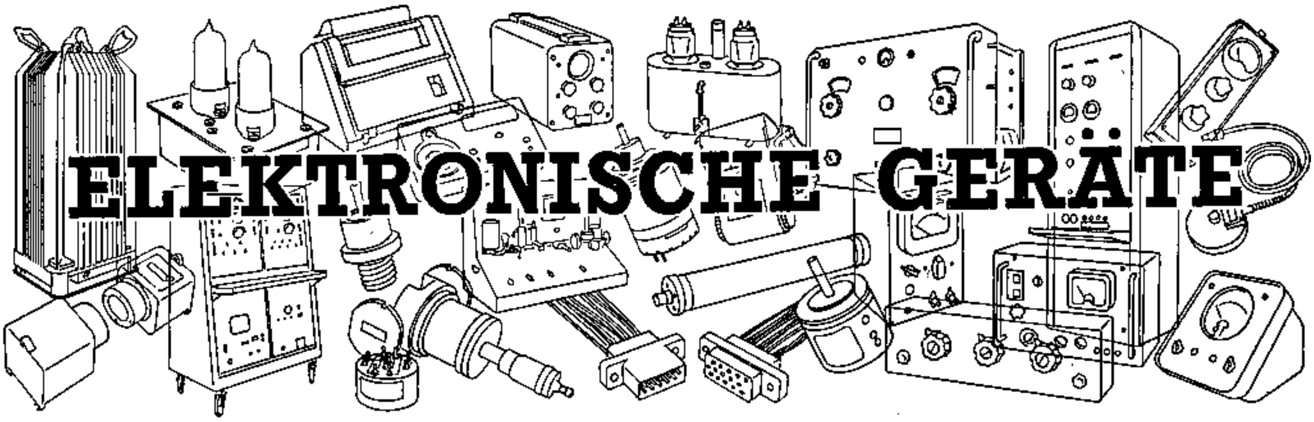
L'auteur décrit un circuit qui convertit une impulsion à répétition en une tension continue. La grandeur de la tension continue est proportionnelle à la largeur de l'impulsion. La réalisation de ce circuit est due à la nécessité de pouvoir disposer d'une impulsion qui puisse entraîner directement un moteur intégrateur. Le circuit est analysé et des détails sont donnés sur son fonctionnement ainsi que sur celui de l'intégrateur.

Un amplificateur transistorisé sensible, à plume enregistreuse par K. G. Beauchamp

Résumé de l'article aux pages 444 à 445

L'auteur démontre qu'en reliant un mouvement à plume enregistreuse à un montage en pont, dont les deux bras du pont consistent en éléments à transistors, une grande sensibilité peut en résulter.

Avec ce dispositif, un courant constant est tiré de la source d'alimentation et, vu que ce courant n'a pas lieu de dépasser le courant maximum d'entraînement de la plume, il en résultera donc une faible dissipation de la puissance des transistors et, partant, une dérive réduite de température.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 450 bis 453

EINE PRÄZISIONS-IMPULSQUELLE

(Abbildung Seite 450)

Communication Systems Ltd, Norfolk House, Norfolk Street, London W.C.2

Diese vollelektronische Präzisions-Impulsquelle hat eine Zeitintervall-genauigkeit in der Grössenordnung von wenigen Sekunden pro Jahr; das Gerät hat sehr kleine Abmessungen.

Ein quarzgesteuerter Oszillator erzeugt im Chronometer eine Grundfrequenz, die in eine Reihe binärer Teiler gespeist wird. Die Ausgänge dieser binären Teiler werden getrennten Stromstossrelais zugeführt, die in den erwünschten Intervallen Impulse hervorrufen. Die Spezialschaltungen sind volltransistorisiert, und gedruckte Schaltungen finden als Einsteck-Bausteine weitgehend Verwendung.

Das Universal-Standard-Chronometer gibt 3 Rechteckwellenimpulse in Abständen von 1,30 bzw. 60 Sekunden ab; andere Impuls-Abgabegeschwindigkeiten sind jedoch auch lieferbar. Die Anordnung für das Chronometer eines Forschungsschiffes sieht z.B. vor, dass 50 Impulse/Sek für Synchronisierung der Krafterzeugung und 1 000 Impulse/Sek zur Triggerung von Oszillograf-Ablenkteilen zusätzlich zu den 1,30 und 60 Sekunden-Impulsen für die Hauptuhreichung abgegeben werden. Ein ähnliches Gerät, das jedoch beliebig nach Stern- oder Weltzeit arbeiten kann, wurde für das Radioteleskop in Jodrell Bank geliefert. Das Chronometer kann auch 20 ms Zeitmarken für elektrokardiografische Untersuchungen abgeben. Diese können gleichzeitig und in Zeitphase an eine grosse Anzahl von Registrierinstrumenten abgegeben werden.

Frequenzstabilität wird durch Betrieb des Quarzes über einen besonderen Teil der Kennlinie erreicht. Dadurch werden automatisch die Schwankungen der Quarzfrequenz stabilisiert, die auf normale tägliche Temperaturänderungen zurückzuführen sind. Ein thermisch kompensierter Oszillator wird überall dort für Chronometer verwandt, wo die täglichen Temperaturschwankungen zu hoch sind oder grosse Temperaturwechsel von Tag zu Tag auftreten, wie z.B. in Flugzeugen und Schiffen. Die Kompensation arbeitet nach einem neuen Prinzip, das weder Heizer noch Thermostaten benutzt und daher keinen zusätzlichen Strom verbraucht. Für gegebene Zeitgenauigkeitsziele kann die Anordnung grössere Temperaturschwankungen vertragen. Andererseits kann für konstante Umgebungstemperaturen eine wesentlich bessere Zeitstabilisierung erzielt werden.

Die Genauigkeit des Chronometers wird durch Erschütterung oder Bewegung nicht beeinflusst. Das Chronometer ist daher besonders für fahrbare Anlagen, z.B. militärische und Rundfunkreportagezwecke, Züge, Schiffe und Flugzeuge geeignet. Standard-Modelle arbeiten mit 24 V Gleichspannung, fahrbare Ausführungen können aber auch für andere Spannungen geliefert werden.

EE 23 751 *für weitere Einzelheiten*

KLYSTRON-NETZGERÄT

(Abbildung Seite 450)

Elliott Brothers (London) Ltd, Valve Division, Borehamwood, Hertfordshire

Dieses Gerät wurde besonders für den Betrieb von Hochleistungs-Drift-Klystrons wie z.B. die Elliott-Typen 8FK1, 8TFK2 und 12FK1 entwickelt. Ein sehr schneller Überlastungsschutz sichert das Klystron im Falle eines internen Funkenüberschlages.

Das Gerät liefert alle zum Betrieb von Hochspannungsklystrons erforderlichen Spannungen. Die Schaltungen sind alle in einem auf Rollen laufenden Standgehäuse untergebracht, auf dessen Vordertafel alle Bedienungselemente, Überwachungsmessgeräte und Ausgangsbuchsen angebracht sind.

Die Heizspannungsversorgung kann stufenlos von 0...100 V Gleichspannung geregelt werden und gibt bis zu 5 A ab. Ausgangsspannung und -strom werden auf Messgeräten auf der Frontplatte angezeigt. Das Gerät ist für Betrieb bis zu 7 kV über Erde isoliert.

Bei Stromstärken bis zu 10 mA kann die Spannung für die Fokussierelektrode von 0...1000 V Gleichspannung stufenlos geregelt werden. Die Ausgangsspannung wird auf einem Instrument auf der Frontplatte angezeigt. Dieser Kreis ist auch für 7000 V isoliert.

Die gemeinsame Stromquelle für Kollektor und Resonator kann stufenlos von 0...5 kV Gleichspannung bei bis zu 150 mA Stromentnahme geregelt werden. Sowohl Kollektor- als auch Resonatorstrom werden auf Messgeräten überwacht.

EE 23 752 *für weitere Einzelheiten*

TRANSISTOR-VERSTÄRKER

(Abbildung Seite 450)

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

Der Transistor-Verstärker Type BCS 2429 ist ein tragbares Modell mit 15 W Ausgangsleistung, das bis zu drei grosse Lautsprecher mit Hornabstrahlung oder zehn bis zwanzig kleinere Lautsprecher mit guter Aussteuerung und Wiedergabequalität speisen kann. Zur Stromversorgung kann jede 12-14 V Batterie benutzt werden, die 2,5 A Spitzenstrom abgibt. Durchschnittsverbrauch ist 1 A.

Durch Verwendung von GEC Transistoren konnten neue Aufbau- und Anordnungsmethoden benutzt werden. Das Hauptgestell des Verstärkers besteht aus einem leichten, versteiften Gussstück, das gleichzeitig als Wärmeableitung für die Hochleistungstransistoren dient und die wichtigen, gewickelten Bauelemente trägt. Durch die Konstruktion wird hoher Widerstand gegen mechanische Erschütterung und Beschädigung erzielt. Auf der schrägen Bedienungstafel sind die drei als Mischer geschalteten Lautstärkeregler, die Toneinstellregler und eine Signallampe. Das Gerät wird in silberblauem und dunkelgrauem Hammerlack ausgeführt.

FREQUENZGANG:

Mikrofonstufen: flach innerhalb ±1,5 dB zwischen 80 Hz und 15 kHz (Mikrofoneingang 1 eingestellt für max. Bass). Phonostufe mit üblichem piezoelektrischem Tonabnehmer für einen im wesentlichen flachen Frequenzgang von 80 Hz..15 kHz korrigiert (Phonoregler in min. Stellung).

LEISTUNG:

Effektivwertmessung der Sinuswellenform an einem an die 15Ω Ausgangsklemmen angeschlossenen 15Ω Belastungswiderstand bei 1000 Hz:

Bei 14 V Gleichspannung—18 W mit 5% Klirrfaktor

Bei 12 V Gleichspannung—14 W mit 5% Klirrfaktor

Bei 14 V Gleichspannung—19 W mit 10% Klirrfaktor

Bei 12 V Gleichspannung—15 W mit 10% Klirrfaktor

Eingangsempfindlichkeit:

Für volle Ausgangsleistung sind erforderlich

Mikrofon 1 oder Mikrofon 2 Eingang 500 μV; Phonoeingang 100 mV.

Rauschabstand:

Alle Regler auf Kleinstwert	−60 dB
Phonoregler auf Höchstwert	−60 dB
Mikrofon 1 oder Mikrofon 2 Regler auf Höchstwert	−50 dB

EE 23 753 *für weitere Einzelheiten*

MESSBRÜCKE FÜR KLEINE KAPAZITÄTEN

(Abbildung Seite 451)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Diese Brücke Type TF 1342 misst Kapazitäten von 1 111 pF bis zu 0,002 pF herunter mit 0,2% Genauigkeit mittels einer dreipoligen Übertrager-Verhältnisarmbrücke. Durch ausserordentliches Unterscheidungsvermögen und Stabilität eignet sich die Brücke besonders zur Messung der Temperaturkoeffizienten von Kondensatoren und von Kapazitätsänderungen zwischen Röhrenelektroden.

Die Brücke misst die Kapazität zwischen zwei beliebigen Klemmen eines dreipoligen Netzwerkes und wird praktisch durch die Impedanz zwischen einer der beiden Klemmen und dem dritten Pol nicht beeinflusst. Daher können Verbindungen zum Prüfling mittels sehr langer Zuleitungen hergestellt werden, ohne die Messgenauigkeit zu beeinflussen. Diese Eigenschaft erlaubt Messung entfernter oder eingelöteter Prüflinge ohne Abschalten zugehöriger Kreise, was besonders wertvoll ist, wo zugehörige Shunt- und Streukapazitäten mitgemessen werden müssen, die für übliche Messungen nicht mit herausgenommen werden können.

Die ohmischen und Blindkomponenten des unbekannten Kondensators werden getrennt in einer Übertrager-Verhältnisarmbrücke ausgeglichen, wobei die unbekannten und Normalimpedanzen in gegenüberliegende Seiten der Brücke gelegt werden.

Dekadenschaltung und -ablesung sind vorgesehen; eine automatische Dezimalpunktanzeige erlaubt direktes Ablesen. Der effektive Shuntwiderstand des Prüflings wird durch einen Normalwiderstand ausgeglichen, der über drei wahlweise angeschaltete Übertragerabgriffe und ein stufenlos geregeltes Potentiometer gespeist wird.

Die Brücke hat einen eingebauten 1000 Hz Oszillator und Detektor. Die Detektorverstärkung kann hoch oder niedrig eingeregelt werden. Zur Erleichterung des Abgleichens ist eine automatische Verstärkungsregelung vorgesehen, die eine fortschreitende Empfindlichkeitserhöhung mit Annäherung des Nullpunktes gibt.

EE 23 754 *für weitere Einzelheiten*

AUFZEICHNUNG VON EREIGNISSEN

(Abbildung Seite 451)

A.E.I. Radio and Electronic Components Division, 155 Charing Cross Road, London W.C.2

Dieser Schreiber gibt eine zweckdienliche und zuverlässige Aufzeichnung von Datum und Zeit, zu der elektrische Anlagen ein- und ausgeschaltet werden, sowie der Zeitdauer, während der sie in Betrieb sind.

Der Antrieb ist unabhängig vom Netz und kann daher auch für Registrierung der Zeit verwendet werden, während der eine Maschine durch Stromausfall stillgelegt ist.

Das eingebaute Uhrwerk läuft 15 Tage mit einer Genauigkeit von ±5 Minuten während dieser Zeitspanne. Das Werk treibt das Registrierpapier mit einer Geschwindigkeit von 25,4 mm/h. Das kupferempfindliche Chromograph-Registrierpapier ist 9½ m lang und 15,9 mm breit, für Antriebszwecke an einer Seite perforiert und entsprechend den vom Benutzer gewünschten Verwendungszwecken vorgedruckt.

Eine Führung ist eingebaut, um den Ablauf des Papiers von einer Spule zur anderen zu erleichtern. Die Spulenkonstruktion ermöglicht leichtes Auswechseln des Registrierpapiers.

Der Schreibstift besteht aus dauerhaftem Kupfer und ist am Anker einer Magnetspule befestigt, die beim Einschalten der zu untersuchenden Anlage betätigt wird. Der Schreibstift wird beim Ansprechen der Magnetspule mit dem Papier in Berührung gebracht.

Die eingebaute Magnetspule arbeitet mit 200 V Gleich- oder Wechselspannung und 5 mA; Magnetspulen für 50...250 V Gleich- oder Wechselspannung können jedoch ohne weiteres geliefert werden. Wicklungen für niedrigere Spannungen müssen im Sonderauftrag hergestellt werden.

Durch Einbau einer weiteren Magnetspule mit Schreibstift kann das Gerät in einen Doppelkanalschreiber umgebaut werden. Auf Wunsch können sowohl Extrakanäle als auch andere Geschwindigkeiten als 25,4 mm/h geliefert werden.

Das Gerät kann auch mit Saphirstift und einem Synchronmotor ausgerüstet werden.

Der Schreiber kann entweder als selbständiges Gerät in Metallgehäuse mit grauer Hammerschlaglackausführung und abnehmbarem, unter Federspannung selbstschliessendem Deckel (Schloss auf Wunsch), oder als Einbau-Baustein für Ausrüstungen ohne Gehäuse geliefert werden.

EE 23 755 *für weitere Einzelheiten*

ZÄHLER FÜR DIE INDUSTRIE

(Abbildung Seite 451)

Airmec Ltd, High Wycombe, Bucks

Die Fertigung des neuen Zählers Type N256 für industrielle Zwecke hat begonnen. Das Gerät, das sehr vielseitige Einsteckbausteine benutzt, ist in einem widerstandsfähigen, staubdichten Gehäuse untergebracht. Zählen sowie Anzeige erfolgt mittels Dekatronröhren, und die höchste Zählgeschwindigkeit ist höher als die in der Industrie normalerweise benötigte.

Einsatzbausteine sind für fortlaufendes Zählen, Zählen bis zu vorgewählten Quantitäten, Eingangslähmung (zur Verhinderung unechter Zählung durch ungleichförmige Gegenstände), Signale für Satzquantitäten und Satzanlauf lieferbar. Bis zu sechs Bausteine für je ein Digit oder eine Sonderaufgabe können in den Hauptrahmen eingeschoben werden; daher kann fortlaufendes Zählen bis 999 999, Satzzählen bis 99 999 oder mit Satzanlaufwarnung bis 9 999 ausgeführt werden. Wird Eingangslähmung benötigt, so werden diese Höchstzahlen um ein Digit verringert. Die gewünschte Satzzahl wird für jeden Baustein von vorn mittels Drehschalter eingestellt.

Für den Zähler können eine Fotozelle und Lichtquelle geliefert werden; verschiedene andere Eingaben wie z.B. ein Schliesskontakt, Wechselspannungen usw. können jedoch auch benutzt werden. Vor leeren Gestellpositionen können Blindvordertafeln angebracht werden, damit das geschlossene Aussehen des Gerätes gewahrt bleibt.

EE 23 756 *für weitere Einzelheiten*

RECHNER-BEZUGSSPANNUNGS-QUELLE

(Abbildung Seite 452)

Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

Das Bezugsspannungsnetzgerät für Rechner Type AS 756·2 ist die nunmehr als getrenntes Gerät lieferbare Präzisions-Bezugsspannungsquelle des "Space"-Rechners. Sie liefert zwei unabhängige positive und negative Ausgangsspannungen; jedem Ausgang kann bis zu 200 mA entnommen werden.

Das Bezugsspannungsgerät besteht aus zwei hochstabilen Netzanschluss-Bausteinen mit driftkorrigierter Gegenspannungssteuerung und benutzt eine Kadmium-Quecksilber-Zelle als Grundbezugsquelle. Zur Gegenspannungssteuerung wird die Spannung der Weston-Zelle mit einem Bruchteil der negativen Quellenspannung verglichen, eine auftretende Fehlerspannung verstärkt und als Berichtigungssignal der Stabilisierungsröhre des negativen Netzgerät-Bausteins zugeführt.

Die positive Ausgangsspannung wird dann unter Bezugnahme auf die "korrigierte" negative über ein Summierungsnetzwerk und einen zweiten driftkorrigierten Verstärker stabilisiert.

KURZSPEZIFIKATION:

Ausgangsleistung

+100 V und −100 V, 200 mA für

jeden Netzgerät-Baustein

Genauigkeit und Stabilität

−100 V langfristig besser als 0,1%; +100 V kann innerhalb 1 mV der negativen Spannung eingeregelt werden und hält diese Spannung innerhalb 5 mV.

Stabilisierungsverhältnis

besser als 1000 : 1 für jeden Baustein

Gleichstrom-Quellenimpedanz

unter 0,05Ω für jeden Baustein

Restwelligkeit und Rauschen

unter 15 mV Spitze zu Spitze.

EE 23 757 *für weitere Einzelheiten*

TRANSISTOR-ANALYSATOR

(*Abbildung Seite 452*)

Avo Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

Der Analysator Type CT446 ist ein kompakter, tragbarer und batteriebetriebener Tester zur Prüfung von npn oder pnp Transistoren für kleine Signale und mittlere Leistung.

Der Tester ist mit dem Transistor in Emitterschaltung für Aufnahme von I_b/I_c Kennlinien mit verschiedenen V_c Spannungen geeignet. β-Werte können für jeden vorherbestimmten Punkt der I_b/I_c Kennlinie gemessen werden, und I'_{co}-Werte werden direkt in μA angezeigt.

Zur Messung des Transistor-Rauschens kann das Spitzenrauschen im ungefähren Bereich 100 Hz...10 kHz mit einem äquivalenten Eingangssignal des eingebauten 1000 Hz-Oszillators verglichen werden, das in dB geeicht ist.

Durch Buchsen auf der Frontplatte sind das eingebaute Überwachungsmessgerät, der 1000 Hz-Oszillator und der Hochleistungsverstärker für Verwendung ausserhalb des Gerätes verfügbar.

Kollektorspannungen sind in 5 Stufen zwischen 1,5 V und 10,5 V mit automatischer Polaritätsumkehrung für npn- und pnp-Typen vorhanden.

Kollektorstrom bis zu 1 A; Basisstrom von 0...1 mA und 0...40 mA. Stromverstärkung β 0...25 und 0...250. Für Rauschmessungen stehen zwei Bereiche zur Verfügung, die für 1...20 dB und 1...40 dB geeicht sind.

EE 23 758 *für weitere Einzelheiten*

DIREKTSCHREIBENDES U.V.-GALVANOMETER

(*Abbildung Seite 452*)

A.E. Laboratories Ltd, Feltham, Middlesex

Die Firma kündigt eine neue Serie direktschreibender U.V.-Galvanometer an. Modelle sind mit Registrierpapierbreiten bis zu 305 mm und für gleichzeitige Aufzeichnung von 50 Kanälen lieferbar. Für die Frequenzbereiche von 0...100 Hz bis 0...5000 Hz steht eine Serie ultrasensitiver Galvanometer zur Verfügung. Die niedrigeren Frequenzbereiche sind für direkten Anschluss ohne irgendwelche Zwischenverstärkung an Thermoelemente, Dehnungsstreifen und andere niedrige Spannungs- und Stromsignale geeignet. Geschwindigkeiten stehen zwischen 1,27 mm/s und 2,03 m/s in 20 Stufen zur Verfügung. Die lange Lebensdauer der Lampe und sehr schnelle Auswechslung einer fehlerhaften von der Vorderseite des Gehäuses aus sind besondere Merkmale.

EE 23 759 *für weitere Einzelheiten*

TRANSISTORISIERTER ZÄHLER

(*Abbildung Seite 452*)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

Der transistorisierte 1 MHz-Zähler TC1 stellt eine Erweiterung des neuen Advance-Fertigungsprogramms für hochwertige Präzisions-Messgeräte dar.

Der interne Frequenzstandard ist ein Quarzoszillator im Ofen, der bei Netzbetrieb und 25°C eine Genauigkeit von ±1 Teil in 10^6 erreicht. Über den Bereich 0° ... 40°C ist die Genauigkeit ±5 Teile in 10^6.

Mit dem Gerät können folgende Messungen ausgeführt werden: direkte Frequenzmessungen bis zu 1 MHz; Messungen über eine Serie von Zeiteinheiten, die 1 ... 10 Perioden der zugeführten Wellenform dauern; Zählung regelmässiger und zufälliger Impulse und Zeitmessungen von 1 µs ... 2 777 Stunden (10^7 Sekunden). Taktimpulse 10^{-1}...10^{-6} Impulse/Sek sind an den Ausgangsklemmen abnehmbar.

Die Frequenzmesszeit ist 0,1, 1,0 oder 10 Sekunden, und die Anzeige ist entweder von 1 ... 5 Sekunden oder „festhalten" mit Handbedienung oder automatischer Wiederholung des Zählens.

Eine leicht lesbare sechsstellige Anzeige in Form eines einzeiligen Dekadentypen-Messgerätes mit senkrechter Eichung ist vorgesehen. Selbstprüfung der Zahlenketten sowie Anschlüsse für einen Zusatzschreiber sind eingebaut. Der untere Frequenzbereich kann durch Verwendung eines Umsetzers für sehr niedrige Frequenzen Type CA1 von 10 Hz praktisch auf Gleichstrom (10^{-7} Hz) erweitert werden.

EE 23 760 *für weitere Einzelheiten*

MEHRZWECK-VERKEHRSEMPFÄNGER

(*Abbildung Seite 452*)

Marconi's Wireless Telegraph Co. Ltd, Marconi House, Chelmsford, Essex

Modell HR120 ist ein Überlagerungsempfänger mit doppelter Übersetzung, dessen erster Quarzoszillator in 1 MHz Stufen geschaltet werden kann. Der erste ZF Verstärker ist von 1 ... 2 MHz stufenlos regelbar und gleichlaufend mit dem zweiten Oszillator angeordnet, der 1,1 ... 2,1 MHz überstreicht und damit eine zweite ZF von 100 kHz ergibt.

Über jedem der vier Dekadenabstimmelemente ist eine deutlich in MHz × 10, MHz × 1, kHz × 100 bzw. kHz × 1 kalibrierte Skala angeordnet, wodurch eine ausserordentlich hohe Frequenzeinstellgenauigkeit erzielt wird.

Unter den auf der Vordertafel für besonders leichte Bedienung gruppierten Bedienungselementen befindet sich auch ein Schalter für Bandbreitenwahl.

Durchweg werden Röhrenvorzugstypen benutzt, die gegen amerikanische und Langlebensdauer-CV-Typen austauschbar sind.

Betriebsarten: Ein- und Zweiseitenbandsprechfunk CW—Ein/Aus-Telegrafie.

Frequenzbereich: 2,1 ... 30 MHz ununterbrochen.

Grenzempfindlichkeit: 8 dB für 30 MHz.

Selektion: Vier Bandpässe können durch Schalter eingestellt werden und sehen Bandbreiten von 1,5-3-6 und 12 kc/s bei −3dB vor.

Spiegelwellensicherheit: 50-75 dB; von Empfangsfrequenz abhängig.

Empfindlichkeit: 1,5 µV in Serie mit 75 Ω über den Gesamtbereich bei 20 dB Rauschabstand für CW und 1 500 Hz Bandbreite.

Oszillatorstabilität: Erster Oszillator 2 Teile in 10^6/°C; Zweiter Oszillator 10 Teile in 10^6/°C.

Einstellgenauigkeit (Frequenz): ±200 Hz nach Korrektur für 10 kHz Vergleichspunkte.

Automatische Lautstärkeregelung: Nicht mehr als 10 dB Ausgangsänderung für 100 dB Eingangssteigerung.

Antenne: Für Anschluss an 75 Ω koaxiales Kabel geeignet.

NF-Filter: Ein auf 1 kHz abgestimmtes NF-Filter mit 100 Hz Durchlass kann eingeschaltet werden.

Ausgangsleistung: 1 W in 2,5 Ω für Lautsprecher; 1 mW für Kopfhörer. 6 mW in 600 Ω Leitung 100 ... 120 oder 200 ... 250 V, 45 .. 65 Hz, einphasiger Wechselstrom.

Kraftbedarf: 120 W.

EE 23 761 *für weitere Einzelheiten*

UNITORS

(*Abbildung Seite 453*)

Ferranti Ltd, Ferry Road, Edinburgh

Ein neuer Unitor für kommerzielle Verwendung ist nunmehr lieferbar, der besonders entwickelt wurde, um den Bedarf für eine zuverlässige mehrpolige Kontaktleiste mit geringer Eindrück- und Ausziehkraft zu befriedigen.

Eine ungewöhnlich geringe Eindrücknennkraft von 113 g pro Pol des Unitors wird durch die einzigartige Konstruktion der Buchsenfederung erzielt, in der unabhängige Federn die Klemmbacken betätigen.

Ein wichtiges Konstruktionsmerkmal ist, dass die Gesamteindrückkraft proportional zur Polzahl ist. Die Ausziehkraft ist ungefähr 2/3 der Eindrückkraft, und die Kontaktleiste verträgt ziemlich schlechte Ausrichtung.

Zur Zeit sind 35polige, 70polige und 91polige Baumuster des Unitors mit versilberten Kontakten in Produktion. Alle Typen sind polunverwechselbar.

Das abgebildete 70polige Modell hat erfolgreich die in D.E.F. 5321 vorgeschriebenen Zulassungsprüfungen bestanden und hat vom RCSC Qualitätszulassung erhalten (Klimagruppe H 1; Temperaturklasse 40/100). Zulassung der anderen Modelle ähnlicher Konstruktion wird erwartet.

Pflichtblatteinzelheiten für alle versilberten Typen sind: Betriebshöchstspannung 500 V; Nennstrom 5A; Kontaktwiderstand 2 mΩ. Gesamtabmessungen der 70poligen Type sind 48,25 mm breit und 90 mm lang.

Diese Unitors werden auch mit hartvergoldeten Kontakten hergestellt und sind in dieser Ausführung besonders für Verwendung in verunreinigter Luft von grossem Wert; sie weisen besonders für sehr niedrige Spannungen überlegene Kontakteigenschaften auf.

EE 23 762 *für weitere Einzelheiten*

EIN ABTASTAUFLÖSER

(Abbildung Seite 453)

The Plessey Co. Ltd, Ilford, Essex

Ein neuer Abtastauflöser, Grösse 23, Type 113D8H wurde von Ketag Ltd. für Verwendung in Radarabtastkreisen entwickelt.

Der Baustein hat einen flachen Frequenzgang bis in den 100 kHz-Bereich mit einer Spitze bei 500 kHz. In vielen Fällen kann die Ausgangsspannung des 113D8H direkt an die Ablenkspulen der Bildröhre in der Rundsichtanlage gelegt werden. Auf dem Stator ist eine Ausgleichs- oder Gegenkopplungswicklung angebracht. Spannungsbereich des neuen Abtastauflösers ist 0-30 V und der Betriebstemperaturbereich $-55°C...+75°C$; Gewicht 800 g.

EE 23 763 *für weitere Einzelheiten*

PRÄZISIONS-POTENTIOMETER

(Abbildung Seite 453)

Mills Electronics Ltd, Shoreham Airport, Sussex

Mills Electronics Ltd hat eine neue Abteilung zur Herstellung von Potentiometern und anderen Bauelementen eingerichtet.

Zwei Präzisions-Potentiometertypen sind zur Zeit lieferbar, die ersten und grössten des umfassenden Programms, dessen Produktion anläuft.

Die abgebildete Type MCD 30/FG ist ein lineares Potentiometer und hat in der Standardausführung zwei Endanschlüsse und 33 Anzapfungen.

Dieses Potentiometer hat einen Aussendurchmesser von ungefähr 76 mm und wurde als Funktionsgeber entwickelt. Durch Parallelschaltung geeigneter Festwiderstände zu Abschnitten der Wicklung können nichtlineare Ausgangsspannungen über einen weiten Bereich erreicht werden. Daraus ergibt sich weiterhin, dass die Funktionen leicht abgewandelt werden können.

Doppelreihen von Kronenklemmen erleichtern den Anschluss von Festwiderständen. Die Klemmen sind sowohl rechtsläufig als auch linksläufig markiert.

Die Normalausführung hat Mehrfach-Edelmetallkontakte für den Schleifkontakt sowie rostfreie Stahlachsen und nichtkorrodierende Kugellager. Bis zu sechs Einheiten können auf einer Achse als Satzpotentiometer zusammengebaut werden. Bauelemente mit linearen Genauigkeiten von ±0,5%; ±0,25% und ±0,2% sind lieferbar.

Type MCD 30/CT ist ein lineares Potentiometer mit einem Aussendurchmesser von 76 mm und ist im allgemeinen Aufbau dem MCD 30/FG ähnlich. Ein Mittelabgriff und Anzapfungen in Doppelreihen sind in der Normalausführung vorgesehen. Eine grosse Anzahl von Anzapfungen können in jeder beliebigen Winkelposition geliefert werden. Satzpotentiometer mit bis zu sechs Einheiten können zusammengebaut werden. Lineare Genauigkeiten von ±0,5%; ±0,25%; ±0,20% und ±0,15% sind lieferbar.

EE 23 764 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Sichtgerät für einen Flugsimulator von A. G. Barnes und W. F. Coulshed

Zusammenfassung des Beitrages auf Seite 402-407

English Electric Aviation Ltd. benutzt einen Flugsimulator zur Bestimmung von Flugzeugeigenschaften in Entwurf- und Entwicklungsstadien. In einem nachgebildeten Führerraum wird dem Piloten auf einem Bildschirm eine Sichtdarstellung des Bodens gegeben. Diese Darstellung stützt sich auf die "Bodensichtanalog"-Methode und wurde in Warton unter besonderer Berücksichtigung des Flugsimulators entwickelt. Die Darstellungserzeugung innerhalb der durch Konstruktion und elektronische Technik gesetzten Grenzenwird zusammen mit einer Kurzbeschreibung der benutzten Darstellung besprochen.

Die Rolle der Raumladungswellen in modernen Mikrowellen-Bausteinen von G. D. Sims und I. M. Stephenson

Zusammenfassung des Beitrages auf Seite 408-412

Dieser Beitrag ist als Einführung zu den Eigenschaften der Raumladungswellen und ihrem Einfluss auf die Arbeitsweise gewisser elektronischer Bausteine gedacht. Teil 1 beschäftigt sich mit den grundsätzlichen Eigenschaften der Raumladungswellen-Ausbreitung und zeigt, wie die Wellen in Begriffen einer Übertragungsleitungsanalogie betrachtet werden können. Teil 2 bespricht, wie Raumladungswellen durch verschiedene Strahlsymmetrien beeinflusst werden können, und Teil 3 zeigt, welche Rolle einfache Raumladungswellen, Ionenwellen und Cyclotronwellen in Rückwärtswellenoszillatoren, Wanderfeldröhren, parametrische Verstärker und einfache Plasmas spielen.

Stromrauscheinheiten von P. L. Kirby und R. H. W. Burkett

Zusammenfassung des Beitrages auf Seite 412-413

Die Beziehung zwischen Stromrauschleistung, der an einen Widerstand angelegten Spannung und dem Frequenzbereich der Messung wird besprochen. Es wird gezeigt, dass μV/V in einer Dekade eine praktische Einheit für den Rauschausdruck darstellt, wenn das Rauschleistungsspektrum 1/f ist. Die Einführung einer derartigen Einheit wird vorgeschlagen. Der durch die Frequenzbeziehung 1/f enstehende Fehler wird für die praktischen Fälle $\overline{E_N^2} \propto 1/f^{0.8}$ *bis* $\overline{E_N^2} \propto 1/f^{1.2}$ *berechnet, und es wird gezeigt, dass der entstehende Fehler nicht gross genug ist, um die vorgeschlagene Einheit unbrauchbar zu machen.*

Temperaturkompensierte Dehnungsmesser von R. L. Chandler und E. J. Dent

Zusammenfassung des Beitrages auf Seite 414-421

Drahtwiderstands-Dehnungsmesser werden im allgemeinen von Forschungs- und Entwicklungsingenieuren benutzt. Ihre Wirksamkeit ist jedoch dadurch begrenzt, dass bei nicht genau bekannten Temperaturschwankungen ein unbekannter Fehler auftreten kann.

Dieser Nachteil wurde im Napier-Forschungslabor überkommen, wo Dehnungsmesser entwickelt wurden, die für Temperaturänderungen über einen weiten Bereich kompensiert sind. Zu diesem Zweck werden Thermoelemente benutzt, die eine Signalspannung hervorrufen, die die durch Temperaturänderungen hervorgerufene Signalspannung ausgleicht, so dass nur das der angelegten Belastung entsprechende Signal übrig bleibt.

Soweit bekannt ist, sind die natürlich patentierten selbstkompensierenden Napier Dehnungsmesser einzigartig. Unter den schwierigen Bedingungen, für die sie entwickelt wurden, sind sie die einzigen Hilfsmittel für genaue Dehnungsmessungen.

Ein Vorverstärker mit negativer Kapazität für elektrophysiologische Zwecke von B. M. Johnstone und Ian D. Pugsley

Zusammenfassung des Beitrages auf Seite 422-424

Eine durch positive Gegenkopplung hervorgerufene negative Eingangskapazität wird analysiert. Die Begrenzungen des Frequenzganges werden besprochen. Ein transistorisierter Verstärker wurde entwickelt, um die Theorie zu prüfen und weitere praktische Begrenzungen aufzuzeigen.

Grenzen der Rechteckformung mit einfachen Röhrenschaltungen grosser Bandweite von R. M. Huey und J. E. Longfoot

Zusammenfassung des Beitrages auf Seite 425-428

Die von der Benutzung der Kurvenform einer Röhrenkennlinie und von gegenphasigen Eingängen an getrennten Steuerelektroden herrührenden Fehler von Rechteckgebern werden miteinander verglichen. Erstere wurden durch theoretische Analyse, letztere durch Messungen erhalten.

Ein Strahlenüberwachungsgerät, mit einem einzigen Transistor bestückt von M. Birnbaum

Zusammenfassung des Beitrages auf Seite 428-429

Ein neues tragbares Strahlenüberwachungsgerät wird beschrieben, das mit einem einzigen Transistor bestückt ist, der die für das G.-M.-Zählrohr erforderlichen Spannungen liefert und gleichzeitig die gemessenen Impulse integriert.

Ein Transistor-Millivoltmeter für Wechselstrom mit hoher Eingangsimpedanz von R. R. Vierhout

Zusammenfassung des Beitrages auf Seite 435-437

Nach einer kurzen Einführung in die Arbeitsweise des Gegenkopplungssystems wird ein Transistor-Millivoltmeter beschrieben, dessen Eingangsimpedanz 1 . . . 6 MΩ ist. Es hat einen Frequenzbereich von 1 Hz . . . 200 kHz und im empfindlichsten Bereich einen Vollausschlag von 10 mV mit einer Genauigkeit von 2% des Vollausschlags.

Ergänzende Transistor-Schaltungen von F. Oakes und C. Thompson

Zusammenfassung des Beitrages auf Seite 438-442

Dass sowohl npn als auch pnp Transistoren zur Verfügung stehen, hat zur Entwicklung von Ergänzungstechniken geführt, die dem Schaltungsingenieur ein leistungsfähiges neues Werkzeug geben. Mehrere für Impuls- und Digitalzwecke nützliche Schaltungen werden beschrieben. Einige derselben haben kein Äquivalent in Elektronenröhrenkreisen; sie arbeiten schneller und sind anpassungsfähiger als Schaltungen mit nur einer Betriebspolarität.

Eine einfache Schaltung zur Erzeugung einer Spannung, die der Dauer eines sich wiederholenden Impulses proportional ist. von H. C. Bertoya

Zusammenfassung des Beitrages auf Seite 442-443

Eine Schaltung wird beschrieben, die einen sich wiederholenden Impuls in eine Gleichspannung umwandelt. Die Gleichspannungshöhe ist der Impulsdauer proportional. Die Schaltung wurde entwickelt, als sich herausstellte, dass die Impulse nicht direkt einen integrierenden Motor treiben konnten. Die Schaltung wird analysiert und Einzelheiten ihrer Leistung und der des Integrators werden gegeben.

Ein empfindlicher Transistor-Verstärker für Schreiber von K. G. Beauchamp

Zusammenfassung des Beitrages auf Seite 444-445

Es wird gezeigt, dass die Empfindlichkeit eines Schreibstiftantriebs erhöht werden kann, wenn er in eine Brückenanordnung geschaltet wird, in der zwei Brückenarme aus Transistorelementen bestehen.

In dieser Anordnung wird der Stromquelle ein konstanter Strom entnommen, und da dieser nicht den Spitzentreibstrom des Schreibers zu überschreiten braucht, werden die Transistoren mit geringerer Verlustleistung und daher mit geringerer Temperaturdrift arbeiten.