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Commentary

A DECISION on the system of colour television to be adopted in Britain has now been officially shelved until after the next meeting of Study Group XI of the C.C.I.R. in Vienna in April 1965. This means that there will be no public colour television service in this country before the end of 1966 or the beginning of 1967.

As already reported in the National Press the Postmaster General, the Rt. Hon. Reginald Bevins, M.P., has taken the advice of the Television Advisory Committee. This Committee, which is under the Chairmanship of Professor Sir Willis Jackson and includes representatives from the BBC, the ITA, the radio industry, certain Government Departments and three independent members, expressed the opinion that a unilateral decision might adversely affect the export of colour components to Europe and the easy exchange of programmes. Consequently it thought that no decision on the system to be adopted should be taken before the European delegates meet next spring. However, it went on to add that, "If that meeting fails to agree on a common system Britain should adopt a system of its choice". Mr. Bevins has fully endorsed this view and in his statement said that he had "...considered this advice in the knowledge that the BBC and the radio industry now takes the view that a colour service could not in any case begin to operate until the end of 1966 or the beginning of 1967 even if a decision were to be taken now. If a decision is not taken until April 1965 it will still be possible to begin a colour service early in 1967."

This decision is in line with the opinion expressed in last month's issue of *Electronic Engineering* (page 215) and it is one with which few people will disagree. While it may disappoint some manufacturers in the radio industry who are ready to push ahead with colour and are indeed rarin' to go, the benefits of a unified system in Europe, or throughout the world if that was possible, will be universally accepted. The present decision to give the C.C.I.R. another chance to make up its mind but with an ultimatum that Britain will not 'hedge' indefinitely seems a fair compromise.

So for the moment we can but wait and see and also hope that the next time the delegates of Study Group XI meet they will do so in a more resolute and business-like mood than they have shown in the past; for their performance to date has done little to foster faith in Mr. Bevin's statement that "...the prospects at next year's meeting are pretty good and we should be able to get an agreement". The choice of the best colour system for Europe is one which should be made on solely technical grounds, but it seems that national loyalties and politics of one sort or another are having an increasing influence on the delegates and there can be no very great hope that this is likely to change within the next twelve months. On the other hand it is a sufficient length of time for developments in any one of the systems under discussion to take place which would make it an unquestionable choice.

In cogitating on the future of colour television in Britain there are two other factors that must be taken into account. The first is that by next April neither Mr. Bevins nor his colleagues may be in a position to implement the decisions which he has recently announced. The second is, what impact will colour have? On this point it is, of course, not possible to be dogmatic but from the evidence available there is no very cogent reason to suppose that its impact will be all that great whether it is introduced now or in five years time. In the present arguments over which system should be adopted it is very easy to get carried away and to see any delay in reaching a decision as a setback to national progress. However, in America, where a colour service has been in operation for some years, its impact can hardly be said to have been enormous. Receivers have remained very expensive and the quality of pictures presented to the viewer often leaves very much to be desired. There is no reason to suppose that this situation is likely to change until a fundamentally new method of picture presentation is evolved and at the moment a more satisfactory and cheaper method than the shadow mask tube isn't even on the horizon.

Colour television, therefore, is temporarily in obeyance and in the next year or two the most interesting thing on the home television front will be the performance of u.h.f., as evidenced in the first instance by the establishment of BBC-2.

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The International Instruments, Electronics and Automation (I.E.A.) Exhibition is due to be held at Olympia, London, from 25 to 30 May.

This year's exhibition promises to be the most comprehensive of its type in the world. Some seven hundred firms, including 130 from overseas, will be exhibiting goods valued at around £20M. The last I.E.A. exhibition attracted 92 000 visitors including 7 000 from overseas. It is anticipated that these figures will be exceeded this year and that over 10 000 foreign engineers and buyers will attend.

From the visitor's point of view the greatest problem will be to do justice to the 700 stands in the limited time at his disposal but from the advance information so far to hand the effort will be well worthwhile.

Pulse Period Meter with Short Response Time, Applied to Cardiotachometry

By P. A. Tove* and J. Czekajewski†

A method of measuring pulse repetition rate with fast response to frequency changes has been developed. The principle is to measure intervals between pulses by alternate charging of two capacitors during successive pulse intervals. The voltage on that capacitor which is not being charged is indicated on a meter or oscilloscope and is a measure of the length of the last period. Direct scale indication of pulse interval length or frequency is achieved and the possibility of obtaining linear and logarithmic frequency-scales is discussed. It is shown that simple exponential charging of the capacitors approximates the desired scale shapes over limited frequency regions. Methods of obtaining linear and logarithmic scales over large regions are described. Practical circuits are described for medical applications in cardiotachometry and registration of respiratory frequency where it is essential to have fast response to changes in pulse rate. Another application is in information transmission systems using pulse length modulation.

(Voir page 352 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

In science and technology one meets very often the problem of measuring the rate of pulses, which are recurrent with a perhaps varying frequency. In medicine, measurements on heart beat and respiratory frequencies are examples. In this article is described a new type of pulse tachometer which allows a direct reading, on a deflexion meter, of pulse repetition rates of low frequencies. A recorder could be added in order to obtain a permanent record of the variations. The characteristic feature of the instrument is a very short response time, because the principle of measurement involves individual registration of each pulse interval and subsequent indication of it during the next interval.

In the 'digital age' the method at hand for measurements of this type would be registration on a counter. However, in addition to implying more instrumental mobilization, this method has the disadvantage that most counters[1] either do not count at all during the display interval (a fraction of a second or more), or in any case require a not negligible rest period between counts. This means that only every other interval can be measured, while the present instrument registers every interval.

A technique used in direct-reading pulse counting-rate meters[2] in nuclear electronics is to reshape the incoming pulses into uniform pulses and let these add up in an average-reading device, which, by measuring the number of pulses per unit time, gives an indication of the mean value of the pulse frequency, and hence the pulse separation. However, it responds slowly to changes as can be seen from the basic circuit shown in Fig. 1. Here, each input pulse transfers a charge C_1E to C_2. In order that the voltage V across C_2 shall give an indication of the mean pulse rate the time-constant R_2C_2 must be many times the separation between input pulses, with an attendant slow response. Principally the same circuit is used in direct-reading frequency meters[3]. The averaging function of the integrating circuit R_2C_2 may often be performed by the mechanical inertia in a galvanometer. Inherently, these types of instruments cannot respond to fast changes in pulse separation.

Instruments have, however, been built according to other principles, which allow immediate response to change in pulse separation. One such instrument[4] is used for information transmission by pulse length modulation, and has been developed for supervisory control of remote electrical power stations. It is in principle a pulse-length-to-voltage convertor. A constant-slope sawtooth pulse, with

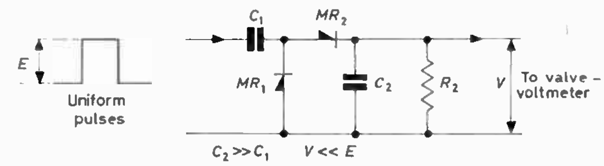


Fig. 1. Basic principle for conventional pulse counting-rate meter or direct reading frequency meter

Because $C_1 << C_2$ the input pulse voltage is nearly all across C_1 and a charge C_1E is transferred to C_2. MR_1 acts as a d.c. restorer

length equal to the input pulse, is generated on an integrating capacitor. At the end of the pulse, the voltage finally reached is sampled and transferred to a small capacitor, which is connected to a valve-voltmeter. The integrating capacitor is then discharged. A disadvantage is that the system requires that the pick-off capacitor should be much smaller than the integrating capacitor, and hence the stored energy of this is only utilized to a small extent in the meter circuit and the deflexion decays rapidly.

The instrument to be described still uses the principle of measuring time by capacitor charging, but the energy is utilized better, and the circuit also has other interesting properties.

Circuit Description

The basic idea in the present circuit is to measure the distance between two adjacent pulses, or the length of one period in a waveform, by putting the information on a memory (a capacitor), which affects the indication meter during the next period. Simultaneously, the length of this is measured and the result transferred to another memory capacitor which in turn affects the indicating instrument during the next period, and the course repeats. That capacitor which is to receive the information about the length of a certain interval is first rapidly discharged at the beginning of it, in order to remove previous information. The principle is shown in Fig. 2. C_1 and C_2 are the memory capacitors. The switches S_1, S_2 and S_5 change state every time a pulse appears. S_3 and S_4 are switches which perform the previously mentioned instantaneous

* *University of Uppsala, Sweden.*

† *On leave from University Marie Curie-Sklodowska, Lublin, Poland.*

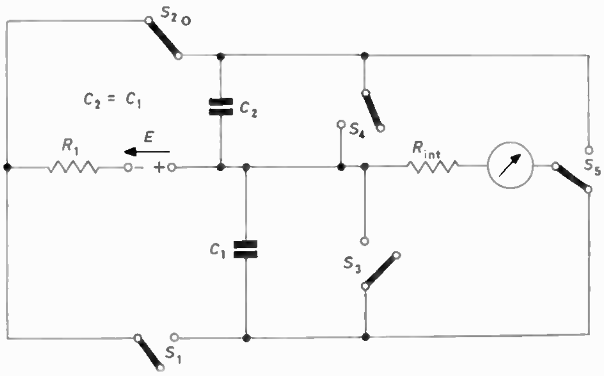


Fig. 2 Principle of the period meter

erasing of information on C_1 and C_2. They work alternatingly, e.g. S_3 is closed a short time by every other input pulse and S_4 by those in between. Starting with the appearance of a pulse, either S_1 or S_2, whichever will be closed until the next pulse appears, will cause a charging of its corresponding capacitor (C_1 or C_2) by feeding a current from the battery supply E through the resistance R_1. (Before charging started, a quick discharge was made by the appropriate one of S_3 or S_4). The length of time between pulses determines the attained capacitor voltage because this follows the well-known relation:

$v = V[1 - \exp(-t/R_1C)]$ where C stands for C_1 and C_2.

If, say, C_2 has been charged, then during the next interval the galvanometer senses its voltage, through the action of S_5. The internal resistance R_{int} of the galvanometer branch is so high that the discharge of the capacitor is negligible and the instrument deflexion remains practically constant during the pulse interval. When the next pulse appears S_5 switches the instrument to measure the voltage of C_1 which was charged during the past interval. If the pulse intervals are equal the instrument deflexion remains the same when S_5 switches, because $C_1 = C_2$. The situation changes, if the pulse separations vary during measurements. Then, the meter measures different voltages on the capacitors every time the relay switches. As a consequence the meter reacts immediately to every change of time separation between the pulses. The speed of reaction is limited only by the inertia of the mechanical system. Two circuits based on the above principle were constructed. The first uses relays for the switching functions, and the second transistors. Both were developed for the medical application of measuring heart-beat rates (cardiotachometry), and respiratory frequency. In these applications it is especially valuable to have a circuit that responds instantaneously to frequency changes.

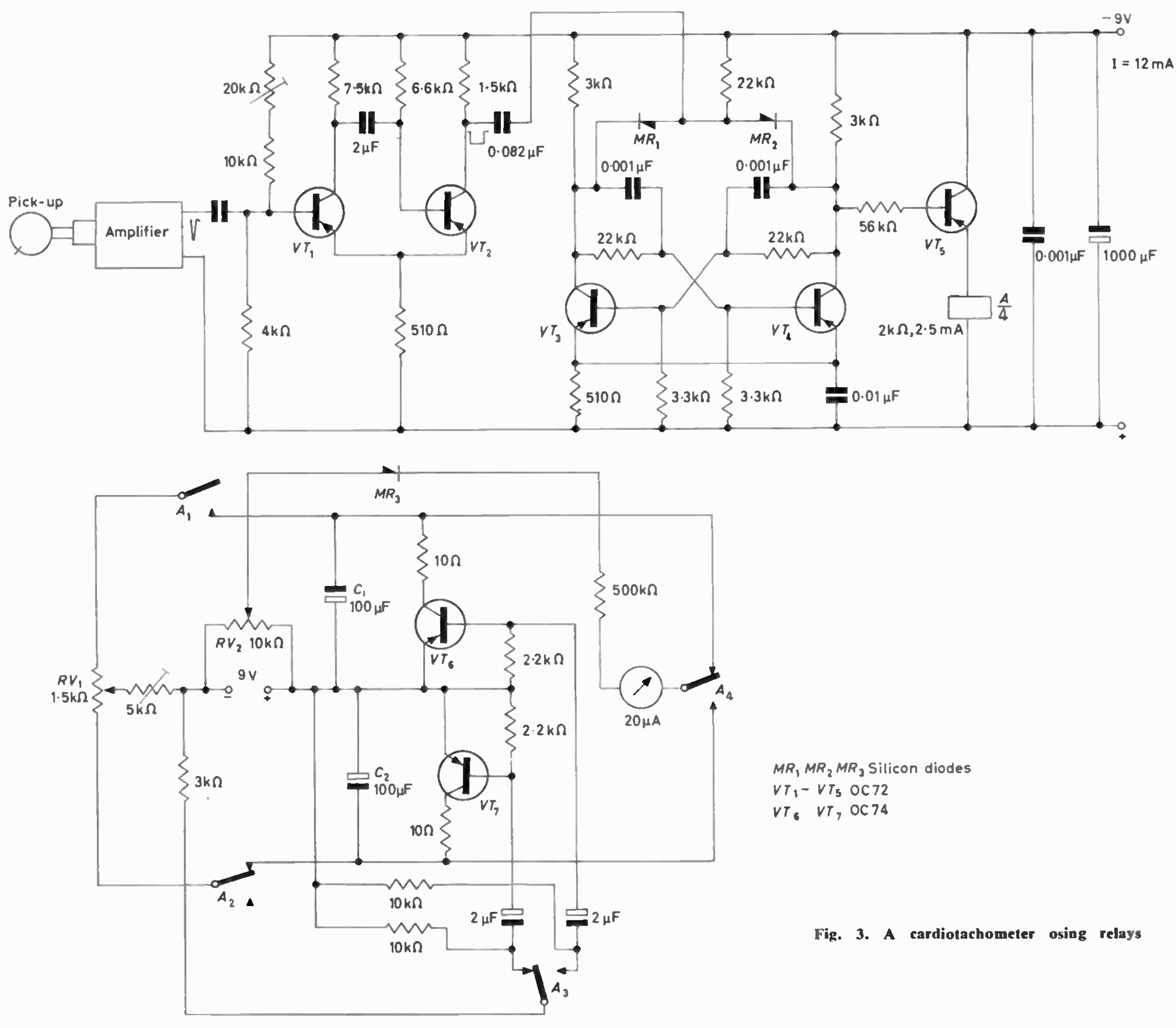


Fig. 3. A cardiotachometer osing relays

Circuit Employing Relays

The function of the first circuit is described by the help of the circuit diagram in Fig. 3. The input pulses, which may come from a pick-up, such as the terminals of an electrocardiograph, are amplified up to a level of a few volts max. and trigger a univibrator (VT_1, VT_2), which gives uniform pulses ($\simeq$140msec long) for switching the flip-flop (VT_3, VT_4). The pulse length is chosen rather long to make the circuit insensitive to disturbing pulses on the input during this time. The switching, which occurs on the trailing edge of the pulses is done via diodes which ensure that only the proper collector (the one being at

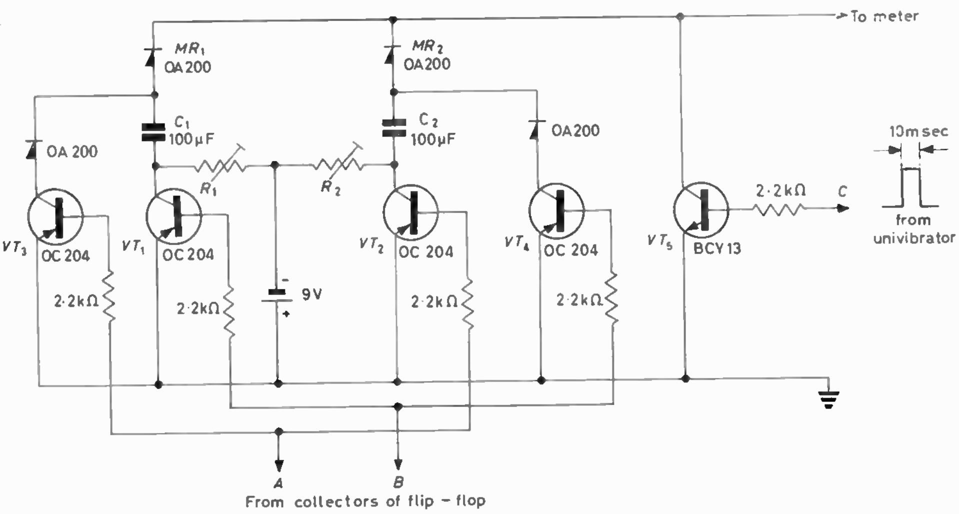


Fig. 6. Transistorized version of the cardiotachometer

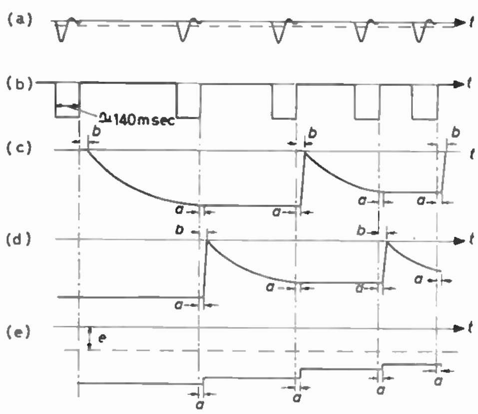


Fig. 4 (above). Voltages in the circuit of Fig. 3 as a function of time for a case of decreasing pulse separation

(a) Triggering pulses for univibrator, obtained from amplifier following pick-up

(b) From univibrator output collector

(c) Across capacitor C_1

(d) Across capacitor C_2

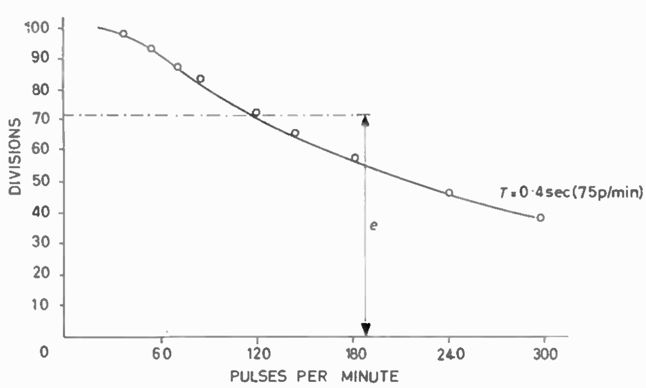
(e) On the meter

e is threshold of sensitivity (set by potentiometer RV_2 in Fig. 3).

Small time intervals a are due to delay in the mechanical action of the relay. Intervals b, in addition, include the discharging time of contacts A_3

Fig. 5 (below). Deflexion of the instrument as a fonction of pulse frequency

Theoretical curve and experimentally measured points are shown. The time-constant chosen = 0·4sec so that best linearity is obtained around 75p/min. The upper non-linear part of the curve can be arranged to fall outside the meter scale



low voltage) is influenced. The relay A is controlled by the flip-flop via the isolating amplifier VT_5. The functions of the switches S_1, S_2 and S_5 are performed by relay contacts A_1, A_2 and A_4 while the action of S_3 and S_4 is performed by the switching transistors VT_6 and VT_7, which are actuated, for a short time only (< 20msec) when the relay contacts A_3 close (alternately), because of the differentiating circuits in the base leads of VT_6 and VT_7. Potentiometer RV_1 is used for matching the time-constants of the two capacitor systems. Their time-constant has been chosen to be $\simeq$0·4sec. As discussed in the appendix treating the relation between time-constant and scale shape this is the proper value to get best utilization of the meter scale for the most interesting range of frequencies, which is around 75p/min in cardiotachometry. The voltages across some circuit elements and their relations are shown in Fig. 4. Fig. 5 shows the deflexion of the meter as a function of the pulse frequency and both experimentally measured points and a theoretical curve are shown. The sensitivity threshold (e in Fig. 4(e) of the meter has been adjusted, using the potentiometer RV_2 and the diode MR_3, to utilize the meter scale better for the interesting pulse frequency range $\simeq$35 to 110p/min. Because of the threshold, higher frequencies do not deflect the meter. If another frequency range is wanted the charging time-constant could be easily changed.

The relay is a normal telephone relay and has a resistance of 2kΩ and a pull-in current of 2·5mA.

A remark should be made on the properties of capacitors C_1 and C_2 and their discharging circuit. Aluminium electrolytic capacitors with 100V rating, i.e. higher than the circuit voltages, were used and their leakage resistance was then several tens of megohms. For highest stability with time tantalum capacitors could be useful. Transistor switches VT_6 and VT_7 should have sufficient current capacity (still having high resistance when cut-off) to allow rapid discharge of C_1 and C_2. If the discharge time is too long, the scale may deviate from the theoretical shape as understood from Fig. 4 and the appendix, if a finite length of intervals a and b has to be taken into account. If one is interested in using less sensitive ammeters, or to measure very low frequencies without any decay of the deflexion in the period between pulses, one may use a higher voltage (which allows higher instrument branch resistance), or use large capacitance, which in turn would require more powerful discharging transistors. It would also be possible to use an emitter-follower ahead of the meter.

Circuits Employing Transistor Switches

More refined versions of the previous circuit were developed by replacing the relays by transistors. It is then advantageous to re-arrange the switches in the basic circuit, Fig. 2, in order to permit the transistors to have one common terminal. One such arrangement that was used is shown in Fig. 6. The transistor switches are controlled from a univibrator and flip-flop arrangement corresponding to that in the relay circuit, Fig. 3, in such a way that when VT_2 and VT_3 are kept conducting, VT_1

The arrangement of another transistorized version of the basic circuit is shown in Fig. 7. In one state of the flip-flop VT_7 is conducting and VT_{10} is prepared to conduct while VT_6 and VT_9 are cut-off. Capacitor C_2 is then charged through R_2 and MR_2. The voltage on C_1 is applied to the meter through MR_5. The flip-flop is switched by the leading edge of the univibrator. With the beginning of the next pulse VT_7 and VT_{10} are cut-off while VT_9 conducts and VT_6 is prepared to. VT_8 is conducting for the length of the univibrator pulse and hence C_1 is discharged through MR_3,

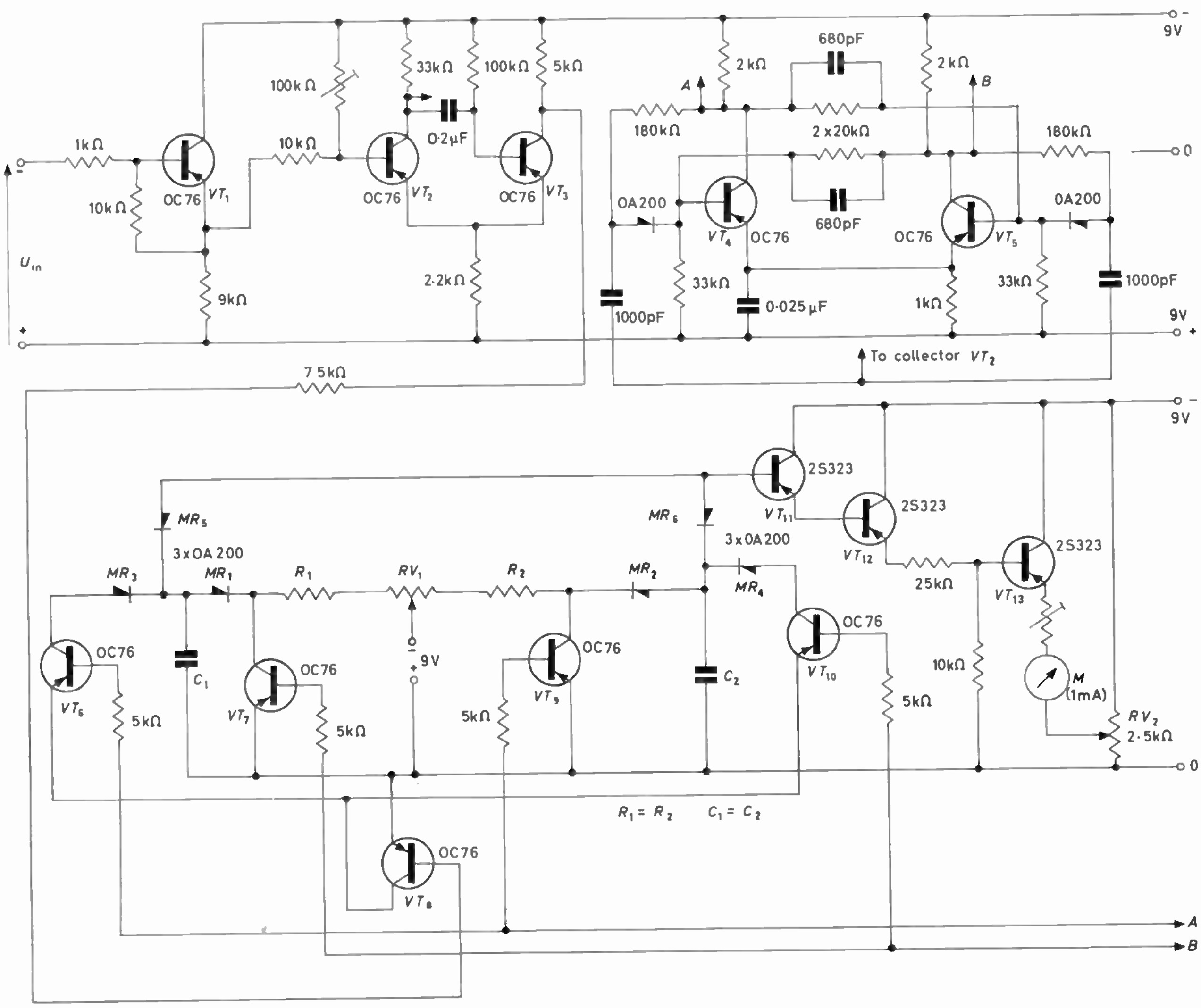


Fig. 7. Practical transistorized cardiotachometer

and VT_4 are cut-off and vice versa. The state VT_2, VT_3 conducting and VT_1, VT_4 cut-off means that C_1 is charged through R_1 from the battery, while C_2, which was charged during the previous period, is connected to the meter through diode MR_2 and transistor VT_2. The flip-flop is triggered by the trailing edge of the univibrator. The pulse of this makes VT_5 conduct for a short time and thus that capacitor which is to be charged is first discharged through VT_5 and VT_1, MR_1, or alternatively VT_2, MR_2. During the 'on' periods of VT_5 there is no output voltage to the meter. Therefore the discharge time should be short to avoid flickering of the meter.

This type of circuit is also suitable for much higher frequencies. In this case it may be desirable to use an oscilloscope instead of the meter.

VT_6 and VT_8. The cascade emitter-follower with low-leakage silicon transistors presents high input resistance to C_1 and C_2. RV_1 serves to equalize the time-constants of C_1 and C_2 and RV_2 is the bias adjustment for the meter.

With this circuit, the action of the discharge of the capacitors is not seen on the output. Also, as soon as the length of a new interval exceeds that of the preceding, indication of this is otained on the meter. This is because MR_5 and MR_6 connect either C_1 or C_2, whichever has the highest voltage, to the emitter-follower. This is in contrast to the previous circuits. The input pulse for the circuit of Fig. 7 should exceed 2·5V. The univibrator pulse is made rather short ($\simeq$ 10msec) in order not to distort the scale, because it takes time from the charging period. C_1 and C_2 were 100μF tantalum capacitors. The circuit was tested for the range from 5p/min to 240p/min which is sufficient for cardiotachometry and respiratory frequency registration. It was covered in five sub-ranges by changing the values of R_1 and R_2 to obtain range centre frequencies f_0 of 10, 20, 40, 80 and 160p/min, respectively. Each range covers from $f_0/2$ to $3f_0/2$, within 10 per cent linearity and the resistors were calculated from $T = RC = 1/2f_0$ (see appendix).

Discussion of Scale Shapes

Two types of capacitor charging characteristics are easily obtained, constant-current charging giving a voltage linearly increasing with time, and exponential charging, with voltage following the function $1 - e^{-t/RC}$. The scale shapes (voltage-frequency characteristics) that may be desirable are, in particular, linear and logarithmic and the question arises to what extent these can be obtained directly from the charging characteristics, or what modifications be necessary.

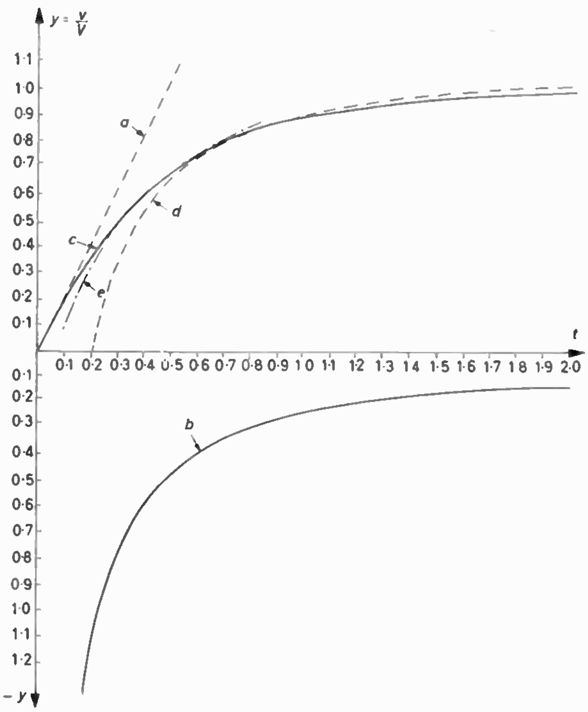


Fig. 8. The relation between desired scale characteristics (linear and logarithmic) and the appropriate capacitor charging characteristics

(*a*) *Depicts constant-current charging and gives a scale linear with time interval*
(*b*) *Corresponds to a scale having deflexion proportional to frequency*
(*c*) *Illustrates exponential charging*
(*d*) *Makes deflexion a linear function of frequency*
(*e*) *Gives a deflexion which is a logarithmic function of frequency*
The parameters of (d) and (e) are chosen so that they are approximated to the best degree by the exponential charging function (c). The proportionality factor of (b) has arbitrarily been chosen equal to that of (d)

Constant-current capacitor charging gives an output voltage v proportional to pulse separation t, $v \propto t$, which is depicted in curve (a) of Fig. 8. This is equivalent to a hyperbolic frequency scale, $v \propto 1/f$. A scale having voltage proportional to frequency, $v \propto f$, would be obtained if the relation between voltage and period time could be made hyperbolic, $v \propto 1/t$, corresponding to curve (b) in Fig. 8 (shown in the fourth quadrant with negative voltage in order to point out that the curve slope is similar to that of the physically realizable charging functions).

A *linear* relation between v and f, $v = c_2 - c_1 f$ is obtained if curve (b) is shifted in voltage (curve (d) in Fig. 8). It is seen that this curve is approximated reasonably over a limited region by an exponential charging curve $v \propto 1 - e^{-t/RC}$ (curve (c) in Fig. 8) and this is the explanation for the approximate linearity over a portion of the scale in the described instrument. In the appendix it is shown that if linearity is desired around a certain frequency f_0, then the exponential time-constant $T = RC$ should preferably be made equal to $1/2f_0$.

A possible way to obtain a more exact realization of the proportional scale, $v \propto f$, would be to start from the relation $v \propto t$, obtaining v by constant-current charging, and then to perform an inversing operation on v, i.e. to produce a new voltage v_1, proportional to $1/v$ by any suitable dividing technique as used in analogue computing[5]. One will then have $v_1 \propto f$.

A *logarithmic* scale could be of the form $v = c_4 - c_3 \log f$ which means that the voltage changes an equal amount for, say, each two-fold increase in frequency. A logarithmic form (plotted as a function of $t = 1/f$, instead of f) is shown as curve (e) in Fig. 8 and it is seen that this is approximated over a limited range by the exponential curve (c) if the constants c_4 and c_3 and thus the contact point between the curves are chosen appropriately, as in the figure. In the appendix it is shown that if logarithmic behaviour is desired around a certain frequency f_0 the exponential charging time-constant T should preferably be made equal to $1/f_0$. The range over which a reasonable logarithmic behaviour could be obtained is small, covering about a factor 4 in frequency.

It is easily understood that the functions $v \propto t$ and $v \propto 1/t$ do not make a logarithmic scale possible (they are equally bad in this respect).

Two possibilities to obtain wider scale ranges, both when linearity and logarithmic behaviour are desired, should be mentioned. Either some correcting network could be used for transforming the voltage output of a simple linear or exponential charging device, to give the desired scale shape, or the charging circuit itself could be modified to give directly the necessary form of capacitor voltage versus time.

A suitable corrective network for obtaining a logarithmic scale would be a logarithmic amplifier fed by the output of a constant-current charger which gives a voltage $v \propto t$. At the output of the logarithmic device one obtains $v \propto \log t = \log(1/f) = -\log f$. A pitch-recorder for music analysis has been developed along these lines[2].

Modification of the charging curve in an RC device could be obtained by the use of voltage-dependent components (varistors and varactors) or by switching different values of R and C into the circuit at chosen times during the charging cycle.

Conclusion

The principle described of registration of pulse intervals has the advantage of being simple and affording instantaneous direct reading of the result on a meter. Along with the described applications as a cardiotachometer or respiratometer for registration in the range 5 to 240p/min, the circuit shows promise of wider application for higher frequencies, when instantaneous response to rapid changes in pulse separation or frequency is desired. It could also find applications in pulse-width modulation systems.

Appendix

Choice of Exponential Charging Time-constant T

(a) *Linear Scale Desired*

A linear scale (charged capacitor voltage versus frequency f) would be of the form $v_{lin} = c_2 - c_1 . f$.

It will be investigated whether the exponential charging function $v = V(1 - e^{-(t/T)})$ may approximate to this, by developing v in a Taylor series of $f = 1/t$ around a certain centre frequency f_0.

The general expression for the series is:

$$v(f) = v(f_0 + \Delta f) = v(f_0) + v'(f_0)\Delta f + (v''(f_0)/2!)(\Delta f)^2 + (v'''(f_0)/3!)(\Delta f)^3 + \ldots.$$

where v', v'' etc. represent derivatives with respect to f, and Δf is the deviation from f_0. Evaluation gives:

$$v(f)/V = 1 - \exp(-1/Tf_0).[1 + \Delta f/Tf_0^2 - (\Delta f)^2.(1 - 1/2Tf_0)/Tf_0^3 + (\Delta f)^3.(1 - 1/Tf_0 + 1/6T^2f_0^2)/Tf_0^4 + \ldots\ldots\ldots]$$

Terms in $(\Delta f)^2$ and higher mean deviation from linearity. Terms of higher order grow smaller. The condition for vanishing second-order term is $1 - (1/2Tf_0) = 0$ and to fulfill this the time-constant T of the RC circuit should be chosen $T = (1/2f_0)$. A rough measure of the deviation from linearity under this condition gives the ratio of third-order and first-order terms which is found to be $-(1/3)(\Delta f/f_0)^2$. As an example, for $\Delta f = (f_0/2)$ this has the value 1/12. The fit to linearity could also be judged from Fig. 8.

In the described cardiotachometer circuit the expected mean frequency f_0 is 75p/min and the above relation $T = (1/2f_0)$ gives $T = 0{\cdot}40$sec.

(b) *Logarithmic Scale Desired*

A logarithmic scale would have the form:

$$v_{\log} = c_4 - c_3 \ln f$$

To see to what extent this form is followed by the charging function $v = V(1 - e^{-(t/T)})$ this may first be rewritten in a form with $\ln f = \ln(1/t)$ as variable instead of t. This gives $x = \ln\left(\dfrac{1}{T\ln(1/(1-y))}\right)$ or $y = 1 - \exp(-e^{-x}/T)$

$x = \ln f$ and $y = v/V$ have been introduced.

When trying a Taylor series expansion of y in x around a certain value x_0 (corresponding to a centre frequency $f_0' = \exp(x_0)$), the derivatives of y with respect to x are of interest. Those of lowest order are:

$$dy/dx = -(1-y)\ln\left(\frac{1}{1-y}\right)$$

$$d^2y/dx^2 = \left[1 - \ln\left(\frac{1}{1-y}\right)\right] dy/dx$$

$$d^3y/dx^3 = \left[1 - \ln\left(\frac{1}{1-y}\right)\right]^2 dy/dx - \frac{1}{1-y}(dy/dx)^2$$

The condition for vanishing second-order term in the series is $1 - \ln(1/(1-y)) = 0$ and this gives $y_0 = 1 - e^{-1}$ and $x_0 = \ln f_0' = \ln(1/T)$ from which follows that the centre frequency f_0' should be chosen equal to $1/T$. A plot of y versus x verifies that this is an inflexion point and that the best fit to logarithmic behaviour (best linearity of plots on logarithmic paper) is obtained here. The value f_0' is twice the frequency f_0 for best linearity. As before, when $f_0' = 1/T$, the ratio of third-order and first-order terms in the series gives an estimate of the deviation from linearity. Here one finds $d^3y/dx^3 = dy/dx$. From this, as an example, one obtains a ratio of third-order and first-order terms of 1/12 for a frequency change with a factor $\simeq 2$ from the centre frequency f_0'. The corresponding frequency range thus extends from $f_0'/2$ to $2f_0'$.

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A Language Laboratory

Following an extensive evaluation programme with educational authorities into the problems of language teaching, W. H. Sanders (Electronics) Ltd has developed a language laboratory which is proving successful in the instruction and tuition of students. Orders for ten of these laboratories have already been received from teaching faculties.

Probably the most common and difficult problem confronting foreign language students is the actual pronunciation of words and phrases. Sanders have tackled this problem by producing a language laboratory which allows the student to compare his own pronunciation quickly and repeatedly with that of the lecturer by the use of the tape recorders. The basic idea is that a student may listen to the lecturer and record the passage, or to a pre-recorded passage, record his own version of the passage, play back both recordings and compare the two. The student can then re-record his own voice until satisfied. Each student is isolated from the rest of the class and is, therefore, able to learn at his own pace, and if guidance is required he can signal the lecturer.

A standard language laboratory consists of a lecturer's control console and between two and thirty-two student booths. The lecturer's console houses a complete tape recording system, a v.h.f. f.m. radio tuner and banks of switches and lamps for communicating with individual student booths.

These booths are of unit construction, heavily sound insulated from each other and fitted with twin tape recorder with combined headset-microphone. A 'call' switch and a 'wait' lamp enable the student to attract the lecturer's attention and the lecturer to acknowledge the call.

The students' booths are designed to be both quick and simple to maintain. Should a tape recorder be damaged as a result of misuse by a student, it can be replaced quickly by unskilled personnel without disturbing the remainder of the class. Maintenance of the Sanders Language Laboratories is undertaken by approved local service organizations.

In practice over many hundreds of hours of operation it has been found that the use of Sanders Language Laboratories has greatly increased the speed of attaining an acceptable standard of proficiency in a foreign language and particularly in pronunciation.

The method of operation is as follows: the lecturer records and edits a lesson and then transfers it to one track of the tape recorders installed in each student's booth. While the lesson is being transferred the students can hear it and rehearse the programme, this facility eliminating the element of surprise and resulting in a better appreciation of the lesson. The recording from the lecturer cannot be erased by the students who play it back and attempt to copy the lecturer's pronunciation. The lecturer can listen to each student individually and if necessary give assistance where there may be difficulty in the pronunciation of part of the lesson. Any student who requires help from the lecturer can attract his attention by operating his 'call' switch, and the lecturer, who may be answering another student's call, can send a 'wait' signal to the enquiring student until he is able to listen to his question. Alternatively the lecturer can address the class as a whole in order to explain a point in pronunciation etc. When speaking to the whole class, the lecturer automatically stops all the students' machines and so prevents anyone from missing part of the recorded lesson. By means of the v.h.f. f.m. radio tuner, use can be made of the language programmes currently being transmitted by the BBC. These programmes can be relayed to the students at the time of the broadcast or recorded on the control console tape recorder and then stored for future use.

It is possible to teach more than one language at once. Inputs are available for several programme sources, e.g. additional tape recorder etc., and a remote control slide projector can be fitted.

Telemetry Sampling Phenomena

By S. Poole*, D.L.C., B.Sc.(Eng.)

In this article various aspects of telemetry sampling are discussed. These are: (1) Bandwidth, (2) Aliasing, (3) A proposed 'search' scan. In general these are considered for the case where visual analysis of the sampled waveforms is required In addition to a theorectical discussion details of laboratory work carried out in connexion with the three topics are given.

(Voir page 352 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

A PREVIOUS article[1] has discussed the principles of time division multiplexing in telemetry systems with particular reference to ambiguities which may arise in the sampling process. It was there stated that two major design criteria in such systems are:

(a) The need to maintain band-limited conditions on all signal channels, i.e. by the use of suitable low-pass filters to ensure that the maximum value of signal frequency on any channel is always less than half the value of the sampling frequency for that channel.

(b) The ever-present need to conserve bandwidth.

It will generally be appreciated that the first of these two criteria has to do with meeting the requirements of the sampling theorem (Nyquist condition for minimum sampling rate). There is however a further condition which has to be faced whenever there is a requirement for the visual analysis of sampled data waveforms and that is that samples of any one channel must be taken at a rate much greater than two per cycle. Some authorities put the requirement as high as seven samples per cycle but this is very much a matter of opinion, varying between various systems and possibly capable of reduction in the light of experience in any one case. The discussion and suggestions of the previous article[1] were essentially concerned with those cases where such visual analysis was called for. It is felt that three topics in that article merit further consideration:

Bandwidth

It is necessary to establish exactly the bandwidth occupied by a sampled data waveform. Consider firstly one signal channel sampled by the unit amplitude pulse train as shown in Fig. 1. By classical Fourier analysis, this sampling pulse waveform may be written as

$$V_p = K\left[1 + 2\sum_{n=1}^{n=\infty}\frac{\sin n\pi K}{n\pi K}\,.\cos n\,\omega_r t\right]$$

where

$K = t_o/T$ and $\omega_r = 2\pi f_r$,

f_r = sampling frequency, $1/T$

This expression immediately indicates the resultant frequency spectrum of the pulse train. Assume a continuous signal variation about a mean value XV as shown. This signal may be written as

$$V_s = V\,[X + \cos\omega_s t] \text{ where } \omega_s = 2\pi f_s$$

The product $V_p . V_s$ may be regarded as a sufficient mathematical description of the sampling process since the resultant top-modulated pulse train is obtained by multiplying V_s by 1 or 0 at any instant according to V_p.

* *Lanchester College of Technology, Coventry.*

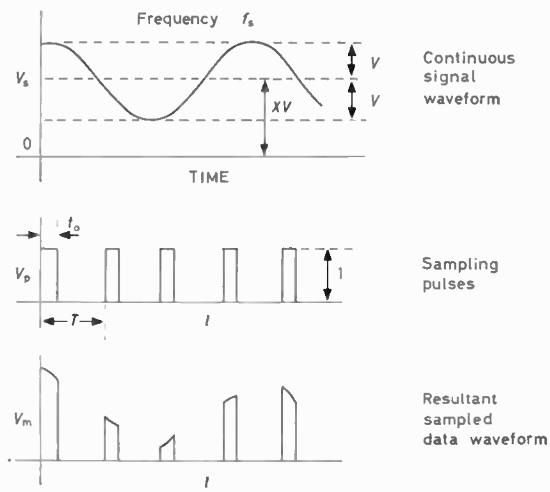


Fig. 1. Single channel sampling

$$\therefore V_m = KV\Big[X + \cos\omega_s t\Big]\left[1 + 2\sum_{n=1}^{n=\infty}\frac{\sin n\pi K}{n\pi K}\,.\cos n\,\omega_r t\right]$$

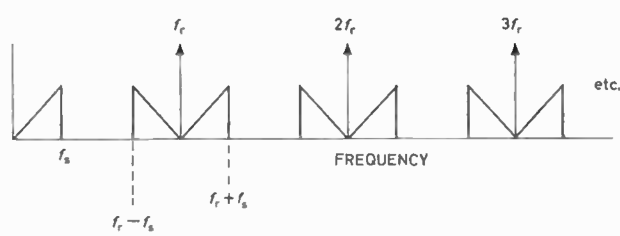
$$= KV\left[X + \cos\omega_s t + 2X\sum_{n=1}^{n=\infty}\frac{\sin n\pi K}{n\pi K}\,.\cos n\,\omega_r t\right.$$

$$+\sum_{n=1}^{n=\infty}\frac{\sin n\pi K}{n\pi K}\,.\cos(n\,\omega_r + \omega_s)t \;+$$

$$\left.\sum_{n=1}^{n=\infty}\frac{\sin n\pi K}{n\pi K}\,.\cos(n\,\omega_r - \omega_s)t\right] \quad ..\ (1)$$

The frequency spectrum indicated by this expression is as shown in Fig. 2. It consists of a component which is actually the original signal variation multiplied in amplitude by K, together with a d.c. term, and components at the sampling frequency and its harmonics plus associated sidebands. The process of de-modulation is easily achieved

Fig. 2. Frequency spectrum for Fig. 1



by passing the complete signal through a suitable low-pass filter which accepts f_s but rejects $(f_r - f_s)$ and above; if the filter output is then amplified by $1/K$, the original signal variation should be recovered without distortion. It may therefore be assumed that the bandwidth of the filter used represents the total bandwidth required for transmission of the signal information, all else above it being irrelevant—i.e. maximum bandwidth required is only up to the value of f_s.

Consider now the effect of other channels with amplitude modulated pulses occurring in the off-time of the one channel previously considered. This gives a complete time-division multiplex waveform as shown in Fig. 3 and it may be assumed that it is possible to reduce inter-channel gap time so much that it may be ignored. The on-time for any one channel is then T/S where S is the total number of channels. Allowing for identical signal bandwidths on each channel, say up to f_s, then the effect compared with

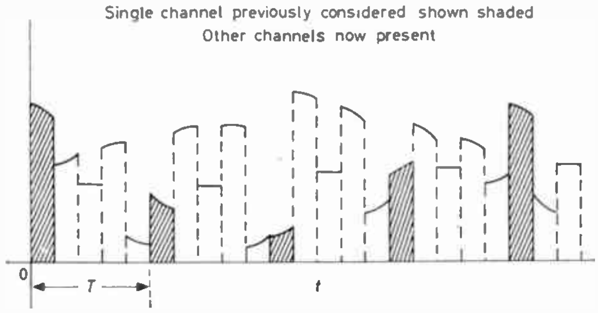


Fig. 3. Multi-channel t.d.m. waveform

the previous case of a single channel is that of now sampling a signal of frequency Sf_s at the same relative rate. Therefore the value of Sf_s, represents the minimum bandwidth required for the transmission of this multiplexed information, but note that individual channel outputs produced are in fact only correct at the instants of sampling.

Such generally accepted bandwidth requirements are however inadequate whenever visual analysis of such t.d.m. waveforms is called for. In addition to the particular requirements for sampling rate as discussed below, there is then a need for much greater system bandwidth if an adequate visual record is to be obtained for analysis purposes. This involves accepting much more of the type of frequency spectrum as shown in Fig. 2. The value of the highest harmonic to be included depends entirely upon the quality of record it is desired to achieve but the tendency is always towards the adequate reproduction of individual pulse shapes. The worst condition in the case of a multi-channel t.d.m. waveform would occur if consecutive channels were at opposite limits of their possible amplitude variations; the necessary bandwidth would then amount to that required for adequate reproduction of a square waveform of frequency $Sf_r/2$.

For the simple case of a single amplitude modulated pulse train, equation (1) indicates how the bandwidth depends upon both pulse duration and sampling rate. It is interesting to verify these facts in practice and compare measured values of the various frequency components with those predicted from theory. This may easily be done using the circuit of Fig. 4. This consists of a common emitter stage through which the signal is passed and which is switched on and off by a sampling pulse train applied to the input of the common-base connected transistor. Any convenient pulse generator may be used, provided that both pulse duration and rate are variable. Typical values used are pulse durations of 10, 25, and 50μsec at rates of 10 and 5kc/s.

With no signal input and sampling pulses applied of sufficient amplitude to drive the common-base transistor into saturation, the output consists of pulses of constant amplitude 4V (between −6 and −10V). The input signal is then applied and increased in amplitude until the measured pulse amplitude variation is ± 1V about the unmodulated value as measured at the output by c.r.o. Using the same abbreviations as in the above theory, this means that $V = 1\text{V}$ and $XV = 4\text{V}$.

∴ r.m.s. amplitudes of f_r and harmonics

$$= (1/\sqrt{2}) \cdot \frac{2KXV}{n\pi K} \sin n\pi K$$

$$= \frac{5{\cdot}66}{n\pi} \sin n\pi K$$

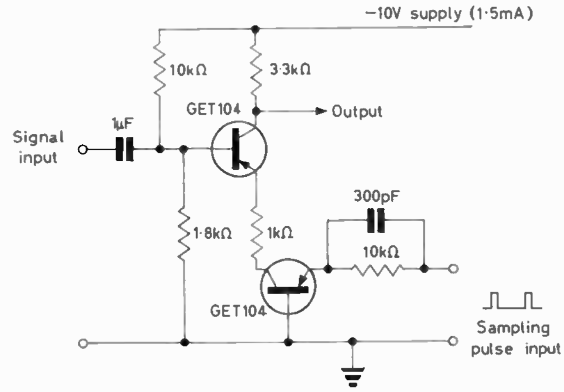


Fig. 4. Transistorized sampling circuit

also r.m.s. amplitudes of sideband frequencies about the above

$$= (1/\sqrt{2}) \cdot \frac{KV}{n\pi K} \sin n\pi K$$

$$= \frac{1}{\sqrt{2} \cdot n\pi} \sin n\pi K$$

Hence, the amplitudes of all terms in the frequency spectrum may be calculated. They may also be measured if a frequency selective detector of adequate performance is available—e.g. Siemens level meter (Pegelmesser) type Rel. 3D 332. Using this apparatus, very close agreement has been found between measured and calculated values with $f_s = 3\text{kc/s}$ for various combinations of t_0 and f_r as listed above. It is particularly interesting to note how the component amplitudes depend upon the term, $\sin n\pi K$; these amplitudes therefore go to zero whenever $nK =$ 1,2,3,4 etc.

The result indicated by expression (1) strictly only applies for the case of p.a.m. with 'top' modulation. This is also implied in Figs. 1 and 3, and telemetry waveforms are often of this type. However, there is sometimes the requirement for a flat-topped form of p.a.m., i.e. where each pulse amplitude is constant at a value governed by the signal at the instant of sampling (beginning of pulse). This is particularly so when p.a.m. is a pre-requisite to other forms of pulse modulation. It will be realized that if t_0 is decreased, the samples considered in the previous analysis will approach closer and closer to being flat-topped pulses. Actual flat top pulses may be regarded as a summation of a large number of trains of such differentially short pulses side by side; the summation must be taken between such limits as required to obtain the desired pulse width. An excellent treatment along these lines is given by Black[2] and it may be shown that the resultant frequency spectrum may be written as

$$V_m = KV\left[X + \frac{2}{\omega_s t_o}\sin\omega_s(t_o/2).\cos\omega_s t + \sum_{n=1}^{n=\infty} 2X.\frac{2}{n\omega_r t_o}\sin n\omega_r(t_o/2).\cos n\omega_r t + \sum_{n=1}^{n=\infty}\frac{2}{(n\omega_r \pm \omega_s)t_o}\sin(n\omega_r \pm \omega_s)(t_o/2).\cos(n\omega_r \pm \omega_s)t\right] \quad (2)$$

where all + signs go together and all − signs go together. This expression applies to a single pulse train with flat-top modulation. It may also apply in cases where individual channel pulses after de-multiplexing are 'stretched', i.e. pulse width artificially increased to raise the signal power level. Equation (2) clearly indicates that this will result in amplitude distortion since the signal frequency term, which may again be separated out by a suitable filter, is now multiplied by:

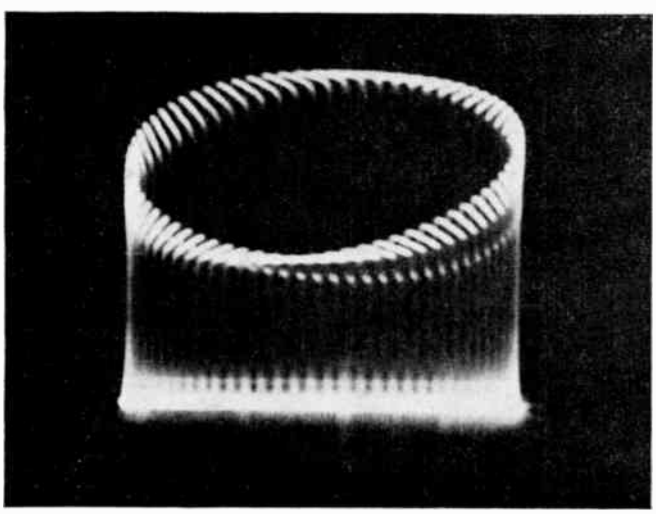
$$\frac{\sin \pi K(\omega_s/\omega_r)}{\pi K(\omega_s/\omega_r)}$$

The amount of distortion clearly depends upon both sampling rate and final pulse width, and may be corrected by the use of a suitable equalizer. The introduction of this distortion in the case of pulse stretching is sometimes referred to as the 'aperture effect'.

Aliasing

This is the name given to the effect which is caused in sampled data systems when the sampling theorem is violated. It must be realized that a train of amplitude modulated pulses may be interpreted in more than one way; the usual envelope shape which may be constructed over such a train is usually that of the lowest possible frequency and this will give a picture of the true signal only if the sampling theorem is not violated. If however the sampling rate is less than twice the signal frequency, then the lowest frequency envelope shape will not give the true signal, (f_s), but a false signal, ($f_r - f_s$). It is seen from Fig. 2 that $(f_r - f_s) < f_s$ whenever $f_r \not> 2f_s$. Any attempt to isolate the f_s term by a low-pass filter must also include $(f_r - f_s)$ as an error term, thus causing distortion. Stiltz[3] gives a full treatment of the magnitude of such errors as related to the amount of signal spectrum violating the sampling theorem. It is the resulting downward spectral transposition of signal energy which is referred to as 'aliasing error'. Stiltz also gives a simple diagrammatic representation of how a p.a.m. train may be variously interpreted and it is felt worthwhile to have a simple demonstration of this effect. This may be done with the circuit of Fig. 4—for example, set to give 10μsec pulses at 10kc/s, amplitude modulated by a signal at 6kc/s; the sampling theorem is thus violated. The modulated pulse waveform is applied to the Y plates of a c.r.o. and a separate variable frequency (f_x) signal applied to the X plates. It is found that single loop Lissajous figures may be formed 'on top of' the pulses as shown in Fig. 5 when f_x is set to 4kc/s ($f_r - f_s$), 6 kc/s (true f_s), 16kc/s ($f_r + f_s$), etc. This clearly demonstrates how the modulation may be variously interpreted—correctly and incorrectly. An incorrect interpretation is of most consequence in a case like this when ($f_r - f_s$) is $< f_s$.

Fig. 5. Lissajous display on c.r.o.



Proposal for a 'Search' Frame Scan

As already stated, it is necessary for signals to be sampled at a much faster rate than 2 samples per cycle, whenever a visual analysis of the sampled waveform is necessary. If the sampling is periodic, i.e. the sampling sequence of channels in each frame (between successive sync pulses) remains unaltered, then it would be necessary to increase the sampling rate in such cases. This may be achieved either by increasing the speed of the sampling device or reducing the channel capacity such that any one channel may be sampled more than once per frame. In either case the result is undesirable, leading to increased bandwidth and reduced system information capacity.

The proposal outlined previously[1] is that the sampling should be non-periodic. If the channel sampling sequence is varied from frame to frame so as to explore the signal waveform over, say, six frame scans, then it is hoped to obtain adequate records for visual analysis with an effective sampling rate much less than would otherwise be required. At first sight this is a very attractive proposition and is theoretically in agreement with a statement quoted by Black[2] as a generalized form of the sampling principle. This states, in effect, that provided $f_r > 2f_s$, a channel may be sampled once per frame in any manner and a knowledge of the magnitude of each sample plus a knowledge of the instant within each frame at which the sample was taken contains all of the information of the original signal. Unfortunately however, the practical out-working of this when faced with the requirement for visual interpretation of waveforms is not as good as had been expected.

Consider a six channel sampling system. Then if the sampling is periodic and allowing for one sync channel between frames, there will always be a time gap between successive samples of any one channel equal to the time occupied by six sampling periods. For convenience and reference this gap time will be denoted as being 6 units.

Consider now the sampling sequence to be as shown in Table 1. It is seen by inspection that this forward movement in time of the sampling instants results in gap times between successive samples on any one channel working out at four of 7 units, one of 1 unit, and one of 8 units. This would appear to worsen the position compared with that for periodic sampling since the majority of gap times are now lengthened and the one case of a very short gap time is of little value.

If however it is arranged that there is a backward movement in time of the sampling instants, then the sequence may be as shown in Table 2. In this case, it is seen that

TABLE 1
Non-periodic Sampling, Forward Shift
S is sync channel; other channels as numbered

S	1	2	3	4	5	6
S	6	1	2	3	4	5
S	5	6	1	2	3	4
S	4	5	6	1	2	3
S	3	4	5	6	1	2
S	2	3	4	5	6	1

TABLE 2
Non-periodic Sampling, Backward Shift

S	1	2	3	4	5	6
S	2	3	4	5	6	1
S	3	4	5	6	1	2
S	4	5	6	1	2	3
S	5	6	1	2	3	4
S	6	1	2	3	4	5

the gap times between successive samples on any one channel work out at five of 5 units, and one of 11 units. It may be reasonably expected that this represents an improvement of the position since the majority of gap times are reduced; the one large gap appearing once in the sequence would however seem to be a serious drawback.

Records have recently been taken by the author of a six channel system, firstly with periodic sampling and then with the sequencies of Tables 1 and 2. From a critical examination of these records, the variations introduced into the sampling sequence seem to make little difference to the overall position, the value of minimum sampling rate required apparently remaining unaltered. Nevertheless, the change from periodic sampling to the sequence of Table 2 appears to be a move in the right direction; it is only prevented from yielding a significant improvement by the large gap time of 11 units occurring with every sixth sample. There is every reason to hope that the position would be improved in a system having more channels—for example a 20 channel system, so that with the proposed variation in sampling sequence, the signal waveform is explored over 20 frame scans rather than 6 as stated in the original proposal. This effect with such a 20 channel system needs further investigation; it may however be stated at this stage that the shift in sampling instants must be backward rather than forward in time, i.e. as Table 2 rather than Table 1. Assuming the 'success' of this proposed search frame scan, it is doubtful whether the sampling rate requirement for visual work could ever be reduced to the order of the Nyquist frequency. A rate of, say 4 to 5 samples per cycle of signal is perhaps the best one may hope for; even this would be a worthwhile improvement.

Acknowledgments

The author wishes to acknowledge the encouragement and advice received from Mr. R. E. Young. Thanks are also due to Mr. R. Barrett, Senior Lecturer at the Lanchester College, for many stimulating discussions during the preparation of this article.

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Magnetoplasmadynamic Direct Generation of Electricity

International Research & Development Co Ltd (IRD), Newcastle upon Tyne, England, have announced a major technical advance in the field of direct electric power. Following three years' investigation and development, IRD have extracted electricity from a closed cycle magnetoplasmadynamic (m.p.d.) generator in which the only moving part is a high temperature gas. To produce electricity the gas is forced at a high velocity through a magnetic field. This is claimed to be the first m.p.d. generator operating on the closed cycle principle (in which the working gas is continuously circulated) to produce significant output power, thus giving Britain a clear lead over the United States and Europe.

Experiments on the open cycle system, eventually based on the combustion of coal, oil or gas, have been successfully operated in the United States and in Britain. These systems require extremely high temperatures (in the region of 3 000°C), which present formidable technical problems, and such devices have only operated for relatively short times. The closed cycle system, in which heat would be provided by a nuclear reactor, operates under highly pure conditions, so that temperature requirements are lower. The recent work at IRD has demonstrated that m.p.d. power can be obtained at temperatures below 1 800°C, a level which may be achieved in advanced types of nuclear reactor.

Substantial financial support for the research programme came from IRD's parent company, C. A. Parsons & Co. Ltd, and later partly from the Advanced Research Projects Agency of the US Department of Defense. The objectives were: to demonstrate the technological feasibility of an m.p.d. direct power generator in a loop closely simulating an ultimate nuclear reactor closed cycle system; and to establish an understanding of the fundamental processes of power extraction by the m.p.d. technique.

The IRD closed cycle m.p.d. loop operates with a mixture of pure helium and caesium. Helium is a non-reactive gas which is compatible with materials at very high temperatures and is a likely choice for the coolant in a very high temperature nuclear reactor. The loop simulates very closely the cycle of operations which would be adopted in a large-scale plant. The helium is compressed and passed through a recuperative heat exchanger to a graphite electrical heater (simulating the nuclear reactor core) which produces gas temperatures up to 2 200°C. Following the heater a small proportion of caesium vapour is introduced (this process is known as 'seeding') which converts the gas to an electrically-conducting 'plasma'. The caesium-seeded plasma is then accelerated to a high velocity and passed through the m.p.d. generator which consists simply of a rectangular-section duct, measuring 0·5in by 1·5in by 6 in and containing five pairs of 0·5in by 0·5in tantalum metal electrodes in the two narrow walls. The walls are fabricated from high density, high purity, aluminium oxide which acts as an electrical insulator at the high operating temperature. Power is extracted from the electrodes and the gas at a lower temperature enters a diffuser, passes over the recuperative heat exchanger to a final cooler, and returns to the compressor.

In these initial experiments the loop was operated continuously for fifteen hours with gas temperatures up to 1 800°C. Twenty series of measurements were made, with varying temperature, caesium content and applied magnetic field, with maximum power outputs of ½W. Even with the present small size of generator duct it is believed that a power output as high as 3kW can be obtained with further development.

Tungsten, tantalum, molybdenum and tantalum-tungsten alloy are employed extensively in the construction of the IRD loop. These metals can operate in contact with helium at very high temperatures but deteriorate rapidly if there is any oxidizing gas present. Consequently, the circulating helium in the loop is continuously purified to maintain levels less than a total of 10 parts per million of other gases in the helium.

The IRD loop has been operated with helium for several hundred hours at temperatures up to 2 200°C and this achievement has necessitated a considerable range of supporting work on the chemical and materials problems and constructional techniques for these exceptional conditions.

A True R.M.S. Voltage Measuring Circuit

By J. C. Cluley*, M.Sc., A.M.I.E.E.

The conventional mean-reading rectifier circuit used in automatic voltage regulators for alternators may be responsible for a considerable error in output voltage if the machine waveform contains harmonic components. The magnitude of this error and its dependance upon the amplitude and frequency of the harmonic is discussed, and a circuit which can produce a direct voltage accurately proportional to the mean square alternator voltage is described. The circuit avoids the time lag associated with mechanical devices or temperature sensitive elements, and uses a recently developed cadmium sulphide dielectric diode which possesses a square-law characteristic.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

ALTHOUGH most industrialized countries now have a fairly extensive public electricity supply network, there is an increasing demand for small alternators where a portable supply is required, where there is no convenient public supply, or where the supply is subject to interruptions. Typical examples of these situations are aircraft maintenance on a large airfield, television outside broadcasting units, microwave or telephone repeaters in remote areas, and the standby generators used in case of supply failure in telecommunications centres, large hospitals, etc.

In most instances the load to be supplied includes electronic apparatus, frequently containing transistors, so that it is essential to maintain the required output voltage accurately, and to reduce the magnitude and duration of the voltage transients caused by changes in load current to acceptably small values. Consequently most users now demand a rapid and accurate electronic voltage regulator which does not involve moving parts with their unavoidable mechanical inertia. This demand is satisfied by voltage regulators using power transistors, silicon controlled rectifiers, or high speed magnetic amplifiers as the controlled element, and purely mechanical devices such as the carbon-pile regulator are generally relegated to less critical applications.

The Reference Element

Unfortunately, despite the greater accuracy and speed of response of the electronic regulator compared with its electromechanical counterpart, its performance may be inferior when the supply waveform contains an appreciable proportion of harmonics. The defect arises from the nature of the reference element in the regulator.

In the electromechanical regulator the reference is the force in a spring or a flexible diaphragm which is balanced by the force on the armature of an electromagnet. Since the movement of armature requires a fairly large air-gap, the iron circuit is generally unsaturated and the mean force on the armature is proportional to the mean square voltage applied to the coil circuit. The regulator consequently senses the mean square voltage and will operate to maintain this quantity, and thus the r.m.s. voltage, as near as possible to the desired value.

However, in an electronic regulator, the reference is usually a constant voltage element such as a Zener diode or a gas-filled stabilizer tube, and the alternator output must be rectified and smoothed before comparison with the reference voltage as shown in Fig. 1.

The Rectifier

In an ideal system the rectifier would deliver a mean output signal proportional to the r.m.s. value of alternator line voltage, but in practice the only simple rectifier arrangements are peak-reading or average-reading circuits. The peak-reading circuit should be lightly loaded, and is particularly sensitive to waveform distortion caused by harmonics. Since the output power of the amplifier of Fig. 1 is determined by the field excitation of the generator, the design of the amplifier can be simplified by operating the error detector at a high power level, and so reducing the power gain required. The detector power level is generally limited by the permissible dissipation of the reference element, of the order of 100mW, but this is frequently obtained through a series resistor from the rectifier output, as shown in Fig. 1. The rectifier may thus be required to supply both the sample and the reference inputs to the error detector, making the mean-reading circuit a more suitable arrangement than the peak-reading circuit on account of the considerable power required. Since the mean-reading circuit is also less influenced by harmonics it is almost invariably used.

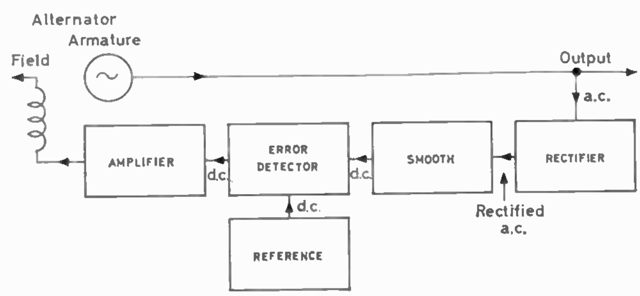


Fig. 1. Arrangement of the voltage regulator

The unavoidable error in the system of Fig. 1, even if the loop gain is very high, arises from the differing influence of alternator harmonics on the r.m.s. indicator by which the user measures its performance, and the average-reading rectifier by which the regulator senses the output voltage. Of course, if the user can be persuaded to measure the regulation using a rectifier voltmeter (perhaps by fitting one to the control panel) no problem arises, but

* *University of Birmingham*

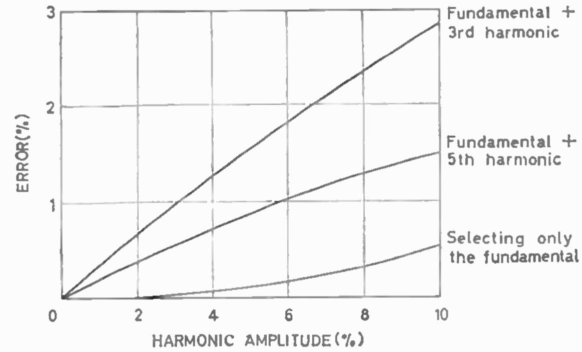


Fig. 2. Measurement error using mean value as a measure of r.m.s. value

written specifications almost invariably mention only r.m.s. values.

Errors Caused By Harmonic Voltages

The magnitude of the error which may arise depends upon the order of the harmonic and its phase relative to the fundamental, as shown in Appendix 1. The largest error arises when the harmonic and the fundamental cross zero voltage at the same instant; this generally occurs at no-load, but at full-load the harmonics in the alternator terminal voltage may experience some phase shift relative to the fundamental. This will arise unless the load is purely inductive, since the harmonics and the fundamental will suffer differing phase shifts due to the alternator reactance when the alternator is supplying power to the load. Fig. 2 shows the error caused by taking the mean value of the alternator voltage as a measure of the r.m.s. value, when various amplitudes of third and fifth harmonic are present separately. These are computed from the following expressions:

$$\text{r.m.s. value} = \tfrac{1}{2}\sqrt{(V_1^2 + V_3^2 + V_5^2 + \ldots)}$$

$$\text{mean value} = 2/\pi \left\{ V_1 + \frac{V_3}{3} + \frac{V_5}{5} + \ldots \right\}$$

where the alternator voltage is represented by

$$v = V_1 \sin\theta + V_3 \sin 3\theta + V_5 \sin 5\theta + \ldots;$$

assuming zero phase shift between fundamental and harmonics.

The maximum harmonic content which may be expected with linear loading is of the order of 10 per cent for machines built to B.S.S. 2613:1955. This permits the machine voltage waveform to deviate as much as 10 per cent from the 'equivalent sine wave' which most nearly fits the actual waveform. If there is only one dominant harmonic, its amplitude may thus be nearly 10 per cent of the fundamental amplitude.

The worst case occurs in single-phase alternators, or in loads connected between phase and neutral in a three-phase star-connected alternator, when the dominant harmonic will be the third, and the error caused by an average-reading rectifier may rise to nearly 3 per cent. It is usual to find other harmonics accompanying the third harmonic, so that for a total waveform deviation of 10 per cent the third harmonic content will be somewhat less, as will the resulting rectifier error. Thus an alternator giving 7 per cent of 3rd harmonic and 3 per cent of 5th harmonic would cause an error of only 2·38 per cent.

The Effect of Load Changes and Non-Linearity

The performance of a voltage regulator however, depends upon the change in harmonic content between no-load and full load, not upon the magnitude of the harmonics, since if the content were invariant with load, it would be possible to adjust the reference signal to compensate for the constant error in the rectifier circuit. Unfortunately this change in harmonic content is not mentioned in alternator specifications, but tests on several small machines indicate that the third harmonic component may be several times greater at full-load than at no-load, and that the fifth harmonic does not increase as much.

A further complication arises when a substantial part of the load consists of electronic apparatus which is supplied through rectifying and smoothing circuits. These impose grossly non-linear loads on the alternator, and thus usually increase the harmonic content of the alternator waveform. The total harmonic content in such a case may be well above the level specified in B.S.S. 2613, which does not mention non-linear loading, and the error caused by an average-reading rectifier in the voltage regulator may thus considerably exceed that shown in Fig. 2. It is clear that the influence of harmonics must be given careful study if the regulator is required to control the output voltage within say, ±1 per cent.

Methods of Reducing Errors in the Rectifier

One improvement which could be used would be a filter placed between the alternator output and the rectifier to attenuate the harmonics. If this transmitted only the fundamental the error caused by a 10 per cent harmonic would be reduced to only 0·5 per cent. However at 50c/s this would require large, heavy and expensive chokes, since they would require air-gaps to avoid saturation and the consequent generation of further harmonics. The filter would also introduce a further time delay into the closed-loop system and make it more liable to instability. A better method would be to use a true r.m.s. sensing circuit, which would give a mean output proportional to the mean square or root mean square of the alternator waveform. Approximate methods based upon linear segment approximations have been used previously, but these are either somewhat inaccurate, or require many segments and are thus complicated and expensive.

Thermal devices for r.m.s. measurement are more accurate but their calibration changes with ambient temperature, and they are slow to respond.

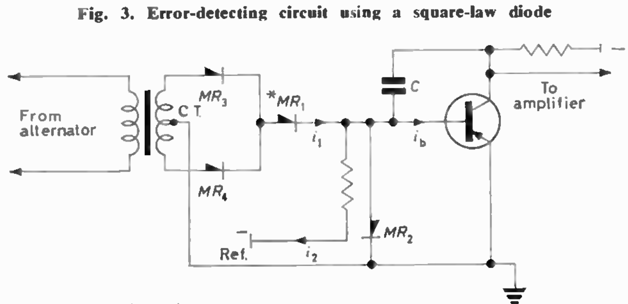
A Square-Law Diode Circuit

A much simpler method using a recently developed square-law diode[1] is shown in Fig. 3. Early diodes had a voltage threshold, with a characteristic of the form

$$i = k(v - v_0)^2 \text{ for } v > v_0$$

where i and v are the diode current and voltage respec-

Fig. 3. Error-detecting circuit using a square-law diode



tively, but further development in the Materials Group of this Department has produced a diode in which V_o is negligible, and the current is thus accurately proportional to the square of the voltage applied[2]. The equilibrium condition for the transistor is arranged so that it is partly conducting, with a base current of 50 to 100μA. The mean current $i_{1(mn)}$ in the square-law diode is balanced by the current i_2 from the reference source, the difference being the base current i_b. Thus

$$i_{1(mn)} - i_2 = i_b$$

Now in a typical square-law diode a mean current of 10mA is permissible, and if the amplifier has sufficient gain to deliver full field excitation for a 50 per cent change in the initial transistor base current of 50μA, the error involved in writing

$i_{1(mn)} = i_2$ is only 25μA in 10mA or $\frac{1}{4}$ per cent.

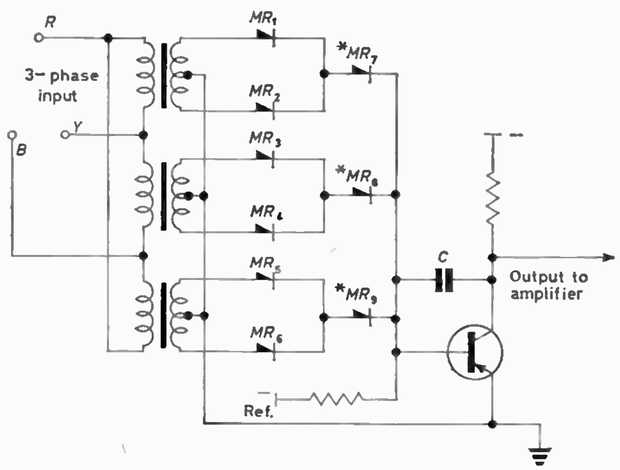


Fig. 4. Three-phase rectifier and error-detector

The regulator will thus adjust the field current to maintain the mean current through the square law diode practically equal to the constant reference current, and so provide a constant mean square output voltage, as required.

In the mean-square sensing circuit of Fig. 3 it is essential that the square law diode MR_1 should receive the unsmoothed rectified supply from the alternator output through a low impedance transformer, and that the potential variations at the base of the transistor should be small compared with the voltage across MR_1. These conditions ensure that the diode voltage will be an accurate sample of the alternator voltage. The second condition is in effect a stipulation of the minimum voltage gain of the error detector and amplifier together. Assuming the figure quoted above for the base current in the error-detecting transistor, the 25μA change in base current required to give full field excitation will correspond to a change in base voltage of 25mV for the typical input impedance of 1kΩ. This is only $\frac{1}{4}$ per cent of the 10V signal which can be handled by a typical square law diode, and will not cause a significant error.

The smoothing of the rectifier current is here performed by the capacitor C connected as a feedback integrator. The value of capacitance used is subject to conflicting requirements; the need to smooth the diode current adequately to prevent the remaining fluctuations from saturating a high gain regulator requires a large capacitor whereas the need for a quick response to load fluctuations requires a minimum time delay and so a small capacitor.

For high frequency machines the longest time-constant in the regulator is almost invariably that of the alternator field circuit, and the smoothing of the measuring circuit presents little difficulty. However for small 50c/s alternators for which accurate regulators are specified, the time-constant of the smoothing circuit may well be the main factor in determining the response time. In such cases the use of a three-phase rectifier system becomes an advantage, since this permits a substantial reduction in the time-constant of the smoothing circuit for the same ripple voltage. The consequent increase in the speed of response of the smoothing circuit may even justify the inclusion of a single-phase to three-phase conversion circuit between the alternator and the rectifier, although this contains reactive elements and must be designed so as not to accentuate the harmonic components.

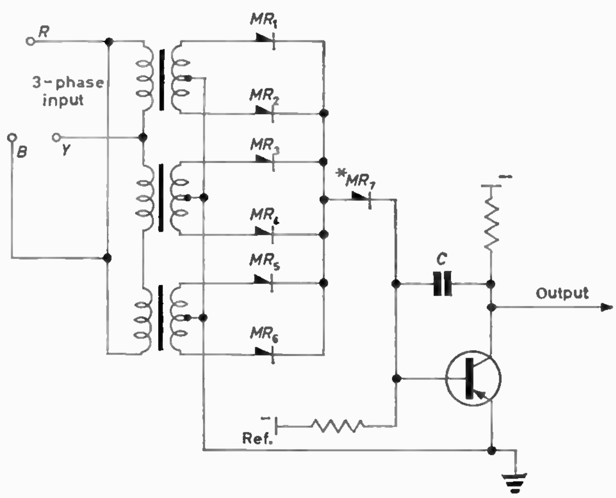


Fig. 5. Three-phase rectifier and error-detector using a single square-law diode

Three-Phase Circuits

The diagram of Fig. 4 shows a three-phase version of Fig. 3 which requires three matched square-law diodes which have the same constant k in the characteristic

$$i = k.V^2$$

The three input transformers can then be identical. If matched diodes are not readily available, it would be possible to achieve the same result by adjusting the transformer ratios to give the same peak diode current in each phase for a particular line voltage.

A simple circuit is shown in Fig. 5 which requires only one square law diode, but since this feeds only the peaks of the waveform to the square law diode it is inaccurate. Calculations given in the appendix show that the addition of 10 per cent third harmonic on all phases will cause a 6·9 per cent change in the root mean diode current, whereas the r.m.s. voltage has changed by only 0·5 per cent. If the three full-wave rectifier circuits are connected in series, in order to give some weight to the parts of the waveform near the axis, the combined waveform presented to the square-law diode is not substantially altered, and the error is in fact rather larger. The root mean diode current in this case falls by about 7 per cent when a 10 per cent harmonic is added to each phase, giving an error of about 7·5 per cent. Both circuits using only one square-law diode thus give results inferior to that given by the mean-reading rectifier circuit and it appears necessary to use three square-law diodes in the circuit of Fig. 4 to obtain a true r.m.s. sensing unit.

The use of Thin-Film Techniques

The first square-law diodes were prepared by depositing cathode and anode electrodes on a CdS crystal, but recent work on the use of thin vacuum-vapour deposited structures[3] is very promising. This technique is very convenient for the fabrication of a diode structure with three anodes and a common cathode, as required for a three-phase measuring unit.

It might be possible either to control the anode area during deposition, or to follow this process by a selective etching or abrading, so as to match the three diode characteristics. Also, since the vapour deposition process is an inexpensive one suitable for mass production, it might be possible to produce six matched diodes on a common cathode, and so dispense with the six semiconductor diodes MR_1 to MR_6 of Fig. 4.

Conclusion

The generally used mean-reading rectifier circuit has been shown to cause considerable errors when used in an accurate voltage regulating system subject to appreciable alternator harmonics. The error can be reduced by careful filtering of the waveform, but this is a difficult operation likely to add considerably to the time-lag of the system. A simpler and more accurate circuit which measures the true r.m.s. value of the alternator voltage may be constructed by interposing a square law diode between the rectifier circuit and the virtual earth of a feedback integrator smoothing circuit. This circuit may also be used in polyphase systems with slight modifications.

Appendix

(1) The output of a mean-reading rectifier with harmonics present

Let the alternator output voltage be

$$v = V_1 \sin\theta + V_9 \sin 3\theta + V_5 \sin 5\theta$$

Then the mean rectifier voltage, neglecting rectifier voltage drop, will be

$$V_{(mn)} = 1/\pi \int_0^\pi (V_1 \sin\theta + V_3 \sin 3\theta + V_5 \sin 5\theta) d\theta$$

$$= 1/\pi \left[-V_1\cos\theta - V_3 . \cos3\theta/3 - V_5 \cos 5\theta/5 \right]_0^\pi$$

$$= 1/\pi . \{2V_1 + 2/3V_3 + 2/5V_5\}$$

$$= 2/\pi \{V_1 + (V_3/3) + (V_5/5)\}$$

The contribution of the harmonic components to $V_{(mn)}$ is maximum when, as here, there is no phase shift between them and the fundamental. If the third harmonic only is present of a magnitude small enough not to alter the zero crossing angle appreciably, the mean output signal becomes

$$V = 1/\pi \int_0^\pi \{V_1 \sin\theta + V_3 \sin(3\theta + \alpha)\} d\theta$$

$$= 1/\pi [-V_1\cos\theta - (V_3/3)\cos(3\theta + \alpha)]_0^\pi$$

$$= 2/\pi \{V_1 + (V_3/3)\cos\alpha\}$$

Thus for a phase shift of $\alpha = 90°$ the contribution of the third harmonic to the rectified current vanishes. If the harmonic amplitude is sufficient to alter the zero crossing angles of the alternator waveform from 0° and 180° on the fundamental component the above result is incorrect, but the general influence of the phase shift α remains the same, and the maximum contribution is still obtained for $\alpha = 0°$.

(2) The Error in Estimating the r.m.s. Value Using the Circuit of Fig. 4.

The alternator output voltage is assumed to consist only of fundamental and third harmonic, represented by

$$= V_1 \cos\theta + V_3 \cos 3\theta$$

The conduction angle considered is that between +30° and −30° and with the postulated condition of zero phase shift between fundamental and third harmonics, the third harmonic makes zero contribution to the voltage at $\theta = \pm 30°$. Thus the instants at which the rectifiers commutate is not influenced by the presence of the harmonic. Since the voltage is symmetrical about $\theta = 0°$, the mean diode current is proportional to

$$V^2 = 6/\pi \int_0^{\pi/6} (V_1\cos\theta + V_3 \cos 3\theta)^2 d\theta$$

$$= 6/\pi [(V_1^2/2)(\theta + (\sin 2\theta/2)) + (V_3^2/2)(\theta + (\sin 6\theta/6)) + V_1V_3((\sin 4\theta/4) + (\sin 2\theta/2))]_0^{\pi/6}$$

$$= (V_1^2/2)(1{\cdot}828) + (V_3^2/2) + 1{\cdot}24V_1V_3$$

If $\alpha = V_3/V$ is the proportion of third harmonic, this is proportional to

$V_1^2/2\,(1 + 1{\cdot}36\,\alpha + 0{\cdot}547\,\alpha^2)$, where a true mean square rectifier should have an output

$$V_1^2/2\,(1 + \alpha^2).$$

The r.m.s. alternator voltage is thus incorrectly measured as

$$V_1/2\surd\,(1 + 1{\cdot}36\,\alpha + 0{\cdot}547\,\alpha^2)$$

instead of $V_1/\surd 2\,(1 + \alpha^2)$

For a 10 per cent third harmonic $\alpha = 1/10$, and the circuit reads $V_1/V\surd 2$ (1·069) instead of $V_1/\surd 2$ (1·005), i.e. an error of 6·4 per cent.

Series connexion of rectifiers

In this case the voltage applied to the square-law diode is $V = V_1\cos\theta + V_3\cos 3\theta + V_1\cos(\theta + 60°) + V_3\cos(3\theta + 180°)$

$$+ V_1\cos(\theta - 60°) + V_3\cos(3\theta - 180°)$$

$$= 2V_1\cos\theta - V_3\cos 3\theta.$$

As before the mean diode current is proportional to

$$V^2 = 6/\pi \int_0^{\pi/6} (2V_1\cos\theta - V_3\cos 3\theta)^2 d\theta$$

$$= 2V^2(1{\cdot}828) + V_3/2 - 4{\cdot}97V_1V_3$$

This is proportional to

$$V_1^2/2[1 - 1{\cdot}36\,\alpha + 0{\cdot}137\,\alpha^2], \text{ where}$$

$$\alpha = V_3/V_1 \text{ as before.}$$

Thus the alternator voltage is again incorrectly measured as

$$V_1/\surd 2\ \surd(1 - 1{\cdot}36\,\alpha + 0{\cdot}137\,\alpha^2) \text{ instead of}$$

$$V_1/\surd 2\ \surd(1 + \alpha^2)$$

For a 10 per cent harmonic component, the circuit reads $V_1/\surd 2$ (0·93) instead of $V_1/\surd 2$ (1·005) i.e. an error in the opposite direction of 7·5 per cent.

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Transistor Low Drift D.C. Amplifiers

By James J. Pinto*, M.Sc.

This article describes the variations of the d.c. parameters of transistors, important in the design of low drift d.c. amplifiers. The conventional 'long-tailed pair' and the modifications necessary for improved performance are described and the results assessed. A new, balanced, 'mirror-image' amplifier is then described and its performance and capabilities are discussed. Finally, the latest trends in direct coupled amplifier design and development are indicated.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE three main parameters of a transistor that vary with temperature and limit the performance of the device in direct coupled amplifiers are:

(1) The forward voltage drop between base and emitter, V_{be}

(2) The d.c. forward transfer ratio, h_{fe}

(3) The collector leakage current, I_{co}.

BASE-EMITTER VOLTAGE—V_{be}

The base emitter voltage of a transistor is given approximately by the expression:

$$V_{be} = v_{be} + I_b r_b + I_e r_e$$

where V_{be} = actual base-emitter voltage
v_{be} = ideal base-emitter voltage
I_b = base current
r_b = base resistance
I_e = emitter current
r_e = emitter resistance.

The ideal base-emitter voltage v_{be} has a negative temperature coefficient of approximately $-2{\cdot}5$mV/°C. The actual base-emitter voltage V_{be} is due, in addition to v_{be}, to the bulk resistance in series with the base-emitter diode and this bulk resistance has a positive temperature coefficient. This resistance is not closely defined and therefore the actual V_{be} and its temperature coefficient are also not closely defined. The temperature coefficient of V_{be} is obviously dependent on the current, being close to the ideal at low currents and becoming less negative and less well defined as I_b and I_e increase. In the planar method of transistor construction, resistivities are more uniform and consequently V_{be} and temperature coefficients of V_{be} are more uniform.

CURRENT GAIN—h_{fe}

The d.c. gain of a transistor depends on various factors such as the type of material, doping levels, type of construction, base width, etc., and it is difficult to arrive at an expression for its temperature dependence. In general, h_{fe} varies at the rate of approximately 0·1 to 0·5 per cent/°C, increasing with increase of temperature.

Another feature, important in the design of d.c. amplifiers, is the variation of h_{fe} with collector current. This is explained as follows[1]: At very low currents, the recombination of electrons and holes in the emitter depletion layer is high compared with the current and the gain is therefore low; as the current increases the recombination is constant, causing the effective gain to increase; at higher currents an increasing electric field develops in the base region and accelerates the minority carriers towards the collector, increasing the gain even more; as the current is increased still further, the high minority carrier density causes an increase in base conductivity, lowering the efficiency and decreasing the gain. In the planar method of construction the gain remains high at collector currents down to the microampere region.

LEAKAGE CURRENT—I_{co}

The leakage current in a transistor is generated in four ways[2]. In the base region at any temperature, a number of inter-atomic bonds break up spontaneously into hole-electron pairs, the actual number depending on the material and temperature, and not on voltage. This causes part of the observed leakage current. A second, similar component is generated in the collector depletion region and since the volume of this region is a function of material resistivity, voltage and junction area, this component is dependent on the method of construction and the voltage, in addition to temperature. A third component of I_{co}, called the surface thermal component, is generated by surface energy states and, like the base region component, depends only on temperature. Surface imperfections, impurities and moisture contribute a fourth, resistive component which is relatively independent of temperature and dependent on voltage. Only the last component of leakage current (and to a certain extent the second) can be minimized by improved construction techniques. In the planar method, junctions are protected by an oxide mask, resulting in extremely low resistive leakage currents.

The leakage current in silicon is very small and the changes in collector currents due to changes in leakage currents are usually small enough to be neglected when silicon transistors are used.

General Considerations

Drift in direct coupled amplifiers may be grouped under two main headings:

(1) *Zero drift*: The drift of the output of the amplifier with no input, or with an unvarying input. It is usually described in terms of the equivalent input signal required to restore the output to its original value. Zero drift cannot be improved by feedback and can only be reduced by compensation.

(2) *Gain drift*: The drift of the amplification factor, voltage or current. It can be described either as a percentage change, or as an equivalent input change. Gain drift can be reduced to any required extent by means of negative feedback. This does reduce the gain, but for a required gain, stability is achieved by having a large open-loop gain (limited by circuit instability) and reducing the closed-loop gain to the required value by means of feedback.

Assuming that gain stability can always be achieved, as mentioned, by means of negative feedback, the problem that remains in the design of transistor d.c. amplifiers is the reduction of zero drift.

Zero drift in transistor amplifiers has two main aspects —voltage and current drift. If the source impedance is

* *Formerly Everett Edgcumbe & Co. Ltd, now with S. Smith & Sons (England) Ltd.*

much larger than the incremental input impedance of the amplifier, then voltage drift of the input has relatively small effects and current drift causes most of the variation.

On the other hand, if the source impedance is small compared with the incremental input impedance of the amplifier then current drift is unimportant and voltage drift is responsible for most of the variation. Voltage drift in transistor d.c. amplifiers is caused by variation of V_{be}, while current drift is caused by variation of h_{fe}.

The main techniques used for the reduction of zero drifts in transistor d.c. amplifiers are:

(1) The use of temperature sensitive resistors to compensate for overall variations.

(2) The use of transistor chopper techniques, which have an inherently lower drift mechanism than ordinary direct coupled systems.

(3) The balanced amplifier technique, in which transistor variations are balanced by similar variations in similar transistors.

It may be observed here that the performance of a transistor amplifier can be extended by enclosing the entire

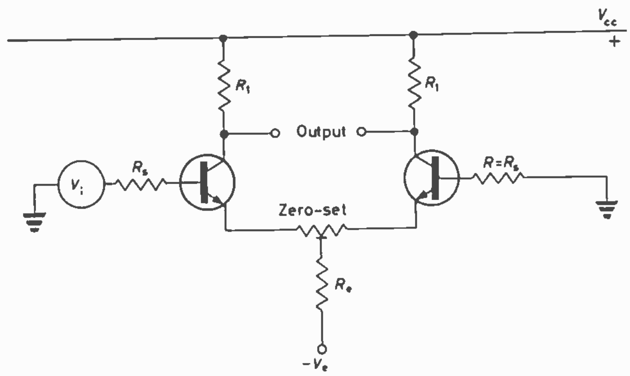


Fig. 1. Simple emitter-coupled differential amplifier

amplifier in a temperature controlled oven, the drift being then determined by the accuracy of control.

The use of temperature sensitive resistors is limited by the fact that tolerances of variation are wide, and for close compensation to be achieved, amplifiers have to be individually calibrated and checked at different temperatures. The chopper technique gives extremely good results and at the present stage of development, represents the best performance available with transistor circuits. The limitations of this method are cost, complexity and limited frequency response. This article is confined to a discussion of the third method named—i.e., the use of transistor drift itself as the compensating mechanism.

The Emitter-Coupled Differential Amplifier

The basic 'long-tailed pair' amplifier is shown in Fig. 1. Assuming identical components, the output should remain invariant under all conditions. In practice the transistors have different individual parameters which have different temperature coefficients. Initial unbalance is removed by means of a zero setting potentiometer and the circuit then has intrinsic drifts as follows:

Zero drifts: (1) The difference of the temperature coefficients of V_{be} of the two transistors causes an equivalent input voltage drift of the difference between the changes. (2) Assuming zero source resistance, the standing output voltage in each half of the amplifier depends on the h_{fe} of the transistor and therefore varies with variations of h_{fe} (see Appendix). The difference of temperature changes of h_{fe} of the two transistors causes a zero drift which may either add to or subtract from the drift due to V_{be} changes. (3) With finite source resistance, a further zero drift is caused by variation of the standing base current through this resistance due to variations of h_{fe}, and also variations of the resistance itself. When an equal resistance is included in the other input of the balanced pair, zero drift is caused by differential variations of standing currents and resistances. Zero drift due to (3) can be reduced by operation at low currents and with small source resistances. Using this amplifier, zero drifts due to (1) and (2) can only be reduced by closer selection and uniformity of transistors.

Gain drift: The gains of each section and of the amplifier depend on h_{fe} and therefore there is a gain drift which depends on the temperature coefficients of the current gains of the two transistors. This can be reduced by means of negative feedback, at the expense of gain.

An improved differential amplifier using negative feed-

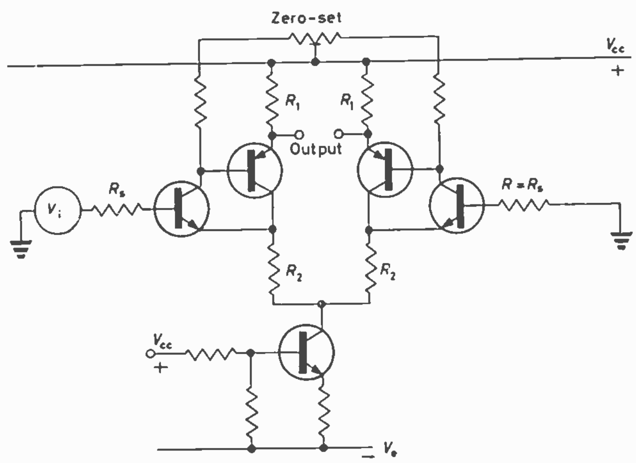


Fig. 2. Improved emitter-coupled differential amplifier

back in each section, due to Beneteau and others[3], is shown in Fig. 2. Each transistor in the differential pair is replaced by a compound transistor having a stable gain, defined by the ratio of two resistors (see Appendix). The circuit has two immediate advantages: (1) The reduction of zero drift due to the fact that the gain of each section is stable and the standing output voltage is therefore stable; and (2) gain stability, achieved by gain stability in each section of the amplifier. Also, since extra current gain is available, the currents in the input pair of transistors can be reduced, thus reducing the effects of standing current through the source resistance. However, the V_{be} differential drift is basically the same as that in the simple differential amplifier and remains the major source of drift in all transistor direct coupled, balanced systems operating from low source resistances.

For improved performance as a differential amplifier[4], R_e should be made as high as possible. This is done, without actually increasing V_e, by the circuit shown additionally in Fig. 2, where R_e is replaced by a transistor operating as a current source.

Some considerations that lead to improved performance can now be discussed. The currents in the input pair of transistors should be reduced as far as possible for three reasons: (1) Reduction of standing base current in the source resistance and consequent reduction of drift due to variation of this standing current and the source resistance; (2) greater uniformity of the temperature coefficients of the base-emitter voltages, since V_{be} approaches the ideal at low currents; (3) reduction of the power dissipated in the transistors and resultant minimization of differential heating. The reduction of current is limited by loss of gain at low currents and the input pair of transistors is usually operated at an optimum current at which the gain/drift ratio is a maximum. Transistors are now available with fairly high gains at collector currents down to the micro-ampere region, e.g., Fairchild 2N2484, Texas 2S502, with minimum gain of 100 at $10\mu A$.

The electrical and mechanical symmetry of the circuit is important. Different voltage and current levels in the first pair of transistors cause differential zero drift and hence care should be taken to preserve the electrical symmetry. Further, a difference of temperature of only 0·01°C between the input transistors would cause a zero drift due to V_{be} alone, of approximately $25\mu V$ and hence care should be taken that the input transistors are always maintained at equal temperatures. This is best done by using two transistors in one encapsulation, which ensures that there is maximum heat sharing. The efficiency of drift elimination is determined mainly by the similarity of transistors and therefore it is desirable that these are closely matched for V_{be}, h_{fe} and temperature coefficients. Double base-emitter voltage of the input transistor, and the current through the simulated source resistance. The input of the second amplifier is connected, in series with the source and a feedback voltage, to this temperature compensated virtual earth, as shown in the figure.

If initial unbalance is removed by means of the zero setting potentiometer, the circuit has a zero drift due to differential variations of V_{be}, identical to that obtained with the emitter coupled amplifiers. Further, since gain is stabilized by means of negative feedback, zero drift due to variations of h_{fe} is minimized and also overall gain is stable. It can be shown that for identical open and closed loop gains the circuit has a stability identical with that of the Beneteau circuit (see Appendix). However, it has the advantage that very large open-loop gains, and consequently improved stability, can easily and readily be achieved. Another important advantage is the versatility of feedback application. If, as shown in the figure, series-current feedback is used, the amplifier has a very high output impedance and approximates a current source. This feature is important in this amplifier, which was delevoped for use with recording instruments which have unspecified

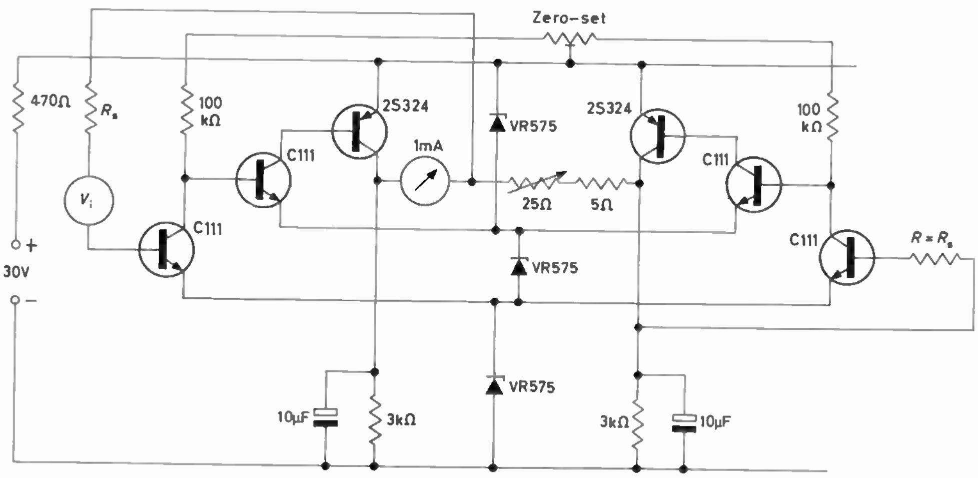


Fig. 3. The 'mirror-image' amplifier

transistors are now available that are matched to within a ratio of 0·9/1 for h_{fe} and with differential temperature coefficients of V_{be} matched to within $10\mu V/°C$. Typical drift figures for balanced amplifiers using these transistors are 3 to $5\mu V/°C$. The present prices of double transistors tend to be prohibitive for common use, but the downward trend is encouraging.

Different methods of zero setting may be used to compensate for initial unbalance. The method used in Fig. 1 compensates for differences between the base-emitter voltages and the value of resistance used should be as small as possible, since it causes decrease of stage gain. The method used in Fig. 2 eliminates the difference between the base-emitter voltages by the adjustment of current and seems to provide better drift performance.

The 'Mirror-Image' Amplifier

An amplifier having drift performance equal to or better than the Beneteau amplifier and having, in addition, very high (or very low) output impedance and the gain stability of an operational amplifier, is shown in Fig. 3.

It consists of two identical compound amplifiers and a balanced output. The input base of one of the amplifiers is connected to its output collector, producing a virtual earth with a voltage temperature coefficient dependent on the or varying resistances. If series-voltage feedback is used, the amplifier has a very low output impedance and approximates a voltage source. In the configuration shown, the circuit also has the advantage of having one terminal common to input and output and therefore may be used as a three terminal operational network.

The design of each section of the amplifier is simple. The overall gain is only limited by the maximum number of available stages and circuit instability. Collector voltages for succeeding transistors and the available output voltage swing may be kept reasonably high, without increasing the supply voltage, by using complementary transistors—i.e., pnp and npn. The standing base currents in the input transistors are made as low as possible to reduce drift due to variations of standing current and source resistance. Decrease of current in the first pair of transistors is not limited by loss of gain to the extent of the Beneteau circuit, since increased gain is readily obtained by using more stages of amplification. The collector currents and voltages for the first and second pairs of transistors are kept nearly constant (the collector currents of the first pair varying by approximately $0{\cdot}5\mu A$ for a full-scale input, the collector voltages remaining almost invariant) thus reducing drift due to differential conditions of these transistors. In this respect, the amplifier compares favourably with the improved second amplifier mentioned by Beneteau and Murari[5]. Stable voltage supplies are obtained from one source by the use of Zener diodes, though accurate stability of the supplies is not required since the circuit is, to a large extent, self-compensating. The feedback potentiometer may serve as a gain control, though very large gains may only be obtained at the expense of stability unless additional gain is introduced by using more stages. The capacitors are connected across the output collectors to eliminate r.f. instability in each section of the amplifier.

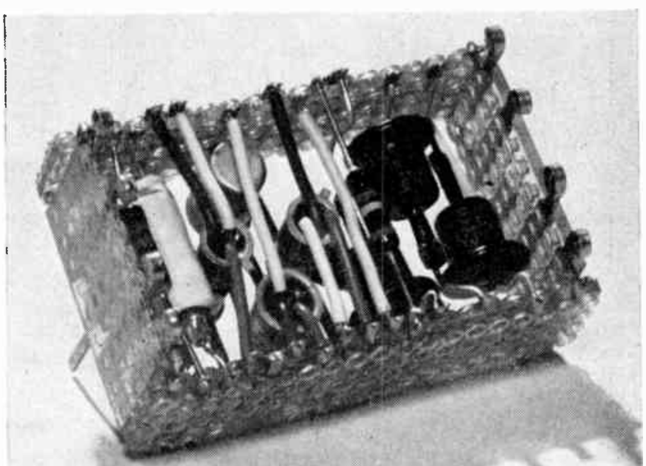
The circuit was built in the 'Ministac' form (Fig. 4) using inexpensive, unselected planar transistors for the first two pairs in the amplifier and ordinary pnp alloy-junction transistors for the third pair. With the input short-circuited, zero drifts of between 10 to 50μV/°C were observed for various input pairs of transistors, and this was roughly the range expected. Using improved, matched devices, the expected performances were obtained. A circuit was built, having 10nA standing base current and with an equivalent input voltage drift of approximately 2μV/°C. The basic amplifier was designed for an output of 0 to 1mA with an input of 0 to 25mV from a source resistance of not more than 100Ω. The incremental output impedance was over 0·25MΩ and the gain stability was within 0·1 per cent over a temperature range of 20 to 60°C.

The performance of d.c. amplifiers may be improved, as previously mentioned, by the 'brute-force' method of enclosing the entire amplifier in a temperature controlled block. An oven was developed for this purpose, which maintained the temperature of the amplifier at 65 ± 1°C. The recorded output of the amplifier showed that the amplifier, using the inexpensive planar transistors, had a short-term stability of ±20μV and a long term stability (due mainly to oven temperature drift) of about twice this figure.

New Trends

The desirability of using matched pairs of transistors in one encapsulation, for the input pair of transistors in balanced amplifiers, has been stressed. With the advent of integrated circuits, matched transistors are being produced on a single silicon chip, as opposed to two separate matched transistors mounted on one header. As the two devices are made by identical processes at the same time and are never physically separated, they have very closely matched characteristics and, important too, very good thermal characteristics. Apart from the larger yield of matched pairs, and consequent potential reduction of cost, single-chip pairs have very close thermal tracking between transistors and exhibit excellent drift performance.

Fig. 4. Method of construction



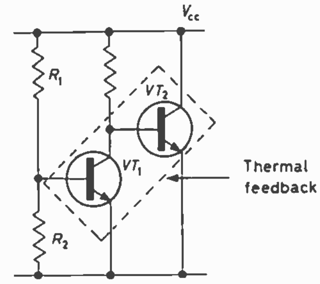


Fig. 5. Thermal substrate

The 'thermal substrate' or integrated oven is a new development that might well enable direct coupled amplifiers to equal conventional chopper amplifiers in performance, with the enormous advantages of simplicity and small size. The thermal substrate is a temperature regulating integrated circuit that can be encapsulated together with a matched pair of transistors, within a single T05 package. A typical thermal substrate consists of a two transistor amplifier with thermal feedback and is shown in Fig. 5. The circuit consists of resistors R_1 and R_2 which provide a reference voltage from the supply voltage V_{cc}. At low temperatures the reference voltage will be lower than the V_{be} of transistor VT_1 which will therefore be turned off, causing transistor VT_2 to conduct and dissipate power. This heats the base-emitter junction of VT_1 through the thermal feedback path and the temperature of the circuit stabilizes at the point at which V_{be} of transistor VT_1 is equal to the reference voltage. This temperature could be set by adjustment of R_1 or R_2. The degree of stabilization achieved using this technique is in the region of 1°C change of substrate temperature with 20°C change of ambient, giving a 20/1 improvement over non-stabilized circuits. At the present time the prohibitive cost has limited the use of the thermal substrate to specialized military applications. However, it is hoped that improvements in the technology of manufacture will enable the temperature stabilized transistor pair to be sold at a commercially realistic price in the not too distant future.

Work on drift-compensated transistor direct coupled amplifiers has been in progress for some time[6]. The fact that permits compensation is that drift is approximately linear with temperature. If the source resistance is known and included in the circuit, the overall drift can be compensated by injection of a linear compensating signal, or actual drift signal from a similar amplifier, the magnitude of which is adjusted by experiment. The remaining drift is then due to the difference between the actual drift and the compensation, and also the differential variation of source resistance. Using this technique, drifts of the order of 0·05μV/°C have been achieved.

Acknowledgments

The author would like to acknowledge that the 'Mirror Image' amplifier (for which a patent has been applied) was developed with the co-operation of and based on an original idea by Mr. G. I. Hitchcox. He would also like to thank the Directors of Everett Edgcumbe & Co. Ltd for permission to publish results of work carried out in their Research Department.

Appendix

Mathematical Analysis of Gain/Drift of Transistor D.C. Amplifiers

Consider the simple grounded emitter transistor amplifier without any form of feedback. In this analysis, a single transistor circuit will be discussed for the sake of simplicity, the results being generally applicable. In Fig. 6, if V_i is the input voltage,

R_s is the source resistance,

h_{fe} is the current gain of the transistor,

R_1 is the collector load resistance.

then:

$$V_i - I_b R_s - V_{be} = 0$$

where

V_{be} is the base-emitter voltage of the transistor,

I_b is the base current.

Therefore:

$$I_b = (V_i - V_{be})/R_s \quad \ldots\ldots\ldots\ldots\ldots (1)$$

If V_o is the output voltage, measured across R_1:

$$V_o = h_{fe} I_b R_1$$
$$= h_{fe} R_1 (V_i - V_{be})/R_s \quad \ldots\ldots\ldots\ldots (2)$$

Then the gain A_1 is given by:

$$A_1 = dV_o/dV_i = h_{fe} R_1/R_s \quad \ldots\ldots\ldots\ldots (3)$$

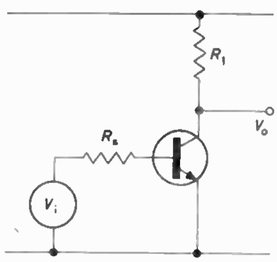


Fig. 6. Simple grounded-emitter amplifier

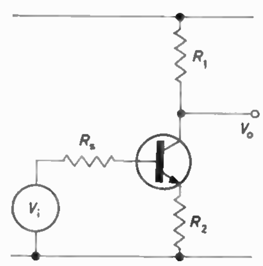


Fig. 7. Amplifier using emitter feedback resistor

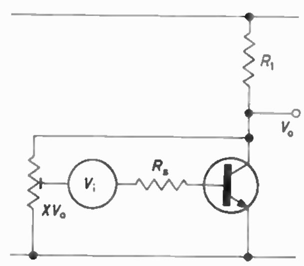


Fig. 8. Amplifier using external feedback

and the drift with h_{fe}, D_1 is given by:

$$D_1 = dV_o/dh_{fe} = (V_i - V_{be}) R_1/R_s \quad \ldots\ldots (4)$$

The figure of merit (drift/unit gain) represented by K_1 is given by the expression:

$$K_1 = D_1/A_1 = (V_i - V_{be})/h_{fe} \quad \ldots\ldots\ldots (5)$$

Next, if an emitter resistor R_2 is included as in Fig. 7:

$$V_i - I_b R_s - V_{be} - I_b (h_{fe} + 1) R_2 = 0$$

or:

$$I_b = \frac{(V_i - V_{be})}{R_s + (h_{fe} + 1)R_2} \quad \ldots\ldots\ldots\ldots (6)$$

V_o is now given by the expression:

$$V_o = \frac{h_{fe} R_1 (V_i - V_{be})}{R_s + (h_{fe} + 1) R_2} \quad \ldots\ldots\ldots\ldots (7)$$

The gain A_2 is now seen to be:

$$A_2 = dV_o/dV_i = \frac{h_{fe} R_1}{R_s + (h_{fe} + 1)R_2} \quad \ldots\ldots\ldots (8)$$

If it is assumed that $(h_{fe}+1) = h_{fe}$ and $R_s \ll (h_{fe}+1)R_2$, then A_2 is defined by the ratio R_1/R_2.

Using no approximations, the drift with h_{fe}, D_2 is seen to be:

$$D_2 = dV_o/dh_{fe} = \frac{(V_i - V_{be})(R_s + R_2)R_1}{(R_s + (h_{fe} + 1)R_2)^2} \quad \ldots\ldots (9)$$

The figure of merit K_2 is given by:

$$K_2 = D_2/A_2 = \frac{(V_i - V_{be})}{h_{fe}} \cdot \frac{(R_s + R_2)}{R_s + (h_{fe} + 1)R_2}$$

$$= K_1 \cdot \left(\frac{R_s}{R_s + (h_{fe} + 1)R_2} + \frac{R_2}{R_s + (h_{fe} + 1)R_2} \right) \quad \ldots\ldots\ldots\ldots (10)$$

If it is assumed that $(h_{fe} + 1) = h_{fe}$ and equations (9) and (10) are derived again, the expression for K_2 reduces to

$$K_2 = K_1 \cdot \frac{R_s}{R_s + h_{fe} R_2} \quad \ldots\ldots\ldots\ldots (11)$$

If R_2 is not included, but external feedback is introduced as shown in Fig. 8, then:

$$V_i - XV_o - I_b R_s - V_{be} = 0$$

where X is the feedback ratio.

Therefore:

$$I_b = \frac{V_i - V_{be} - XV_o}{R_s} \quad \ldots\ldots\ldots (12)$$

The output voltage is now given by the expression:

$$V_o = h_{fe} I_b R_1$$
$$= h_{fe} R_1 \frac{(V_i - V_{be}) - XV_o}{R_s}$$

or:

$$V_o = \frac{h_{fe} R_1 (V_i - V_{be})}{R_s + h_{fe} R_1 X} \quad \ldots\ldots\ldots\ldots (13)$$

The gain A_3 is given by:

$$A_3 = dV_o/dV_i = \frac{1}{R_s/(h_{fe}R_1) + X} \quad \ldots\ldots (14)$$

If $R_s/h_{fe}R_1 \ll X$, then A_3 is approximately $1/X$.

Using no approximations, the drift with h_{fe}, D_3 is given by:

$$D_3 = dV_o/dh_{fe} = \frac{R_1 (V_i - V_{be}) R_s}{(R_s + h_{fe} R_1 X)^2} \quad \ldots\ldots (15)$$

The figure of merit, K_3 is now:

$$K_3 = D_3/A_3 = \frac{(V_i - V_{be})}{h_{fe}} \cdot \frac{R_s}{R_s + h_{fe} R_1 X}$$

$$= K_1 \cdot \frac{R_s}{R_s + h_{fe} R_1 X} \quad \ldots\ldots\ldots\ldots\ldots (16)$$

Equation (16) is identical with equation (11) if $X = R_2/R_1$.

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Low Voltage, High Input Impedance Hybrid Amplifier Circuits

By F. H. Laishley*, A.M.I.E.R.E.

A family of hybrid d.c. coupled amplifier circuits is described. The circuits operate from low voltage supplies and have very high input impedances. The technique used is to employ a single valve requiring a low anode voltage in the input stage of each of the circuits and to follow this stage with conventional semiconductor circuits. This allows simple high impedance circuits to be obtained of small physical size and of low power consumption. The circuits were initially developed for measuring voltages stored on low value capacitors.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

TWO of the more recent additions to the range of solid state devices are the field effect transistor[1] and the dielectric triode[2]. In contrast with semiconductor amplifying devices they require a voltage input rather than a current input and they therefore allow simple circuits to be built having very high input impedances, of the same order as can be obtained with valves. Unfortunately, the former is only just becoming available commercially and is very expensive, while the latter is not yet commercially available at all.

A small number of valves, including electrometer types, is designed to operate from low voltage power supplies and these are easily incorporated into transistor circuits. Using this hybrid technique, it is possible to design simple very high input impedance circuits, using a single valve and conventional transistor circuits. Admittedly, even a single valve may not be permissible in some applications but in many cases a hybrid circuit of this type will provide a neat and simple solution.

Direct Coupled Buffer Amplifier Circuit

Consider first the non-inverting amplifier circuit shown in Fig. 1. This circuit has a very high input resistance, a low output impedance and good linearity. The voltage gain V_{o1}/V_{in} is almost unity and the d.c. level of V_{o1} may be set to earth potential by means of RV_1. Alternatively, a higher voltage gain V_{o2}/V_{in} is available from terminal T_2 at a higher d.c. level. The voltage gain at this point is approximately equal to $(R_1 + R_2)/R_2$.

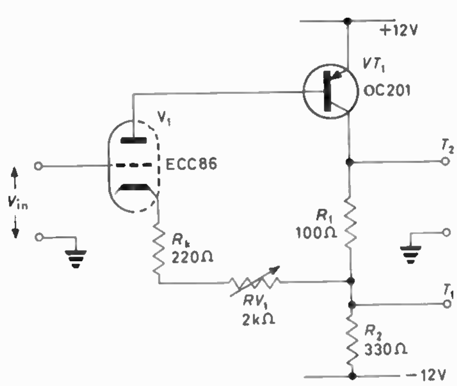

Fig. 1. Direct coupled buffer amplifier

Circuit Analysis

For analysis purposes, consider the simplified a.c. equivalent circuit shown by Fig. 2. The current generator represents the valve V_1 and the transistor VT_1, having a combined mutual conductance value of g_m', where $g_m' = \alpha' . g_m$, α' = grounded emitter gain of VT_1, and g_m = the mutual conductance value of V_1.

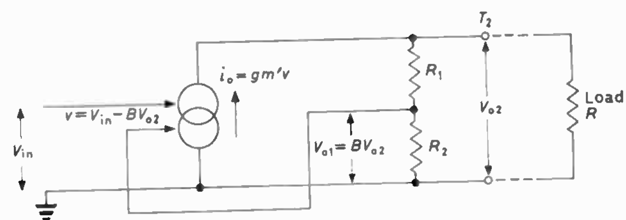

Fig. 2. A.C. equivalent circuit of Fig. 1

VOLTAGE GAIN

With no load resistance R connected, all the output current from the generator flows down the resistors R_1 and R_2 causing a voltage $B . V_{o2}$ to be developed across R_2, where $B = R_2/(R_1 + R_2)$. This voltage is fed back to V_1 cathode where it is in series opposition to the input voltage V_{in}. Hence the voltage gain of the circuit with negative feedback applied is $V_{o2}/V_{in} = A/(1 + AB)$ where A equals the open loop gain of the circuit and is equal to $g_m'(R_1 + R_2)$. Therefore:

$$V_{o2}/V_{in} = \frac{g_m'(R_1 + R_2)}{1 + g_m' R_2} \simeq \frac{R_1 + R_2}{R_2}$$

provided $g_m' R_2 \gg 1$.

It should be noted that in the above analysis it is assumed for simplicity that all the signal voltage developed across R_2 is fed back to V_1 cathode. In the actual circuit of Fig. 1, a small loss of signal occurs across the cathode resistor R_k and the series potentiometer RV_1.

OUTPUT RESISTANCE

To find the value of the output resistance R_g, of the current generator in Fig. 2, consider the case with V_{in} kept constant and a load resistance R connected between terminal T_2 and earth. This additional load will cause the current i_o to increase to $i_o + \delta i_o$, and V_{o2} to decrease to $V_{o2} - \delta V_{o2}$.

From Fig. 2; $i_o = g_m' v$

$$= g_m'(V_{in} - BV_{o2}) \quad \text{.......} \quad (1)$$

$$\therefore\ i_o + \delta i_o = g_m'(V_{in} - B(V_{o2} - \delta V_{o2}))$$

$$= g_m'(V_{in} - BV_{o2}) + g_m' B . \delta V_{o2} \quad .. \quad (2)$$

Subtracting equation (1) from equation (2) gives:

$$\delta i_o = g_m' B . \delta V_{o2}, \quad \text{.......} \quad (3)$$

and the output resistance of the current generator:

$$R_g = \delta V_{o2}/\delta i_o = 1/g_m' B \quad \text{.......} \quad (4)$$

* *Joseph Lucas Ltd.*

This resistance R_g is effectively in parallel with $R_1 + R_2$ thus the output resistance of the circuit at terminal T_2 is:

$$R_{g(T2)} = \frac{R_1 + R_2}{1 + g_m' B(R_1 + R_2)} = \frac{R_1 + R_2}{1 + g_m' . R_2}$$

$$\simeq \frac{1}{g_m' . B} \text{ provided } g_m' R_2 \gg 1.$$

Note

The expression obtained for the output resistance of the current generator with negative feedback applied, agrees with Dr. E. de Boer's theorem[3]. His theorem states that the effect of applying negative feedback to a current generator is to reduce its internal resistance to $1/g_m k$ where k is the fraction of the output signal fed back to the input.

INPUT RESISTANCE

The input current taken by the circuit of Fig. 1 is determined entirely by the amount of grid current drawn

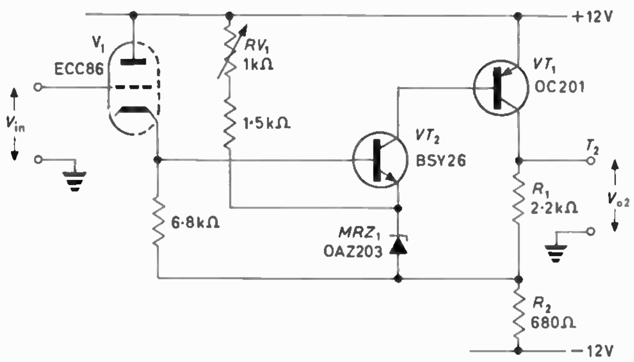

Fig. 3. An improved circuit

by the valve and the leakage paths provided by any components or wiring connected across the valve grid and earth. For an electrometer type valve, e.g. ME1400 series, the operating grid current is typically in the order of 10^{-13}A, and this can be further reduced with certain bias conditions. A disadvantage of this type of valve is that it is rather delicate and sensitive to mechanical vibrations. Also, it has a very low mutual conductance value.

Where a higher value of grid current can be tolerated, the electrometer valve can be replaced with one of the more conventional types of valve designed to operate from low voltage supplies; e.g. ECC86, EF98. The advantages to be gained from this type of valve are:

(1) More rugged

(2) Less sensitive to vibration

(3) A higher value of g_m

(4) Indirectly heated cathode.

An Improved Circuit

The circuit shown in Fig. 3 is slightly more complex but has the following advantages:

(1) The quiescent current through V_1 is stabilized by the Zener diode MRZ_1 and the base-emitter diode of VT_2. Therefore, the value of this current can be chosen to give the best working value of g_m, and it will not be affected by variations in gain of transistor VT_1 due to temperature changes.

(2) The output terminal T_2 can be adjusted to earth potential by means of RV_1. Hence voltage gain is available, while the zero d.c. level of the input signal V_{in} is maintained.

The gain of the circuit is $\simeq (R_1 + R_2')/R_2'$ where R_2' is equal to R_2 in parallel with RV_1 and the series 1·5kΩ resistor shown.

(3) The additional transistor VT_2 provides an increase in the open loop gain of the circuit, resulting in:

(a) Lower output impedance.

(b) Higher gain stabilization factor.

(c) Improved linearity.

The expression for the output resistance R_g is $1/g_m' B$ as in the case of Fig. 1 except that:

$$g_m' = g_{m(V1)} . \alpha'_{(VT1)} . \alpha'_{(VT2)}$$

$$B = \frac{R_2'}{R_1 + R_2'}$$

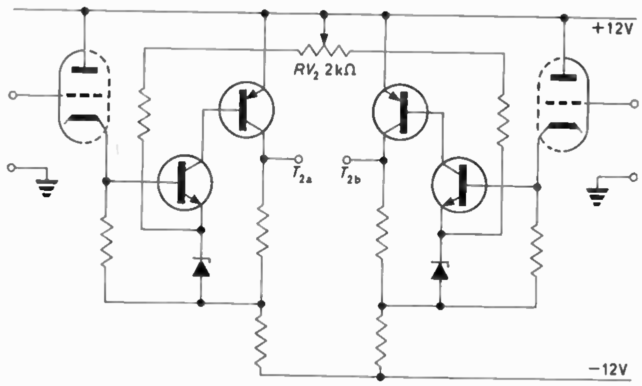

Fig. 4. Circuit arrangement for low drift

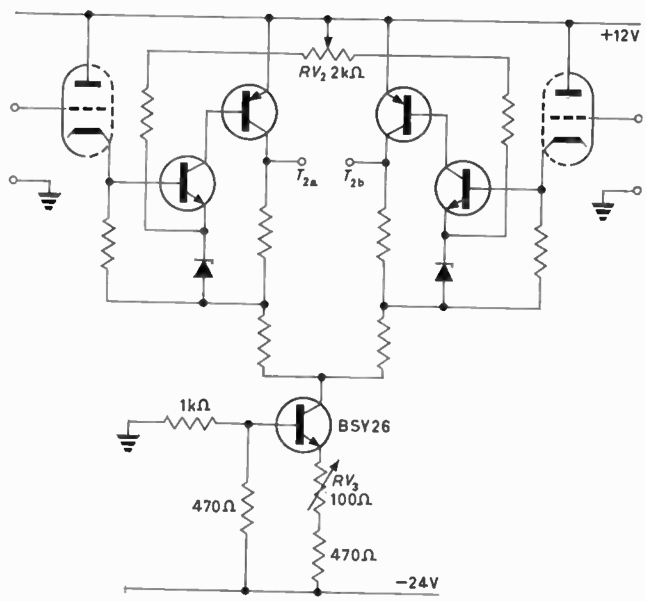

Fig. 5. Long tailed pair circuit

Circuit Arrangement for Low Drift

Where low drift performance is desired two circuits, such as either Fig. 1 or Fig. 3, may be connected as shown in Fig. 4. Drift in both halves of this circuit, due to similar variations in valve and semiconductor parameters, or h.t. supplies, will tend to cancel out.

For best performance the semiconductor devices should be mounted in a common heat sink of sufficient thermal capacity to eliminate temperature differences between transistors.

It should be noted that the common balancing potentiometer RV_2 balances the potentials of the output terminals

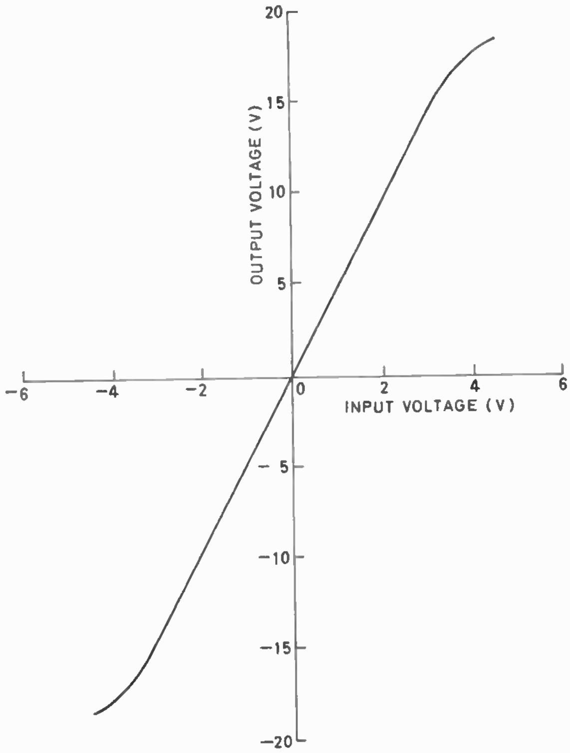

Fig. 6. Push-pull input-output characteristic of Fig. 5

T_{2a} and T_{2b} with respect to each other and does not balance them about earth.

Long Tailed Pair Arrangement

If the common negative h.t. supply point of Fig. 4 is fed from a constant current source, the familiar long tailed pair circuit results, see Fig. 5. The inclusion of RV_3 in the constant current circuit allows the potenials of the output terminals to be adjusted to earth. Balancing is effected by means of RV_2 as in Fig. 4.

Some of the applications of this circuit are as a differential amplifier, phase splitter or a d.c. amplifier, where high input impedances are required.

With the circuit values shown, a drift figure of 5mV/h was obtained (referred to the input terminals).

The input-output characteristic for a push-pull input and push-pull output is shown by Fig. 6. The linearity was better than 1 per cent within an output voltage swing of ± 10V.

Conclusions

The hybrid circuits discussed provide a means of realizing the high input impedances obtainable from valves while retaining the advantages of compactness and low power consumption offered by semiconductor circuits.

The drift performance obtainable from these circuits is similar to that obtained from their valve counterparts. Lower drift figures are obtainable from fully transistorized circuits[4], but these circuits have much lower input impedances.

Some useful applications of the hybrid circuits shown are:

(1) As a capacitor memory voltmeter circuit, i.e. where it is desired to read the voltage stored on a small capacitor without discharging it.

(2) As a high to low impedance buffer amplifier stage.

(3) As a d.c. valve-voltmeter circuit.

Acknowledgments

Thanks are due to Mr. D. Dixon who constructed and tested the circuits, and to Mr. M. J. Wright and Mr. M. J. Osborne for their helpful comments when checking the manuscript.

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A New Broadcasting Centre for France

Situated in the Quai de Passy, a new broadcasting centre has recently been brought into operation by the French broadcasting authorities bringing most of the Paris department of the Radiodiffusion-Television Francaise and the sound broadcasting studios under the one roof.

As can be seen from the photograph, the centre consists of two concentric circular buildings with a central tower. The outer circular building consists of ten floors mainly for the administrative sections for television and for the Paris sound broadcasting authorities.

Contained also in this building are three large orchestral studios for performances before public audiences. In the inner building of six floors is the main sound broadcasting distribution centre with its studios and control rooms.

The central tower of twenty-one floors houses the sound library and the disk and tape recording equipment.

There are in all some fifty-eight studios and seventeen control rooms at the broadcasting centre.

Aerial view of the Broadcasting Centre at Quai de Passy, Paris

ARIEL 2—The Second Anglo-American Satellite

THE second Anglo-American Satellite, Ariel 2, was launched on 27 March by an American Scout rocket from Wallops Island, Virginia, and is now orbiting successfully.

The rocket was launched in a south-easterly direction and the four stages carried it to a height of about 170 miles, where at a speed of about 18 000 mile/h, the satellite was injected into orbit, spinning at about 160rev/min. The spin was gradually reduced by the release of a special de-spin device, the extension of booms and paddles and finally a long wire aerial, until after 12min it came down to its planned rate of about 5rev/min.

The planned orbit of the satellite lies between latitudes 51°N and 51°S. It completes a revolution of the earth about every 103min. During each revolution its altitude varies between about 175 and 1 000 statute miles above the earth.

Ariel 2 is basically a cylinder about 2ft in diameter, 3ft in length and weighing about 150 lb. In orbit, it unfolded booms and paddles carrying solar cells, and unreeled a wire aerial 130ft long. The main satellite structure and ancillary equipment (telemetry equipment and power supplies) were designed, constructed and tested jointly by NASA Goddard Space Flight Centre and the Westinghouse Electric Company at Baltimore, Maryland, U.S.A.

The Experiments

The satellite contains experimental equipment designed by Cambridge University, Manchester University and the Meteorological Office. The experiments were chosen by the Royal Society's British National Committee of Space Research, in consultation with NASA. They are:

GALACTIC RADIO NOISE (*Dr. F. Graham Smith, Mullard Radio Astronomy Observatory, Cavendish Laboratory, University of Cambridge*)

The measurement of galactic radio noise in the frequency range 0·75 to 3Mc/s, and the exploration of the upper ionosphere.

The aims of this experiment are:

(a) To measure the intensity of the galactic radio background at frequencies which do not penetrate to ground based receivers under known conditions of receiver sensitivity and under known conditions of ionospheric effects on propagation and aerial impedance.

(b) To watch for temporal or spatial variations of galactic radiation.

(c) To explore the electron density in the upper F region.

The equipment was developed and produced by G.E.C. (Stanmore) Ltd. The wire for the long aerial was specially produced by British Insulated Callender's Cables.

ATMOSPHERIC OZONE (*Dr. R. Frith and Dr. K. H. Stewart, Air Ministry, Meteorological Office, Bracknell*)

This experiment has as its main objective the measurement of the vertical distribution of ozone in the earth's atmosphere as often and in as many places as possible. These measurements should add to our knowledge of the processes forming and destroying ozone, of the air motions which distribute it and of the effects of ozone on the thermal equilibrium of the upper atmosphere. Measurements are being made of the intensity of the radiation received from the sun at selected wavelengths in the ozone absorption region in the ultra-violet, at times when the satellite is entering or leaving the earth's shadow and the solar rays have to pass through the earth's atmosphere to reach the satellite.

The optical-mechanical equipment was manufactured by R. & J. Beck Ltd. A photocell with a special cathode was developed by Rank-Cintel Ltd.

The electronic equipment, which was designed and manufactured by the Guided Weapons Division (Filton) of the British Aircraft Corporation consists of four amplifiers, an e.h.t. generator and two h.t. generators.

One of the h.t. generators is on stand-by so that the overall reliability is increased through redundancy. These units are built on printed circuit cards and are encapsulated in foamed resin.

The maximum input to each amplifier is variable to allow for differences in detector outputs. The maximum input acceptable to the photocell amplifiers can lie between 0·5 and 5·0μA. The maximum input to the photomultiplier amplifiers can lie between 0·1 and 1·0 μA. The maximum output from the amplifiers is always 5·0V. The bandwidth of the amplifiers can extend from d.c. to 1·5kc/s and linearity over the working range is better than 1 per cent. Short term stability is better than 1 per cent and long term stability is better than 5 per cent over a temperature range of -15°C to $+60$°C. The e.h.t. generator provides 2kV stable to 0·1 per cent for the photomultipliers and the h.t. generators provide 100V for the photocells.

The complete equipment functions from a power supply of less than ¼W.

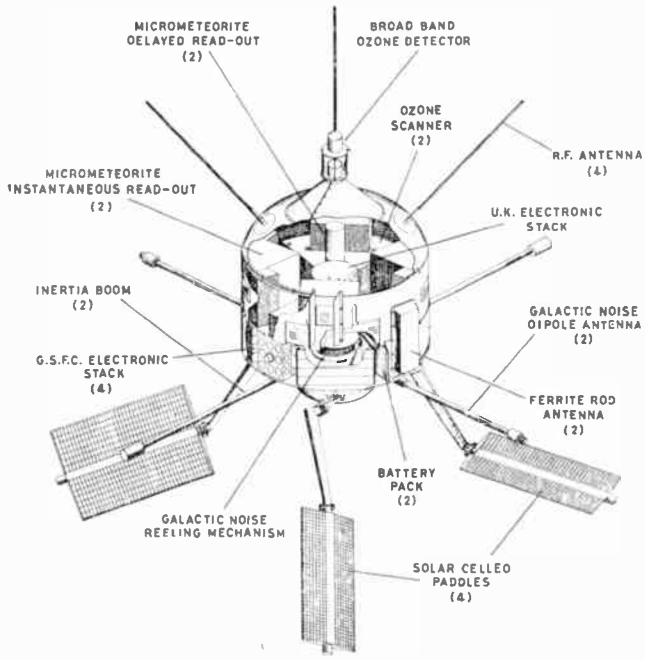
MICROMETEORITE FLUX (*Dr. R. C. Jennison, Nuffield Radio Astronomy Laboratories, University of Manchester, Jodrell Bank, Cheshire*)

This experiment has as its objective the detection and measurement of micrometeorites encountered by the satellite.

The electronic equipment was designed and manufactured by Ferranti Ltd, Wythenshawe in conjunction with Manchester University.

Basically the equipment consists of Instantaneous Read-Out Detectors (IROD) and Delayed Read-Out Detectors (DROD) mounted in Ariel 2.

The IROD consists of an aluminium foil, moving across a window, transversely to the spin axis of the satellite. When the micrometeorites make contact at about 1 000 000mile/h they punch holes through the foil. The sun's rays now shine through the holes made by the micrometeorites and are picked up by a bank of Ferranti n on p type solar cells mounted a short distance behind the foil. As the satellite is rotating the sun's rays shine through the holes only for a very short period of time every revolution but this is adequate for a pulse of electrical energy to be formed in the solar cells. Amplifiers in the satellite developed by Ferranti Ltd then amplify



Outline diagram of Ariel 2

the pulses to a level which can be transmitted. The number of pulses recorded determines the concentration, and the height of the pulses give the relative sizes of the micrometeorites. The exact size of each hole is determined by comparing the pulse height with those generated by two holes of a known size made before launching and placed one each side of the window containing the foil.

So that the foil does not become peppered with holes it is wound on about $\frac{1}{8}$in every revolution of the satellite.

Although intended for the separate experiment of detecting finer dust particles the DROD's also provide information relating to larger particles. The principle is the same as for IROD except that the foil strip (now a metallized mylar film) is wound on in a direction parallel to the spin axis of the satellite. These units are therefore less sensitive to satellite spin and in this respect also act as a fail safe device for the IROD units if the correct satellite spin rate is not achieved. The smaller micrometeorite particles abrase the metal surface of the film so that sunlight is transmitted through it. The amount of light transmitted before and after exposure to the space environment is monitored by Ferranti solar cells positioned at both ends of the foil window. The illumination of the film is restricted by spring loaded flaps on either side of the window such that as the sun sweeps across it a pulse output is obtained from the solar cells. The magnitude of the pulse is dependent upon the integrated area of the smallest particles penetrating the metal surface. Calibration holes again ensure the automatic checking of the system.

The complete equipment has been developed to operate in a temperature range of $-15°C$ to $+50°C$ in a near perfect vacuum.

The sensitivity is such that the holes formed by particles 1 micron in diameter or greater may be detected.

Telemetry and Tracking

Measurements taken by the satellite while in orbit are transmitted back to earth as soon as they are made but some are stored in the satellite on a miniature tape recorder and played back on receipt of a command signal.

The signals from the satellite are received by the D.S.I.R. Radio Research Station at Singapore and the sub-station in the Falkland Islands as well as by the world-wide chain of NASA Minitrack Stations, one of which, at Winkfield, Berkshire,* is operated by the staff of D.S.I.R.'s Radio Research Station at Slough. Using the radio signals received from the satellite telemetry transmitter, the Minitrack Station will also make observations of the satellite's direction each time it passes within their range. These tracking results collectively are used to determine the orbit and therefore the position of the satellite at any time. This knowledge of the satellite's position is essential for the subsequent interpretation of the experimental measurements.

Data Processing

Information received from Ariel 2 by the NASA Minitrack Stations including Winkfield and the D.S.I.R. Stations at Falkland Islands and Singapore are ultimately recorded on magnetic tapes which are now being sent to the D.S.I.R. Radio Research Station at Slough where preliminary data processing is being carried out.

The tapes contain measurements from the satellite experiments, intermingled in a pre-arranged sampling pattern and recorded in a pulsed frequency modulation code. The average pulse repetition rate is 55 per second while the signal frequency of each pulse, representing the magnitude of the experimental quantity concerned, falls within the band 5 to 15kc/s. The time of reception at the receiving station is recorded continuously in a coded form on two other tracks alongside the data from the satellite.

The tapes are first inspected in one section of the processing system and the satisfactory parts which have an adequate signal-to-noise ratio are selected and their usefulness is assessed.

Selected telemetry tapes are then replayed into the main section of the processing system. This essentially digitizes in binary form the experimental magnitudes represented by the signal frequencies of the pulses. The signals are

* A Satellite Tracking Station in Britain. *Electronic Engng.* 33, 160 (1960).

General view of the data processing equipment at the Radio Research Station, Slough. The control console is seen in the foreground and to the right is the tape editing console

first separated from the background noise by a bank of 128 filters. When a frequency falls within the pass-band (100c/s wide) of a filter the output from this filter suppresses the noise outputs from all the other filters. By this means signals as much as 12dB below noise level can be recovered. Each filter is connected to its own binary number generator and the number, in binary coded decimal form, corresponding to the active filter is passed to the output. Conversion accuracy from frequency to digital number is about 1 per cent.

The overall design and manufacture of the satellite data processing equipment was carried out by Plessey UK Ltd in conjunction with the D.S.I.R. and the NASA Space Flight Centre at Goddard.

Maximum flexibility has been the aim of the design and the equipment will be capable of dealing with any pulsed frequency modulation data format that may be used in future joint U.S.-U.K. satellites. The system has also been designed in such a way that it would be possible to process data many times faster than real time by a simple substitution of part of the equipment. In this way, data gathered during transmission periods totalling four hours, for example, can be compressed into a one-hour period.

The equipment consists of six 6ft × 19in rack cabinets containing the electronic circuits, two control consoles and three tape handlers. The racks contain about 750 plug-in circuit boards. These are of 23 different types, although there are more than 100 of each of the two commonest types.

In the main digitizing section, signals are selected from background noise by means of a unit containing a bank of 128 'comb' filters, each of which has a pass-band 100c/s wide. This enables a data frequency range of 5 to 15kc/s to be covered and also includes additional filter units above and below those limits to cater for possible drift and for the decoding of synchronization and format identification pulses transmitted outside the 5 to 15kc/s range. Each of the output wires of the comb filter is connected to a 3-character number generator which produces numbers suitable both for further decoding and for processing by the computer.

Data is assembled number by number at a rate of 55 per second in a sequentially addressed ferrite core store until one telemetry sequence of 256 data samples is present. The store is then emptied at a rate of 8 000 numbers per second on to a digital tape transport where one sequence forms a block of data. As each number is generated at the beginning of the process a parity bit is added and checked as it leaves the core store and again after it has been written on tape.

The equipment is provided with an internal simulator which enables it to be completely checked out without the necessity of using an actual tape recording. In addition, a system of marginal checking is embodied; this is intended to reveal incipient faults before they can cause a failure during processing.

Future Satellites

U.K. 3, the third in the Anglo-American series, is being built entirely in Britain, and should be ready for launching by NASA in about three years' time. Five groups of experiments have been chosen by the Royal Society's British National Committee on Space Research, in consultation with NASA, for the payload.

In addition, a number of British experiments have been accepted by NASA for inclusion in large U.S. 'observatory' satellites to be launched in the next few years.

A Low Cost Closed Circuit Television System

A new, low cost system of closed circuit television has been introduced by Associated Electrical Industries Limited. Fully transistorized, portable, and capable of being run from two 12V car batteries, it has evolved from the high performance equipments supplied by AEI for military purposes.

The two simple units in the system—a camera and control monitor—contain the minimum circuits necessary for high definition. It is a system that is controlled from the monitor, so allowing an installation to be supervised from a central or remote viewing point. There are no preset controls on the camera itself.

In schools, banks and industrial plants; in prisons, parks and post offices, closed circuit television is becoming a vital aid to scan, inspect or control. This new system from AEI, though simple in its basic form, is flexible enough to be extended for multi-channel work where a customer requires an elaborate system.

Transistorization, and the use of printed circuit boards, ensures reliability. This, in turn, eliminates any need for constant servicing and allows the equipment to be sited in the most inaccessible places without the need for continual removal.

With the camera weighing only 4 lb and the 14in monitor 30 lb—both having comparably small dimensions—the system is highly portable and can be set up comfortably in a confined space.

One of the most interesting features offered in an equipment in this price range—under £500 for a basic chain—is automatic sensitivity control. This control automatically retains constant picture quality in changing light conditions over a scene brightness of 60 to 1. A normal picture is possible with a scene illumination of 2ft-candles and 50 per cent subject reflection at $f/2$ lens aperture.

A wide range of power supplies can be used. These extend from two car batteries—a 24V 3A supply—to normal mains supplies covering 110V to 250V at 50 to 60c/s. Power consumption is 45W, including camera, and dissipation within the camera is only 5W.

The system has been designed to operate on the 625, 525 or 405 lines scan with a 2:1 interlace. The horizontal resolution is 600 lines and the bandwidth 8Mc/s, an important factor for high definition.

The camera can be supplied with a standard 16mm lens or any other lens can be provided to meet specific requirements—wide angle, telephoto and the like.

A distribution unit can be supplied where an application calls for a multi-channel system. This unit is used to select any picture available on control monitors—where more than one camera is used—and to distribute the picture to the required viewing monitors. Viewing monitors can take the form of either conventional domestic television receivers, or video receivers.

A switching unit can also be supplied to control the accessories available for the camera—such as zoom lens, pan and tilt, and windscreen wipers—which strengthens the remote control features of the basic system.

This new, versatile closed circuit television system has been designed and developed by the Electronic Apparatus Division within AEI's Electronics Group.

Tuned Detectors Using the Twin-T Circuit

By T. Spencer*, B.Sc.

This article deals with a specific application of the twin-T circuit to transistor amplifiers. The twin-T is used as the shunt feedback element of a stage, amplifying a narrow band of frequencies centred about the notch frequency of the twin-T, with a stage Q_F of approximately 5 at 10c/s; by cascading two such stages the Q_F is squared. An abstract of the complete theory of the twin-T using finite source and load impedances is given in an Appendix. A nomogram for the optimum design of this network is given. That the Q_F of a tuned amplifier with a twin-T network in the feedback path is directly proportioned to the d.c. transmission of the network, its intrinsic Q and the stage gain of the amplifier is proved, hence cascading stages not individually tuned unless a band-pass effect is desired and using only one twin-T network in the overall feedback path is preferable. Tuning is very simple, stability extremely good and almost any bandwidth is possible. Universal curves are given to facilitate the design of a tuned amplifier and it is significant that little departure from the design of valve circuits exists.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE use of printed circuits with transistors has led to the packaging of many more circuits than is otherwise possible in a given volume. The tuned circuit employing inductors still presents a problem where space is limited even though ferrite cores are available having very high Q-factors, but they are otherwise bulky, particularly at low frequencies.

The availability of 'Stantelum' capacitors, because of their very small size, makes the twin-T circuit as a feedback element in a transistor tuned amplifier an attractive proposition, without the problem of inductive coupling and regeneration, leading to simplicity in design and layout.

The intrinsic Q of the twin-T is low, but when used in an amplifier, a stage Q_F, sufficient for good selectivity is possible, comparable with that using inductive circuits. A single stage with inductive tuning may provide sufficient Q_F, compared with the necessity to cascade when using the twin-T, but cascading the former would require feedback to avoid instability and the overall Q_F and number of transistors used will probably be the same using either method. Irrespective of tuning, the variation of resonant frequency with temperature is a function of the transistor and must be minimized in the normal manner; good design will reduce the variation of stage-gain and selectivity. A very low temperature stability factor is most effective in stabilizing the gain, remembering that with 100 per cent feedback $A = f(\alpha)$, therefore never temperature independent; with inductive tuning the necessity to stabilize the operating conditions will reduce the stage Q_F.

The advantage of tuned amplifiers using the twin-T must be demonstrated with a practical circuit and a detector incorporating a 10c/s tuned amplifier is described. The majority of design information available on the twin-T applies to its use with valve circuits i.e. having an infinite load and zero source impedance. The theory for its use with transistor circuits is summarized in an Appendix, from which the feedback elements in the tuned stages of Fig. 1. were designed, using the nomogram. The universal curves derived in a second Appendix facilitate the stage design.

A Practical Circuit

The outputs from the tuned circuits in Fig. 1 are used to trigger an amplifier or blocking oscillator switching two lamp circuits. Detection occurs in the emitter-follower biased to cut off, with the average current indicated by a microammeter. The circuit is stable, linear and operates from $-5°C$ to $+55°C$ with a drift referred to the input of $14\mu V/°C$ and a gain of 500.

Because of space limitations, the detector is directly-coupled throughout. For good selectivity, high gain is required limited only by drift. The tuned amplifiers use the twin-T as a shunt feedback network at all frequencies other than resonance. Complementary circuits with Ge npn and Si pnp type transistors are used because of the voltage levels employed and the low cost of currently available Si npn types.

Circuit Adjustment

At the resonant frequency of the twin-T, transistors VT_1 and VT_3 act as grounded emitter amplifiers with resistive loads. Each tuned amplifier is separately adjusted by just permitting, under no-signal conditions, the stage to oscillate by reducing the twin-T shunt resistor R_3. The oscillations are then stopped by current feedback produced by a resistor in the emitter circuit, which although increasing Z_o of the twin-T reduces the closed-loop gain of the stage. A signal at the resonant frequency is then applied and only very slight adjustment of R_3 is necessary for maximum stage gain; two stages approximately square Q_F i.e. improve the selectivity and sensitivity. If close-tolerance components are not used for the twin-T, the gain at resonance and Q_F will be down, because the signal response after adjusting R_3 does not correspond to that when the stage was permitted to oscillate i.e. the gain is not optimized. It may be found that the base potential of VT_6 is not sufficiently close to the reference; adjustment of the input stage current-feed resistor will correct this, or alternatively the next preferred reference voltage may be used.

Experimental Results

The individual circuit Q_F's were found to be such that a 10 per cent frequency change produced approximately a 10dB change in output. The overall response curve of Fig. 2 reveals a Q_F per stage of 5 approximately. A drawback is the variation in resonant frequency with temperature, consequently the bandwidth. The linearity of the transfer characteristic measured at a frequency off resonance, with temperature as the parameter is good and constant over the full detector range. The sensitivity is almost doubled at sub-zero temperatures, with little change at elevated temperatures. For a four-stage direct-coupled circuit preceding the reference point, at the base of VT_6, the output drift is exceptionally low.

* *Birmingham University, formerly English Electric Aviation Ltd.*

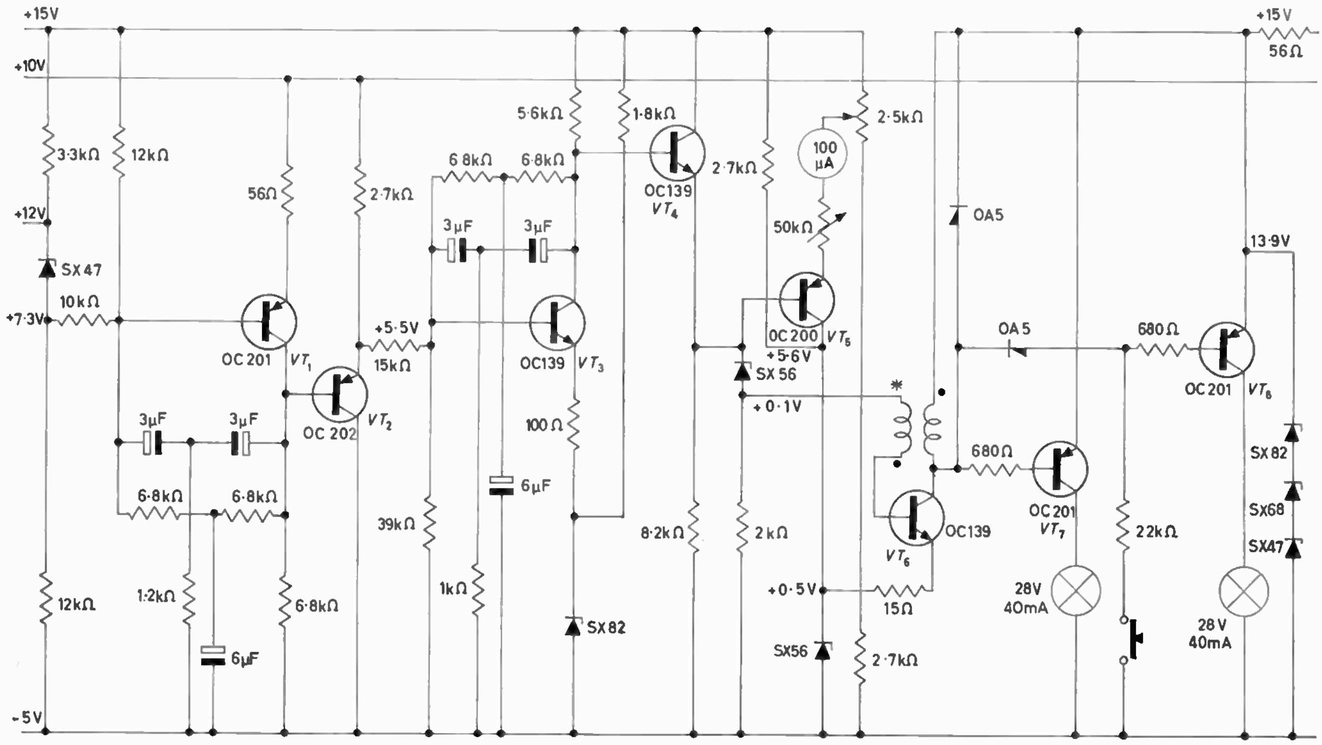
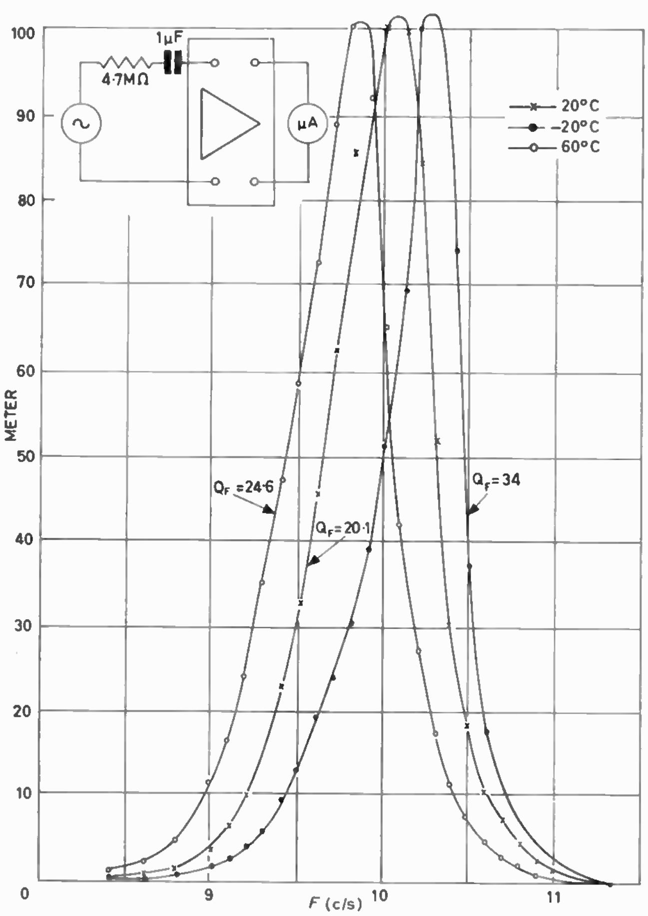


Fig. 1. 10c/s tuned detector

* *Ferroxcube LA1 : N_1 = 4118t 46 s.w.g. Lewmex 'F';*
N_2 = 1282t 46 s.w.g. Lewmex 'F'

Fig. 2. Overall frequency response



The Twin-T and Circuit Response

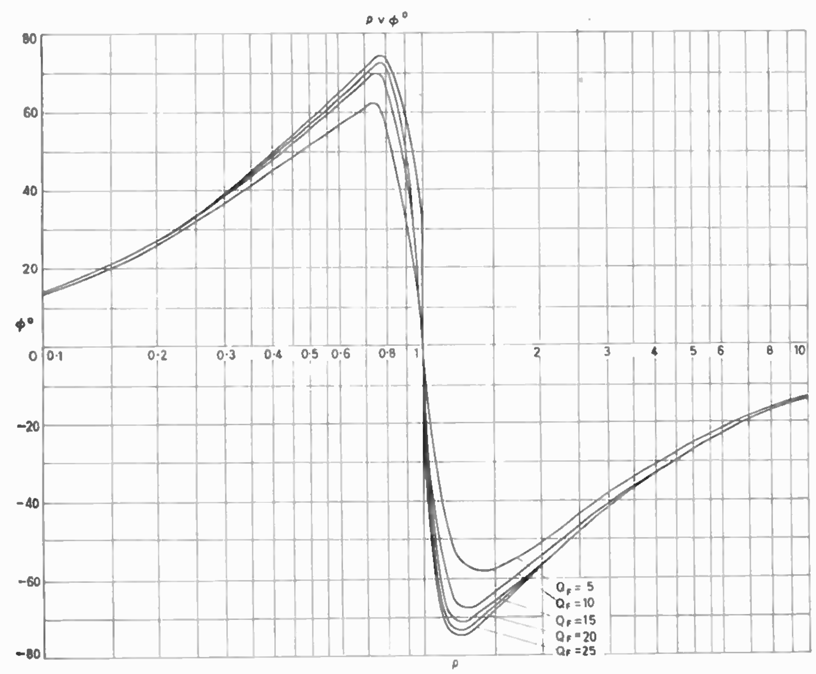
With reference to the tuning of the amplifiers by adjusting R_3 for maximum Q_F, its value at resonance is less than that theoretically derived when considering the twin-T network only and confirms that $n > 1$, since $R_3 = R/2n$ and R is fixed as it forms part of the d.c. stabilizing network of the stage.

The concept of Q with reference to the twin-T is justified since the series branch of the T_1 network following transformation of the twin-T is equivalent to a rejector circuit while the shunt arms are series RC. The Appendix shows that for the tuned parallel-RLC circuit near resonance, when the circuit Q is high, equations (1) and (2) are identical. Defining Q_F for the frequency-selective amplifier using a twin-T in the feedback loop the Appendix shows that for large values of Q_F:

$$Q_F = Q_{max} \,.\, T\,(o) \,.\, A \text{ if } T\,(o) \,.\, A \gg 1$$

where $T\,(o) = \dfrac{Z_o}{(Z_1 + Z_o) + 2R}$ is the d.c. or zero-frequency transmission.

Since $Q_{max} = 1/4$ and $T(o) = 1$, for a symmetrical filter having zero source and infinite load impedances with $n = 1$, $Q_F = A/4$ which is in agreement with the stage Q of a frequency selective amplifier using valves. Figs. 3(a) and (b) are 'universal' amplitude and phase curves as a function of ρ with Z_o/Z_1 as parameter for the symmetrical twin-T i.e. when $R^2 = Z_1Z_o\,(1 + n)$ with $n > 1$. These curves enable the gain-ratio to be determined for a required stage Q_F and frequency ω; for a required non-resonant gain A' the resonant gain A and phase shift ϕ are known. The curves were obtained from equations (10) and (11) Appendix (2). Q_F is fixed by the required resonant



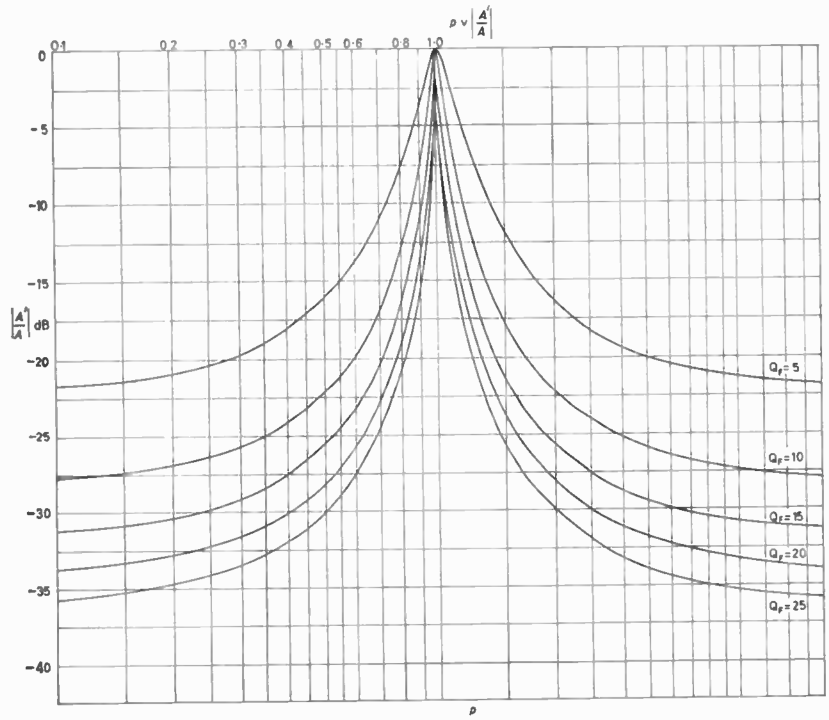


Fig. 3. Universal phase and amplitude curves

$Z_0 = Z_1$; $Q_{max} = 0{\cdot}392702$; $K = 1{\cdot}0$; $n = 1{\cdot}61$; $R = 1{\cdot}615Z_1$

frequency and bandwidth i.e.

$$Q_F = \omega_0 / 2\Delta\omega$$

and the universal curves are identical to those derived by Fleischer[3]. In fact, the analysis given in the second Appendix agrees in all respects with that given by him and shows that the gain-ratio, equation (12). is correct to within 1 per cent provided that the frequency range is restricted to:

$$(\rho - 1/\rho) = 0{\cdot}15/Q_{max}$$

With $Q_{max} = 1/4$ this range is $0{\cdot}744 \ll \rho \ll 1{\cdot}344$. The above ratio is identical to that of a high-Q rejector circuit, replacing the circuit Q by the overall stage Q_F.

In Fig. 3, the gain-ratio is little affected by Z_0/Z_1 particularly for low values of Q_F, hence only one set of curves is given.

Hyde[4] has analysed the unbalanced or asymmetrical twin-T with infinite load and zero source impedance and verifies that deliberate unbalancing of the filter by reducing R_3 when the former is used as a series feedback network in a single stage amplifier improves the selectivity and a Q_F of 20 is easily obtained. He has shown that oscillation occurs at that value of n for which the real part of β for anti-phase transmission is approximately $-1/A$, where A is the stage gain at $\rho = 1$. Also that for a balanced filter ($R_3C_3 = RC$) when $n = 1$ with series negative feedback the value of Q_F obtainable is $A/4$, in agreement with the Appendix.

It has been deduced from Appendix (1) that for a symmetrical notch $Q_{max} = 1/4$. White[2] has shown that by deliberately loading the twin-T, its Q is improved and that for $T(o) = T(\infty) < 0{\cdot}4$, Q increases with n; improvements in Q are at the expense of insertion loss. Bolle[5] has shown that an asymmetrical network provides a $Q_{max} = 1/2$ and that $n = 1$, $RC = R_3C_3$ are necessary conditions for oscillations. He shows that variations of R_3C_3 in such a manner that it remains constant, produces anti-phase as well as in-phase transmission for the same frequency for which $\beta = 0$ before detuning.

Lynch and Robertson[6] deduce an expression equal to that derived in Appendix (1) viz: $4R_3C/RC_3 = 1$. As their analysis holds for an infinite load and zero source impedance, it may appear that for $R_3 = R/2n$ and $C_3 = 2C/n$, these components are independent of the load and source impedances, but the cubic curve in the nomogram is a function of n, Z_1 and Z_0 and they affect R_3 and C_3; the ratio $4R_3C/RC_3$ is independent of n.

Smith[7] states that the null point for a balanced network is independent of variations in source and load impedances and always occurs at $\rho = 1$. From Appendix (1) $\omega_0 = \sqrt{n/RC}$; the nomogram shows the variation of n with Z_1 and Z_0, therefore ω_0, since RC is constant. The use of normalized frequency will always produce a null at $\rho = 1$, irrespective of n and obscures the fact that ω_0 is dependent on it.

The experimental results in Table 1 Appendix (1) correspond to networks having ω_0, R, Z_1 and Z_0 fixed; from K, n is known hence R_3, C and C_3. The measured values of R_3 are in error, not because of inaccuracy in determining n from the nomogram (Fig. 5) or that $R^2 \neq Z_1Z_0(1 + n)$, but due to component tolerances resulting in $4R_3C/RC_3 \neq 1$ and a sensitivity of 50mV/cm being used for the measurements. For a given value of R_3 using n from the nomogram $\omega \neq \omega_0$ or for a given $\omega = \omega_0$, R_3 is not equal to the calculated value. The error in n is taken as the difference between R_3 measured and calculated. The results in Table 1 reveal the important fact that n obtained from the nomogram results in accurate design, irrespective of symmetry, although the nomogram applies to $R^2 = Z_1 Z_o (1 + n)$.

The theoretical curves of Fig. 4 are plotted from equations (5) and (6) in Appendix (1), which are perfectly general and true only when:

$$R_1 = R_2 = R$$
$$C_1 = C_2 = C$$
$$R_3 = R/2n$$
$$C_3 = 2C/n$$
$$\omega_o = \sqrt{n}/RC$$
$$4R_3C/RC_3 = 1$$

Not only does the equality $R^2 = Z_1 Z_o (1 + n)$, when substituted in the equations produce symmetry, maximum attenuation β and Q_{max}, but optimizes n where:

$$(n^3 - n^2 - n + 1) = \frac{4 Z_1 Z_o}{(Z_1 + Z_o)^2} = K$$

which is presented in the nomogram used in the design of the filter. The curves of Fig. 4(a) comply with $4R_3C/RC_3 = 1$ irrespective of symmetry and the equation in terms of R_3 and X_3 would reveal a departure from this condition, with $\rho \neq 1$ and $\beta \neq 0$, if either one of these components of the filter be altered, keeping the other constant. The effect of $R^2 \neq Z_1 Z_o (1 + n)$ is apparent in Fig. 4 i.e. $T(o) \neq T(\infty)$ and a null will always occur at $\rho = 1$.

Should an initial condition where $R_3 = R/2n$ and $C_3 = 2C/n$ be changed by varying R_3 with C_3 R and C fixed, n is effectively changed since:

$$4R_3C/RC_3 = 4R/2n \;.\; C/RC_3 = 2C/C_3 (1/n) = 1$$

For variations in C_3: $4R_3/R \;.\; nC/2C = (2R_3)/(R)\, n = 1$ Reducing R_3 makes $\rho < 1$ since $\omega_o^2 = 1/2C^2RR_3$; increasing C_3 makes $\rho > 1$ since $\omega_o^2 = 2/R^2C_3C$. In general therefore $1 \geqslant 4R_3C/RC_3 \geqslant 1$ and $1 \geqslant \rho \geqslant 1$; transforming the above equation for β in terms of R_3 and C_3 to demonstrate this is unnecessary. According to Bolle, R_3 and C_3 should be altered so that the product remains constant, therefore ω_o.

The Twin-T and Phase Shift

The phase curves Fig. 4(b) show the limits of θ as $\pm\,\pi/2$ at $\rho > 1$ when $R^2 > Z_1 Z_o (1 + n)$; conversely when $R^2 < Z_1 Z_o (1 + n)$ the limits are $\pm\,\pi/2$ at $\rho < 1$, the angle decreasing to zero as ρ extends in either direction to zero and infinity.

With β expressed in terms of R_3 and X_3 the effect of $4R_3C/RC_3 \neq 1$ may be proved, but apparently[7] with $4R_3C/RC_3 < 1$, either by increasing or decreasing n keeping C_3 or R_3 constant, θ decreases continuously from 0 to $-\pi$ for $\rho < 1$, followed by an extremely rapid change

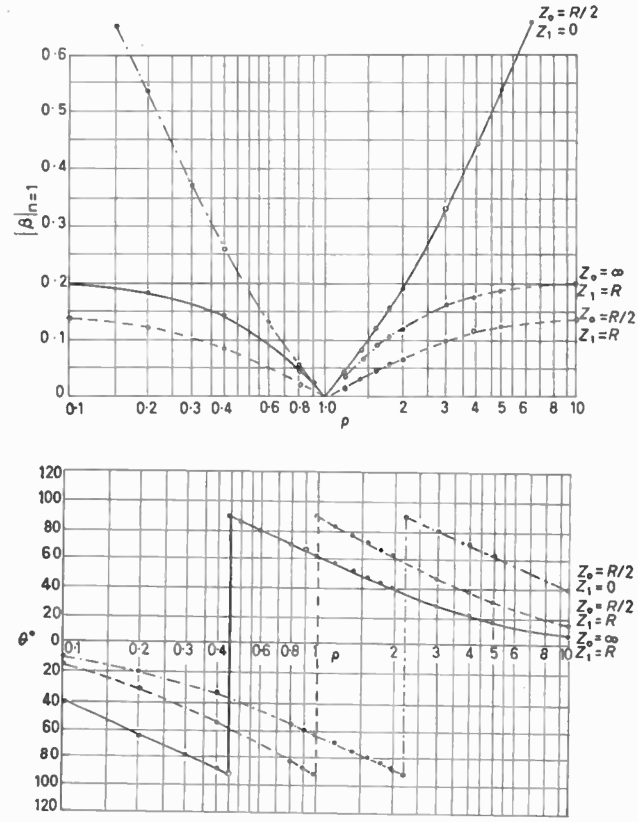


Fig. 4. Theoretical curves

from $-\pi$ to $+\pi$; as $\rho \to \infty$, $\theta \to 0$. When $4R_3C/RC_3 > 1$, θ no longer decreases from 0 to $-\pi/2$, but is less as ρ increases to unity; $\rho < 1$ a positive change in θ occurs, no longer instantaneous but gradual, the slope between the limits depending on the degree of departure of the above ratio from unity. The closer this ratio is to unity the steeper will the gradient of the phase curve become and the greater its limits until they equal $\pm\pi/2$ at unity ratio.

Asymmetry and $\beta \neq 0$ affects the stage gain at resonance; if $R^2 > Z_1 Z_o (1 + n)$ the net phase shift $\phi < 180°$, conversely if $R^2 < Z_1 Z_o (1 + n)$ the net phase shift $\phi > 180°$ and when the filter is balanced A' is independent of β at $\rho = 1$. This conforms with Fig. 3 which assumes a balanced filter, hence the limits of ϕ being close at $\rho = 1$. The loop phase-angle ϕ includes and is opposite in sign to that of the filter, because of the amplifier phase-angle π, particularly if $Z_o \gg Z_1$. At $\rho = 1$ in Fig. 3(b) ϕ is about $\pm 30°$ instead of $\pm 90°$ and decreases with Q_F; as Z_o/Z_1 increases ϕ approaches $\pm\pi/2$ becoming independent of Q_F.

Although when $4R_3C/RC_3 = 1$, Q_F is a maximum for $R^2 = Z_1 Z_o (1 + n)$ at $\rho = 1$, in practice a reduction of R_3 such that $4R_3C/RC_3 < 1$ improves Q_F still further. For $\theta > \mp\pi/2$, the closed-loop response increases[7]; when $\phi = 0$ the stage will oscillate and if $Z_1 = 0$ and $Z_o = \infty$, the frequency of oscillation $\omega_o = 1/RC = 1/R_3C_3$ with $n = 1$. Smith[7] confirms that variations in R_3 or C_3

TABLE 1

f (c/s)	Z_1 (kΩ)	Z_2 (kΩ)	K	n	R (kΩ)	C (μF)	C_3 (μF)	R_3 (Ω) CALC.	R_3 (Ω) MEASD.	Δn (PER CENT)	$4R_3C / RC_3$	$R_2 / Z_1Z_o(1+n)$
10	3·90	1·50	0·804	1·560	6·8	3·00	3·750	2180	2080	−4·59	0·954	3·09
200	0·60	0·60	1·000	1·618	10	0·10	0·124	3090	3100	+0·324	1·003	106·10
200	1·82	1·50	0·990	1·615	10	0·10	0·124	3096	3000	−3·226	0·968	14·00
1000	1·00	−0·338 −j0·943	3·000	2·000	1·732	0·13	0·130	433	410	−5·31	0·946	1·00

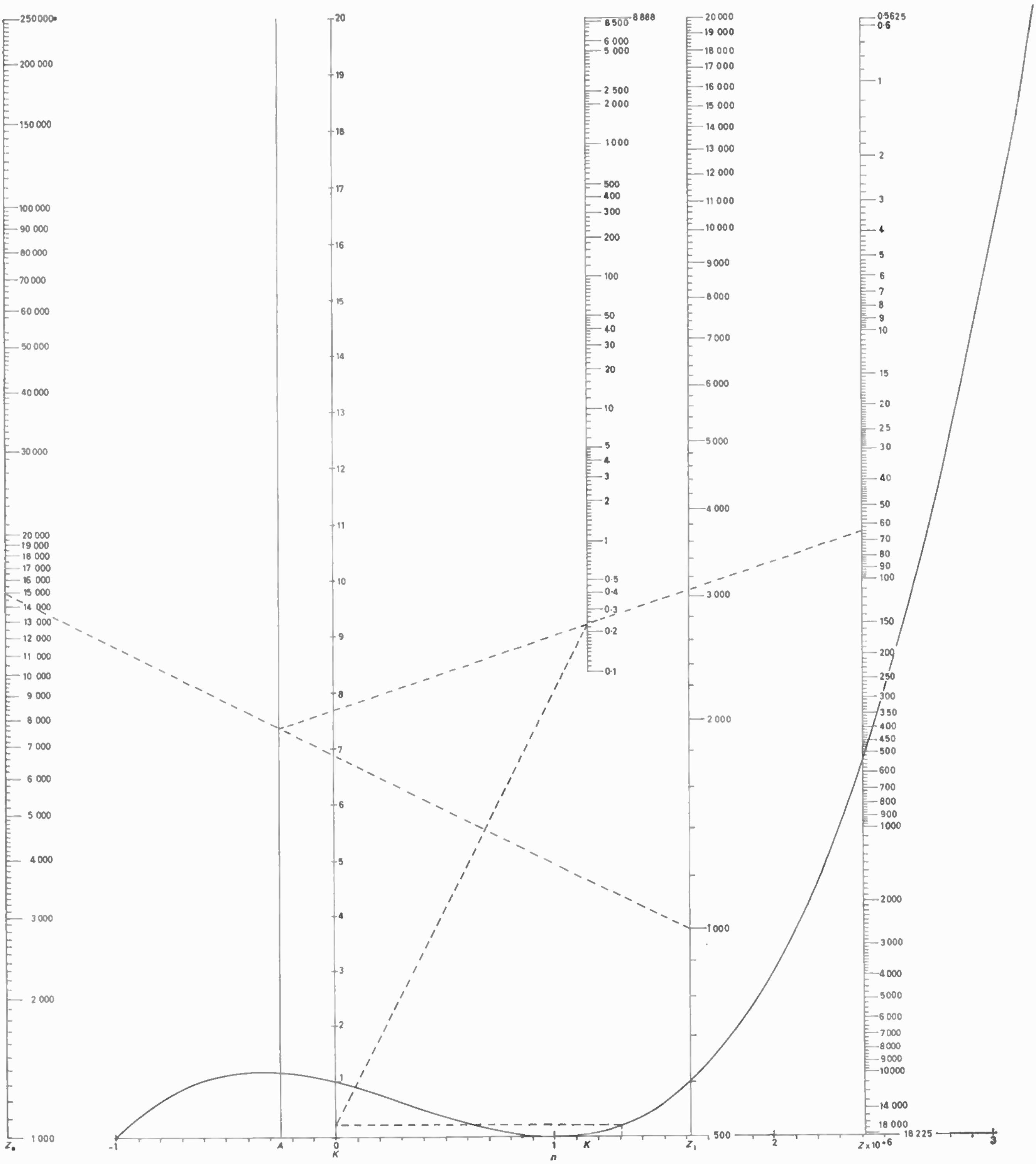


Fig. 5. Twin-T nomogram

Z_0 = modulus of load impedance; Z_1 = modulus of source impedance;
$Z = ((Z_1 + Z_0)/2)^2$; $K = 4Z_1Z_0/(Z_1 + Z_0)^2 = Z_1Z_0/Z$; $R^2 = Z_1Z_0(1+n)$

produce oscillations and because of the steepness of the phase gradient, that the limits of θ are $\pm\pi$ and the symmetry of this curve when $4R_3C/RC_3 \leqslant 1$, a reduction in R_3 rather than an increase in C_3 is to be preferred, as good frequency stability and ease of design will result.

Smith's value of $K = 0{\cdot}42$ (where $K = 1/2n$ and is not to be confused with K in this article) confirms the optimization of the filter designed from the nomogram, for substituting the value of $n = 1{\cdot}19$ corresponding to this figure in the nomogram produces from it a value of $K = 0{\cdot}089$. Checking this by substitution of $Z_1 = 0{\cdot}1R$ and $Z_o = 5R$ used[7] in the equation $4Z_1Z_o/(Z_1+Z_o)^2$ gives an alternative value for $K = 0{\cdot}077$; the inequality will be due to the inaccuracy of the figure 0·42 quoted. Thus instead of obtaining the best value of n by successive approximation[7], the nomogram gives it directly.

Impedance and Temperature

The impedance looking into the filter from the source is $R/\sqrt{(1+n)}$ at $\rho = 1$ irrespective of symmetry and equals that looking into the filter from the load. For low values of R, its loading effect is almost independent of n and because of this Q_F will be low, confirmed by the response curves of Fig. 2. The effect of Z_1 and Z_o at low and high frequencies should be considered in design since asymmetry reduces undesirable transmission above or below $\rho = 1$, i.e. produces a low- or high-pass filter effect.

For low values of R the input impedance contributes to the frequency stability of the amplifier with temperature. Z_1, approximately equal to the collector load, decreases little with temperature assisted by the shunt filter impedance and is almost temperature independent. Z_o will increase with temperature and the larger the source impedance looking back from the base of the transistor, the smaller ΔZ_o. But for the filter, the source resistances in Fig. 1 are high; changes in Z_o are damped by the filter impedance and consequently little change in K and therefore in n and ω_o will occur as is evident from Fig. 2. By deliberately making $4R_3C/RC_3<1$ when tuning at room temperature, n is effectively increased to values greater than unity with a consequent reduction in incremental changes with K, which in general changes very much less than Z_o.

The phase of the output varies with respect to the base input, if the emitter is grounded, and its effect on ρ is considerably reduced if the phase curve is steep at $\rho = 1$, resulting in good frequency stability with temperature.

Conclusions

Variations in gain with temperature will occur, and the transistor frequency-selective amplifier using the twin-T filter is extremely stable, although Q_F has been deliberately increased to the point of regeneration; the frequency and bandwidth changes can be reduced through design. The tuned stages of Fig. 1 are very simple and the use of emitter-followers as source and load to a filter network will improve the frequency and gain stability. An equivalent improvement is possible using the 'long-tailed pair' with an emitter-follower output, but the latter can be dispensed with if large load resistors are used. Unfortunately the unbalanced output gain is half that of a single-ended grounded-emitter stage; twice the output of such a stage could be obtained using a 'virtual earth' if the input and output are balanced. Emitter-followers are used in Fig. 1 to maintain selectivity.

A saving in cost and components will be brought about by cascading three stages to obtain Q_F, using only one notch filter in the feedback loop. Since $Q_F \simeq Q_{max}\,T(o)A$, the use of one filter increases $Q_{max}T(o)$ and A will be large because of a greater gain per stage with the necessity to current-feed the first stage only. Using series voltage feedback, two stages will be necessary with an apparent improvement in Q_F^4 as compared with shunt feedback.

Acknowledgment

The author wishes to thank the Management of English Electric Aviation Ltd for permission to publish this article; to Mr. S. C. Dunn, Head of Electronics and Control for his interest and encouragement; to Mr. P. Rooke and Mr. J. Fox of the Mathematics and Physics division for their respective derivations of the equation optimizing n, and Mr. H. Lutterlock for his assistance with the experimental work.

Appendix (1)

Following the star-delta transformation (Fig. 6), let the

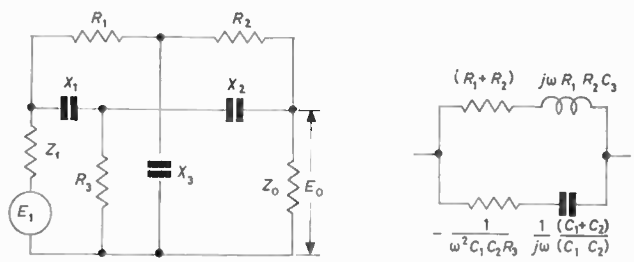


Fig. 6. Star-delta transformation

series branch impedance

$$Z = \infty :$$

$$\therefore\ Y = 1/Z = 0$$

If:

$$\left.\begin{matrix} R_1 = R_2 = R \\ C_1 = C_2 = C \end{matrix}\right\}\text{then } Y = \frac{1}{2R + j\omega R^2C_3} - \frac{1}{1/(\omega^2C^2R_3) + j2/\omega C} = 0$$

$$\therefore\ \omega^2 = \frac{1}{2C^2RR_3} = \frac{2}{R^2C_3C}$$

$$\therefore\ 4R_3C/RC_3 = 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

or:

$$4\,.\,(C/R)\,(R/2n)\,(n/2C) = 1$$

$$\therefore\ R_3 = R/2n \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

$$C_3 = 2C/n \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

$$\omega_o = \sqrt{n}/RC \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

$$\beta = E_o/E_1 = \frac{Z_o(\rho^2-1)}{\left[(Z_1+Z_o)(\rho^2-1) + \frac{2\rho^2Z_1Z_o(1+n)}{R} - 2R\right] - j2\rho\left[(Z_1+Z_o)\left(\frac{1+n}{\sqrt{n}}\right) + (Z_1Z_o/R)\left(\frac{1+n}{\sqrt{n}}\right) + (R/\sqrt{n})\right]} \quad \ldots\ldots \quad (5)$$

$$\text{Tan}\,\theta = \frac{2\rho(Z_1+Z_o)\left(\frac{1+n}{\sqrt{n}}\right) + 2\rho\,(Z_1Z_o/R)\left(\frac{1+n}{\sqrt{n}}\right) + 2\rho R/\sqrt{n})}{(Z_1+Z_o)\,(\rho^2-1) + 2\rho^2\,(Z_1Z_o/R)\,(1+n) - 2R} = \frac{\rho}{Q(\rho^2-1)} \quad \ldots\ldots \quad (6)$$

$$Q = \frac{(Z_1+Z_o)\,(\rho^2-1) + 2\rho^2\,(Z_1Z_o/R)\,(1+n) - 2R}{2(Z_1+Z_o)\,(\rho^2-1)\left(\frac{1+n}{\sqrt{n}}\right) + \frac{2(\rho^2-1)}{\sqrt{n}}\left[\frac{Z_1Z_o(1+n)}{R} + R\right]} \quad \ldots\ldots \quad (7)$$

Using these equations in the transformed transfer function with $\rho = (\omega/\omega_o)$

For a symmetrical notch and maximum Q:

$$R^2 = Z_1Z_o\,(1+n) \quad \ldots\ldots\ldots\ldots \quad (8)$$

For the notch to be as narrow as possible with respect to n, the slope at a point on the β characteristic as a function of ρ, is to be a maximum with respect to n for ω_1

very close to ω_0. If ω_1 is a fixed distance from ω_0, then the slope decreases as the notch narrows, unless ω_1 is very close to ω_0.

If $\omega_1 = (\omega_0 + \delta\omega_0)$ i.e. $\rho = (1+\delta)$, $\omega_0 = \sqrt{n}/RC$, $R^2 = Z_1 Z_0 (1 + n)$

by substituting for ω_1 i.e. $\rho = (1 + \delta)$

$$\left(\frac{d\beta}{d\rho}\right)_{\substack{\omega_1\\ \delta \to 0}} = \frac{jZ_0}{(Z_1+Z_0)\left((1+n)/\sqrt{n}\right)+2\sqrt{\left[Z_1 Z_0\left((1+n)/\sqrt{n}\right)\right]}}$$

For

$$\frac{d}{dn}\left[\left(\frac{d\beta}{d\rho}\right)_{\substack{\omega_1\\ \delta \to 0}}\right] = 0:\ (n^3-n^2-n+1) = \frac{4Z_1Z_0}{(Z_1+Z_0)^2} = K \quad \ldots\ldots\ldots\ldots (9)$$

APPENDIX (2)

Assuming Q of the rejector circuit (Fig. 7) large:

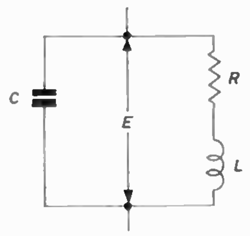


Fig. 7. Rejector circuit

$$\beta' = E/E_{max} = \frac{1}{1 + jQ(\rho^2 - 1)/\rho} = \frac{1}{1 + j\gamma} \quad \ldots\ldots (1)$$

For the twin-T with $Z_0 = \infty$, $Z_1 = 0$ and $n = 1$:

$$\beta = \frac{1}{1 - j\rho/Q_{max}(\rho^2 - 1)} = \frac{1}{1 + 1/j\gamma} \quad \ldots\ldots\ldots (2)$$

with $Q_{max} = Q$

$$\text{If } \delta = \frac{\omega - \omega_0}{\omega_0} = \Delta\omega_0/\omega_0 = (\rho - 1):\ \beta' = \frac{1}{1 + j2\delta Q} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

$$\therefore \lim_{\delta \to 0} |\beta| = 2\delta Q\,|\beta'| \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

For a frequency-selective amplifier: The gain ratio

$$A'/A = \frac{1}{1 + \beta A}$$

From Appendix (1):

$$\beta = \frac{Z_0/[(Z_1+Z_0) + 2R]}{1 - j\rho/Q_{max}(\rho^2 - 1)} = \frac{T(o)}{1 - j/\gamma} \quad \ldots\ldots\ldots (5)$$

$$\therefore A'/A = \frac{[1 + T(o)A + 1/\gamma^2] - j/\gamma T(o)A}{[1 + T(o)A]^2 + 1/\gamma^2} \quad \ldots\ldots\ldots (6)$$

$$\therefore 1/\gamma = Q_F/Q_{max} = \sqrt{\{[1 + T(o)A]^2 - 2\}} \quad \ldots\ldots (7)$$

where $Q_F = \omega_0/2\Delta\omega_0 = 1/2\delta = Q_{max}/\gamma$ is defined in terms of the frequencies at which $|A'/A| = 1/\sqrt{2}$.

For $T(o)A \gg 1$: $Q_F \simeq Q_{max}\,T(o)A$ $\ldots\ldots\ldots\ldots\ldots\ldots$ (8)

$$|A'/A| = \sqrt{\left\{\frac{1 + \gamma^2}{1 + \gamma^2[1 + T(o)A]^2}\right\}} = \sqrt{\left[\frac{1 + 4Q_{max}^2\delta^2}{1 + 4Q_F^2\delta^2}\right]} \quad (9)$$

$$\therefore |A'/A|_{dB} = 10\log(1 + 4Q_{max}^2\delta^2) - 10\log(1 + 4Q_F^2\delta^2) \quad \ldots\ldots\ldots\ldots (10)$$

Tan ϕ =

$$\frac{-T(o)A}{\gamma[1 + T(o)A] + 1/\gamma} = \frac{(1 - Q_F/Q_{max})}{Q_F((\rho^2-1)/\rho) + \dfrac{\rho}{Q_{max}(\rho^2-1)}} \quad \ldots\ldots (11)$$

Simplifying:

$$|A'/A|_{dB} = -10\log(1 + 4Q_F^2\delta^2) \quad \ldots\ldots\ldots (12)$$

with an error of 1 per cent if: $0{\cdot}01 = \log(1 + 4Q_{max}^2\delta^2)$

$$\therefore ((\rho^2 - 1)/\rho) = 0{\cdot}15/Q_{max}$$

For $Z_0 = \infty$, $Z_1 = 0$ and $n = 1$: $0{\cdot}744 < \rho < 1{\cdot}344$.

$$|A'/A| = \frac{1}{\sqrt{(1 + 4Q_F^2\delta^2)}},\ \tan\phi = -2Q_F\delta \quad \ldots. (13)$$

which is identical to the modulus and argument of β', equation (3).

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An Electronic Security System

Decca Radar Limited are supplying an entirely new form of electronic alarm for the detection of intruders and saboteurs. The first installation, capable of detecting any movement within a specific volume of space, has been made at Stoke Heath prison in Shropshire, where it is now being evaluated.

The Decca system employs Doppler techniques to enable movement to be detected by ultrasonic transmissions. Any movement, however slight, which occurs in the area being protected, results in the generation of ultrasonic Doppler frequencies, which are detected and cause an alarm to operate. The design of the equipment makes it equally suitable for the protection of a single vulnerable area or for several areas; with multiple installations any trouble spot is immediately pinpointed on an illuminated diagram at a central control point.

A high degree of reliability is provided with good protection against false alarms; the equipment is also inherently resistant to sabotage. Trials of the first installation have shown that repeated attempts to deceive it can be withstood, not only by those who did not know of the equipment's location but also by men familiar with its characteristics.

Large and complex groups of buildings can be effectively monitored from a single control position. The basic elements of the Decca system are now ready for production, and its potential for military applications as well as in the civil security field is being evaluated.

The Decca Intruder Detector is suitable for the protection of practically any type of room, doorway, corridor, or a particular section of floor. Another version has been developed for the protection of windows. The pattern of radiated ultrasonic energy is fan-shaped near the floor; as the performance is independent of volume coverage the contents of a protected room can be changed without having to adjust the equipment.

The whole system employs transistors and other solid state devices to give high reliability. All circuits are fail-safe, and sabotage will cause the appropriate alarm signal to operate. The design is based on replaceable plug-in circuit boards which can easily be changed for maintenance purposes.

The sensing heads incorporate transducers in the form of tubular ceramic elements for ultrasonic transmission and reception. These are arranged to give a wide arc fan coverage up to 15ft radius horizontally. Alternatively the beam can be arranged to cover windows, safes and valuable equipment.

The alarm signal can be remoted to any required distance. The central cabinet used in multi-point installations receives signals from the separately installed sensing heads. The cabinet contains detection units, false alarm filters, battery and mains operated battery charger. The battery is floated continuously and maintains the equipment in operation without interruption in the event of a mains failure. A 'hold' device is incorporated with the detector units so that, once started, the alarm remains on until reset by a push button.

A Simple Balloon Technique for Measuring the Radiation Patterns of Radio Aerials

By C. S. L. Keay*, M.Sc., A.Inst.P., and R. E. Gray†, M.Sc.

The radiation patterns of aerial systems operating at frequencies from 20Mc/s to 500Mc/s are readily measured by using a small self-contained transmitter and dipole suspended from a meteorological balloon. The balloon is tethered and controlled by four nylon lines which constrain it to move in a constant-radius arc centred on the aerial system under investigation. A complete pattern accurate to within 10 per cent may be obtained very quickly—sometimes in less than an hour, depending on the working frequency and the amount of fine detail required.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

In most radio communications systems it is desirable that the aerials should be directed along the optimum propagation path, with as little lateral or vertical radiation as possible in order to make the best use of the available power. The only conclusive way of checking this requirement is to measure the actual radiation patterns *in situ.* Such measurements, in which the aerial and its surroundings are considered as an integrated radiating system, not only provide a check on the aerial design but also indicate whether there is any need for adjustments.

In the past, h.f. and v.h.f. aerial measurements have been made either on the actual full-sized system by using aircraft or on a scale model of the aerial and its surroundings. The techniques in current use are classified and described in a survey paper by Cumming[1].

The use of aircraft is expensive and demands accurate knowledge of the aircraft's position as a function of time. If the airborne test aerial is mounted on the aircraft, interaction between it and the aircraft severely complicates its radiation pattern in a manner that is not easily allowed for. This complication may be minimized by towing the test aerial on a stabilized drogue at the end of a long cable[2]. If, however, the position of the test aerial is determined from the position of the aircraft, by radar fixes for example, allowance must be made for the drogue drifting in the wind. Furthermore, if the flight is made in a straight line at constant height, as is often the case, the receiver and measuring equipment must be capable of handling wide variations in signal strength, especially when highly directional aerial systems are being measured. Generally speaking aircraft techniques are best when the design frequency of the aerial is under 20Mc/s.

The use of scaled models involves close tolerances in dimensions as well as in the conductivity of the ground plane. If the scaling factor is large this method may prove to be impractical, especially for Yagi aerials which are by their nature very dependent on the self-reactance of the various elements, and hence on the element diameters[1]. Reproducing to the correct scale the surroundings of the aerial at its final site often rules out the use of scaled models, *in situ* measurements being the only alternative.

In a recent paper Brueckmann[3] has shown how radio transmissions from a satellite were utilized for determining the radiation pattern of a large multiple-beam array, and advocates the launching of an aerial calibration satellite. This approach may have advantages for calibrating some specific aerial systems, but lacks the flexibility needed to cope with a wide variety of aerial designs and frequencies.

* *University of Canterbury, Christchurch, New Zealand.*
† *Now at Royal Alexandra Hospital for Children, Sydney, N.S.W., Australia*

General Description

This article discusses the merits of a hitherto unpublished, inexpensive technique, whereby a small self-contained transmitter and dipole aerial is suspended from a

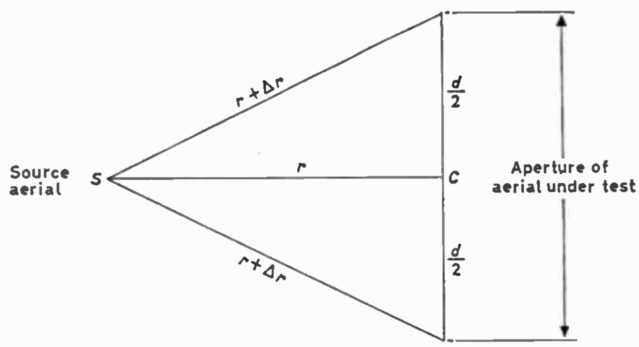


Fig. 1. Diagram for calculating the variation in phase across the aperture of an aerial

meteorological balloon flown at constant distance from the centre of the aerial array to be measured. The variation in signal strength indicated by a receiver connected to the array gives a direct measure of the radiation pattern as the position of the balloon is varied. By the reciprocity theorem the radiation pattern of an aerial is the same whether it be transmitting or receiving, provided non-linearities are absent. Thus radiation pattern measurements may be carried out equally well by probing the transmitted field of the aerial or by measuring the angular variation of the aerial's response to an incident wave. The latter approach was adopted in this case simply because it is much easier to construct a lightweight transmitter than lightweight receiving and recording equipment.

An important advantage of the balloon technique over most methods employing aircraft lies in the fact that the balloon and its cables can be kept entirely non-metallic. The absence of nearby reflecting surfaces preserves the cylindrically symmetrical free-space radiation pattern of the balloon-borne dipole, thereby eliminating a troublesome correction factor which would otherwise be involved.

Basic Requirements

In general, measurements of the spatial distribution of the far-field (true radiation pattern) of an aerial system must be undertaken with a test device which is effectively outside the induction field (or near-field) of the aerial. At a distance of 16 wavelengths from a current element the induction field is down to 1 per cent of the radiation field: a negligible contribution for most practical measurements.

Furthermore, all measurements of the far-field must be carried out at a sufficient distance from the aerial system to ensure that the true free-space radiation pattern is approached. To meet this requirement the incident wave from any external source must be very nearly uniform in amplitude and phase across the effective aperture of the aerial system under test. The variations which can be tolerated across the width of the aperture are commonly taken to be $\pi/8$ radians in phase and 1/4dB in amplitude[1].

If d is the effective subtended aperture of the aerial under test and r is the distance to the external source aerial as shown in Fig 1, then

$$\Delta r = d^2/8r$$

neglecting terms in Δr^2. The effective subtended aperture

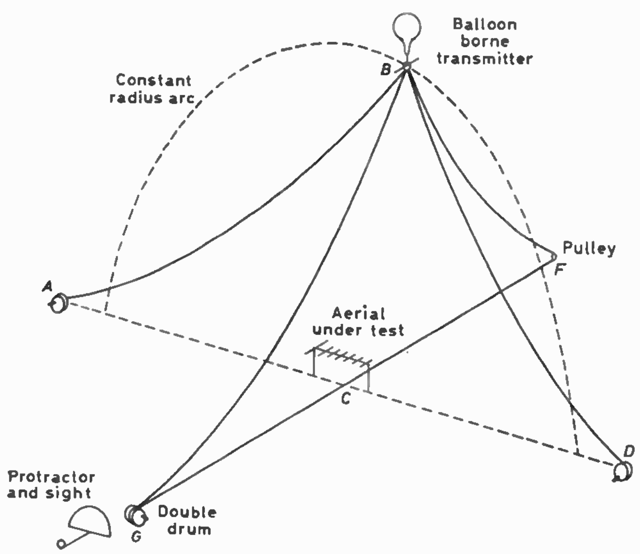


Fig. 2. Physical layout of the cables and sighting apparatus for controlling the balloon and measuring its position

d refers to the aerial dimension perpendicular to the distance r, and as the position of the source is changed during a series of measurements so the value of d is liable to change.

The condition that the phase difference between the incident wave at the edge of the array and that at the centre be not greater than $\pi/8$ radian is given by

$$2\pi\Delta r/\lambda \leqslant \pi/8$$

Eliminating Δr from the previous equation gives the condition

$$r \geqslant 2d^2/\lambda$$

which must be satisfied for all positions of the source.

Nevertheless this condition may be relaxed whenever measurements are made at fixed radius in a plane containing the centre of the aerial array under test. The minimum radius given by the above equation is obtained by taking for the aperture width d the maximum dimension of the array in the plane of measurement, even if the array length is considerably greater normal to the plane of measurement.

The results will be quite valid provided the phase *differences* between the various contributions from along the length of the array remain constant within the above limits for all positions of the source. When dealing with extensive arrays this property of fixed-radius measurement techniques can be very advantageous.

The Balloon Technique

Consider a balloon-borne transmitter B tethered at two points F and G equidistant horizontally from the centre C of an aerial under test, as shown in Fig. 2. Measurements of the radiation pattern of the aerial are to be made in the plane containing B and C normal to the line joining F and G. The tethering cables BF and BG, provided they are fairly light, constrain the balloon to a constant radius arc within the measurement plane as the lengths of the traversing cables BA and BD are altered. This is not at all difficult to achieve provided A and D are located several yards outside the arc of constant radius and the traversing cables BA and BD are not pulled so tightly that the tethering cables BG and BF slacken.

If a light wind is blowing the balloon is kept within the plane of measurement by a small differential adjustment of the tethering cables. These cables, BF via a pulley

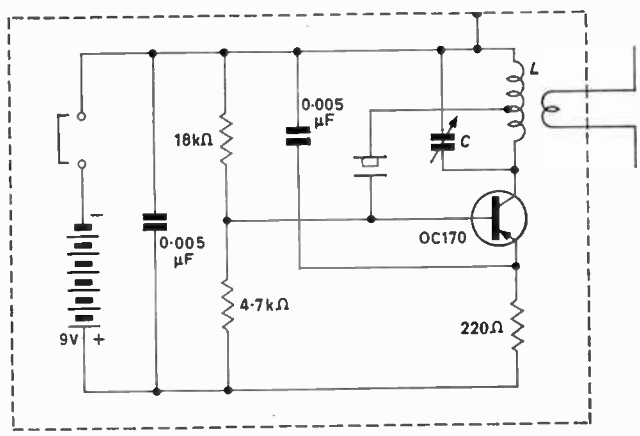


Fig. 3. 69Mc/s balloon-borne transmitter

at F, and BG, are returned to opposite sides of a double winding drum at G to enable a single operator to wind one cable in as the other is wound out. This adjustment has the effect of altering the cable tensions (to cope with the wind force) by varying the relative amounts of sag in BG and BF. For full effectiveness in this respect the distance FG should be approximately 2 to 2½ times the arc radius. It should also be mentioned that the tethering cables BG and BF are attached one at each end of the balloon-borne transmitter in order to provide a restoring couple for keeping the transmitting dipole in correct alignment.

The winding-drum operators at A and D alter the length of the traverse cables under instructions from a controller at G who monitors the balloon's position with the aid of a protractor and cross-wire sight. Using a large transparent protractor the angle BCD may be obtained to well within one degree, and if concentric circles are engraved on the protractor the radius of the arc followed by the balloon may also be checked, and adjustments or allowances made accordingly.

The signal level from a receiver connected to the aerial being tested should be displayed near the controller at G who can then record signal strength as a function of balloon elevation. The receiver should have a dynamic range of the order of 60dB if best results are to be obtained with high gain aerials. A receiver with logarithmic response is ideal but a conventional type of receiver with linear response is perfectly satisfactory provided that it is

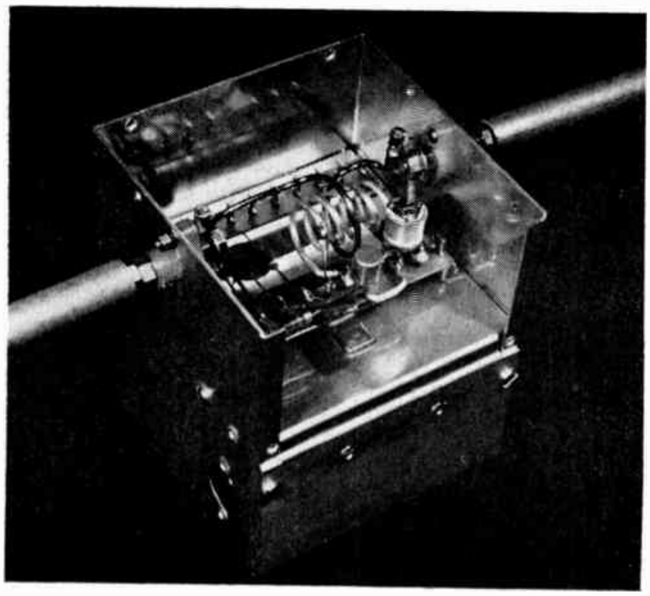


Fig. 4. Balloon-borne transmitter

well shielded and a calibrated attenuator is available to insert in the input for extending the receiver's dynamic range without risk of overload.

The Balloon-Borne Transmitter

The balloon-borne transmitter should be kept as simple as possible, consistent with good stability. A circuit employing a single transistor delivering 10mW of r.f. power output will be ample for almost any aerial measurement using this technique. The circuit of the source transmitter used in obtaining the results about to be described is shown in Fig. 3. It uses a single OC170 transistor operated in common emitter mode with a small overtone crystal as a frequency determining feedback element. The simplicity of the transmitter construction is revealed by the photograph shown in Fig 4. The underneath compartment contains the battery which was of sufficient size to sustain an r.f. power output of 12mW for at least twelve hours without any detectable departure from constancy. The variation of output power with changes in ambient temperature was not very great and its effect, if present, was largely cancelled by immediately repeating every balloon run in the opposite direction.

Data on Balloons

Balloons capable of lifting up to 57kg when filled with hydrogen are readily available as they are used constantly in meteorological work. The sizes of typical balloons and their lifting capacities are given in Table 1.

The Nylon Cables

Thin nylon line is by far the best material for the tethering and traversing cables. It is light yet strong and does not absorb moisture. Its low surface friction enables it to slide over the ground and snag less readily than string or twine, but does make extra care necessary whenever knots need to be tied.

A 45-pound nylon line was found to be very suitable for use with 800 or 1 000g balloons. 3 000ft of this line weighs only half a kilogram (density of nylon is about $1{\cdot}14\text{g.cm}^{-3}$) and is sufficient to allow, from the geometry of Fig. 2, an arc radius of 500ft. This leaves ample lift for a transmitter and dipole, and for providing the tension in the cables necessary for good horizontal control of the balloon.

TABLE 1
Balloon Sizes

EMPTY WEIGHT (*grams*)	INFLATED VOLUME (*metres*^{3})	(*feet*^{3})	PAYLOAD CAPACITY (*kg*)	(*lb*)
650	3·2	115	2·6	5·7
800	4·3	150	3·5	7·7
1000	5·8	205	4·7	10·4
1750	12·0	425	10·4	23·0
2400	18·7	660	16·5	36·4
7000	63·7	2250	57	126

Results

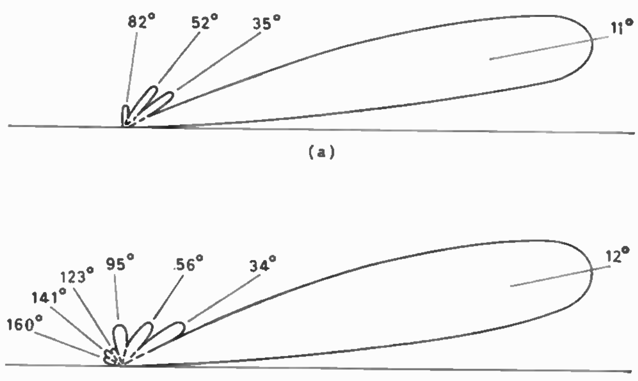
The majority of the measurements using this technique were made at frequencies near 69Mc/s on a variety of aerial systems. A typical result is shown in Fig. 5, which represents the vertical radiation pattern of an array of twelve horizontal half-wave dipoles spaced one eighth wavelength in front of a wire net reflecting screen. The configuration of the dipole is shown in Fig. 6.

It is apparent from Fig. 5 that the actual radiation pattern follows the calculated pattern very closely, with most of the radiated energy being sent in the intended direction. A small amount of energy is lost to the rear due to an imperfect reflector screen and the fact that it did not extend high enough to be fully effective for the uppermost row of dipoles. The high elevation lobes of the pattern are shifted somewhat as a consequence of this.

Fig. 5(b) is the average of the results from two complete traverses of the array by the balloon; one in each direction. The difference of one degree between the calculated and measured values of elevation of the main lobe might be due to experimental errors although it could have arisen from a faulty design assumption concerning the depth below the surface of the ground at which reflection occurs. However, it is obvious from Fig. 5 that this particular aerial system is performing very close to expectations.

It is not difficult, using this balloon technique, to obtain results which have an overall accuracy of better than 10 per cent. This figure depends mainly on the signal strength calibration accuracy of the receiver and attenuator, as well as on the ability of the field personnel to keep the balloon close to the required flight-path. In this respect it is worth mentioning that even the poorest of the balloon flights yielded results which were better than those obtained when a helicopter was employed for making the same measurements.

Fig. 5. Vertical radiation pattern of a 4 × 3 array at 69Mc/s
(a) Calculated
(b) Measured



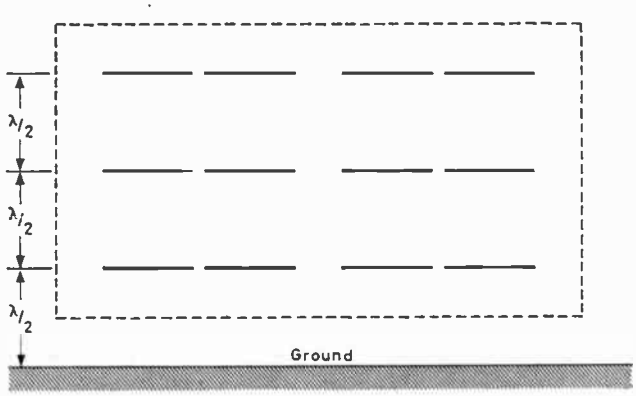


Fig. 6. Configuration of the 69Mc/s 4 × 3 dipole array

Furthermore the balloon technique is very inexpensive in terms of manpower and equipment when compared with other techniques. Under favourable weather conditions it is normal for a team of four to measure both the vertical and horizontal radiation patterns of an aerial system in well under half a day. This assumes that the necessary apparatus is properly prepared and the positions *A, D, F,* and *G* (Fig. 2) have been surveyed and pegged out beforehand.

The range of frequencies at which it is profitable to use a balloon technique extends from roughly 20Mc/s to 500Mc/s. Below 20Mc/s the necessary arc radius exceeds 800ft, which is about the limit for satisfactory control over a tethered balloon. Above 500Mc/s the arc radius becomes small enough to permit the use of rigid non-metallic booms for supporting the source transmitter and dipole.

Acknowledgments

The authors wish to thank the New Zealand Meteorological Service for their valuable advice on the handling of hydrogen-filled balloons.

This work was required for a programme of meteor research sponsored by Grant NsG-219-62 from the National Aeronautics and Space Administration, U.S.A.; and was aided by the New Zealand Universities Research Grants Committee.

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The Allen Clark Research Centre

H.R.H. The Duke of Edinburgh opened the Allen Clark Research Centre of the Plessey Group at Caswell, Towcester, on Friday, 20 March, 1964.

The research laboratories at Caswell were first established in November, 1940, by the late Sir Allen Clark, founder and Chairman of The Plessey Company Limited.

Many important developments—especially materials for electronics and telecommunications—have taken place in the laboratories. Today the programme is mainly directed at research into new purity materials, magnetic and dielectric materials, radar and high temperature materials plus solid state chemistry and physics development. Out of a total payroll of 50 000 people in the Plessey Group, over 4 000 are employed on research and development.

In addition to the central activities at the Allen Clark Research Centre, special research programmes are carried out at the Group's establishments at Roke Manor (Plessey-UK) and Taplow (British Telecommunications Research Limited, a subsidiary of AT & E). There are also extensive facilities for environmental testing, especially of professional components at Titchfield.

The laboratories at Caswell were set up by a small team of seven men who had been working in materials research at the company's laboratory in Ilford, and six of these founder members—including the Director, Mr. G. C. Gaut—are still associated with the establishment today. The nucleus of seven men has grown to some 360, of whom about 40 per cent are graduate scientists.

Caswell House was originally a private residence in 230 acres of ground and was itself the home of some of the early research, but stables and other buildings were gradually converted into laboratories and workshops.

By 1956 all available space had been absorbed—and the first new building (of 4 000ft^2) was added. Expansion continued until during the last year of Sir Allen's life, the plans were laid by him personally for a basic research block. It is this building which H.R.H. The Duke of Edinburgh commemorated to Sir Allen at the opening ceremony on 20 March.

To all those who knew him, this is a most appropriate memorial. It perpetuates not simply his interest in research, but his quest for new knowledge as the basis for his actions and decisions in taking his company into new fields of technology.

The development of any new component at Caswell starts with a thorough investigation into the basic chemistry of materials involved. This is followed by an equally comprehensive study of the physical properties before mechanical problems are considered. In this way complete background information is obtained on any new component or process, greatly assisting future development.

When the process and a device has reached a technically acceptable stage, a pilot plant operation is initiated to develop the process mechanically and produce quantities of the device for life testing and market assessment. It is usual, and preferable to carry out the pilot plant operation near the production unit likely to take over the work. Background information and liaison continuity is provided by transferring the technicians concerned, together with the experimental equipment which they used to achieve the results, to the factory so that there is complete integration between the research and development and production stages.

Most of the scientists at the laboratory are chemists and physicists, in approximately equal numbers. They are split into the following groups, each with the appropriate number of technicians:

Research	New high purity materials
	Magnetic and dielectric materials
	High temperature materials
	Radar materials
	Solid state chemistry
	Solid state physics
Development	Semiconductor device development
	Semiconductor pilot plant

Administratively, the laboratory is self-contained. The services cover all aspects of its business, including accounting and the production of technical papers and reports. A well-equipped central technical service deals with all analytical and engineering requirements.

Scientific information and advice is provided by maintaining close contact with the universities through a consultant staff of lecturers and professors.

The laboratories which now become the Allen Clark Research Centre have always occupied a leading position in industrial research and development. For example, in August 1956 the first active research in the field of solid circuits began there in collaboration with the Royal Radar Establishment, Malvern. Other important work includes thin film base electrode ceramic capacitors; solid track volume controls; tantalum capacitors and high temperature silicone rubber for plugs and sockets. The centre is also an authority on radar absorbing materials. Recent achievements in entirely new materials include vitreous carbon and dense silicon nitride.

Analysis of a Feedback Amplifier

By C. D. West*, B.Sc.

The analysis of a pulse amplifier circuit is presented. The circuit, which has a low temperature coefficient of gain, consists of a d.c. coupled difference amplifier and a common emitter stage. Component values are given for an amplifier with a gain of 10 and a temperature coefficient of $50 \times 10^{-6}/°C$. Several stages may be cascaded.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE transistor circuit shown in Fig. 1 was suggested by Collinge[1] and Lloyd[2] to construct a stable pulse amplifier. It is believed that this circuit configuration has several attractive features; in particular it exhibits a smaller change of gain with temperature than amplifiers previously described.

There is no direct equivalent of the circuit although the principle of a d.c. coupled amplifier with low d.c. gain and high a.c. gain has been used to construct a very stable valve amplifier[3].

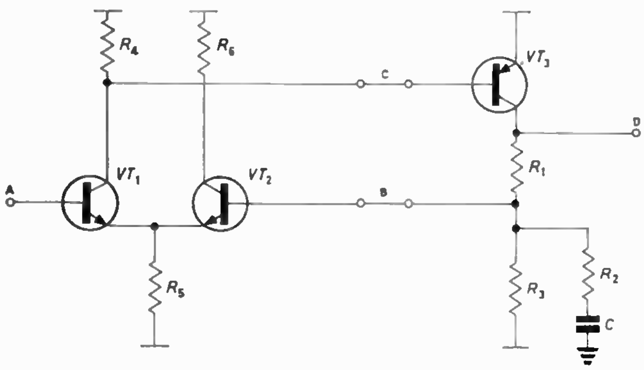

Fig. 1. The basic loop

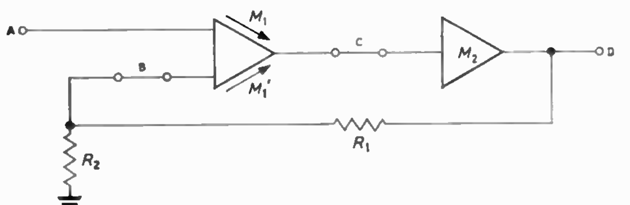

Fig. 2. Block diagram of the loop

Analysis

General Expression for the Feedback Gain

Consider the loop drawn as in Fig. 2. The voltage gain between the input (A) of the difference amplifier and its output (C) is M_1 and between its input (B) and output (C) is M_1'. The gain between the input (C) and output (D) of the amplifying stage is M_2.

Since the feedback voltage appears at (B) rather than directly at the input terminal, a modification of the usual picture of a virtual earth point at the input where feedback and input voltages are added must be used. The general method given by Vincent[4] may be applied.

However, from Fig. 2:

$$V_o = M_2 (M_1 V_1 + M_1' V_f)$$

where V_1 is the input voltage at (A), V_f is the feedback voltage at (B), V_o is the output voltage at (D). Suppose that the difference amplifier is not perfect, but that $M_1' = -M_1\gamma$ where γ is a constant approximately equal to unity.

Then:

$$V_o = M_2 (M_1 V_1 - M_1 \gamma V_f)$$

now if the input impedance at the base of VT_2 is much greater than R_2, and neglecting the reactance of C,

$$V_f \simeq V_o \cdot \frac{R_2}{R_1 + R_2}$$

(strictly R_2 should be replaced by R_2 in parallel with R_3) so:

$$V_o \simeq \frac{M_1 M_2 V_1}{1 + (\gamma R_2/(R_1+R_2))\, M_1 M_2} \quad \ldots\ldots\ldots (1)$$

Writing $\beta = \gamma R_2/(R_1+R_2)$ and identifying it with the feedback factor and $M_1 M_2 = M_o$, the open loop gain of the amplifier, one may write the feedback gain, called M_f, as:

$$M_f = \frac{M_o}{1 + \beta M_o} \quad \ldots\ldots\ldots (2)$$

To assess the temperature performance of the circuit one requires M_o, the open loop gain and β, the feedback factor and their temperature coefficients.

Calculation of Open Loop Gain, M_o

The open loop gain is the gain between input A and output D if the link is broken at B, (Fig. 1). On the left-hand side of the break is put a resistance R_2 to earth, and on the right-hand side an impedance, equal to the input impedance of VT_2, at its base, to earth. This last impedance is much larger than R_2 and negligible in this calculation.

The difference amplifier is treated as a common emitter transistor VT_1, perturbed by the presence in its emitter lead of the input impedance of VT_2 in parallel with R_5. VT_2 in its turn is treated as a common base transistor perturbed by the presence of R_2 in its base lead (see Fig. 3).

R_3 and R_1 + output impedance of VT_3 are neglected, since they are much larger than R_2. Finally, VT_3 is a transistor in common emitter.

The common emitter hybrid parameters of VT_1 and VT_2 are called h^* and h^{**} respectively. An approximate expression for M_1 may then be found as outlined above giving:

$$M_1 \simeq \frac{-\alpha^* Z_{L1}}{h_{ii}^* + \dfrac{1+\alpha^*}{1+\alpha^{**}}(h_{11}^{**} + R_2)} \quad \ldots\ldots (3)$$

Z_{L1} is the load of the difference amplifier, consisting of the parallel combination of R_4 and Z_3, the input impedance of VT_3.

If $R_2 \ll h_{ii}^{**}$ and $\alpha^* \gg 1$ this equation reduces further to:

$$\frac{-Z_{L1}}{h_{ib}^* + h_{ib}^{**}} \quad \ldots\ldots\ldots (4)$$

Here h_{ib}^* and h_{ib}^{**} are the common base parameters of VT_1 and VT_2. It is easily shown that M_2, the voltage gain of VT_3, is given by:

* *University of Liverpool.*

$$M_3 = \frac{-\alpha'}{Z_3\,((1/Z_{L3}) + h_{22}')} \quad \ldots\ldots\ldots\ldots (5)$$

where h' are the common emitter parameters, Z_{L3} is the load and Z_3 is, as explained above, the input impedance of VT_3. M_o is obtained from equations (3) and (5).

$$M_o \simeq \frac{\alpha'\alpha^* R_4}{\left\{h_{11}^* + \frac{1+\alpha^*}{1+\alpha^{**}}(h_{11}^{**} + R_2)\right\}\{(1/Z_{L3}) + h_{22}'\}\,\{R_4 + h_{11}'\}} \quad \ldots\ldots\ldots\ldots (6)$$

This expression has been compared with results obtained experimentally from circuits containing transistors of known (measured) parameters. Normally $R_2 \ll h_{11}^*$ and for these measurements and calculations R_2 was made equal to zero.

The temperature coefficient of open loop gain is obtained by differentiating equation (6) with respect to temperature. For most of the transistor types tested the manufacturers gave h_{fe} (α'), h_{ib} (h_{11}) and h_{ob} (h_{22}) as a function of temperature. Accordingly the expression will be most convenient in terms of the temperature coefficients of these parameters.

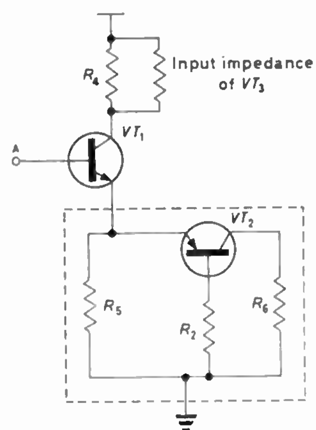

Fig. 3. The difference amplifier

Many modern transistors have a small tolerance on h_{ib}, say ± 15 per cent, and so $h_{ib}^{**} \simeq h_{ib}^*$.

Writing h_{ib}' etc. as common base parameters of VT_3:

$$1/M_o\,(\partial M_o/\partial T) \simeq (1/\alpha')\,(\partial\alpha'/\partial T)\,(1 - h_{22}'Z_{L3} - h_{11}'R_4) - 1/h_{ib}^*\ (\partial h_{ib}^*/\partial T) - (h_{11}'/R_4)\,.\,(1/h_{ib}')\,(\partial h_{ib}'/\partial T) - h_{22}'Z_{L3}\,.\,(1/h_{ob}')\,(\partial h_{ib}'/\partial T) \quad \ldots\ldots\ldots (7)$$

Now for the transistors finally chosen, the first (positive) term is about $+0{\cdot}5$ per cent/°C and the sum of the other (negative) terms is about $-0{\cdot}3$ per cent/°C. Thus small variations in the temperature coefficients of the parameters lead to quite large variations in the temperature coefficient of M_o. The transistor types were, of course, chosen for just this reason, to make the temperature coefficient of M_o as small as possible while maintaining a high minimum value of βM_o.

Calculation of β, The Feedback Factor

It has been shown that $\beta = \gamma R_2/(R_1 + R_2)$. In fact R_2 is shunted by R_3 and the input impedance of VT_2 at its base.

R_3 has a negligible temperature coefficient in parallel with R_2 and so allowance for its presence is made simply by assuming that for the numerical value of R_2 one writes the value of $R_2 \| R_3$. However, both the input impedance mentioned, $Z_{in(2)}$, and γ are temperature dependent and must be calculated.

Calculation of γ

γ was defined by $M' = -\gamma M_1$. Assume a voltage v at the input A. The output of the difference amplifier is $M_1 v$. With a voltage at the inputs A and B the output is $(M_1' + M_1)v$. Defining δ the common mode rejection ratio, as:

$$\frac{\text{Output with a signal at one input only}}{\text{output with same signal at both inputs}}$$

$$\text{It is seen that } \delta = \frac{M_1 v}{(M_1' + M_1)v} = \frac{1}{1-\gamma}$$

so:

$$\gamma = 1 - (1/\delta)$$

If δ is calculated then γ may easily be found.

In a difference amplifier of the type used, the common mode rejection ratio would not be infinite even if its transistors were identical and the emitters were fed from a true current source, since the output impedance of the transistor is finite. In practice, this effect can be ignored compared with those arising from dissimilarities of the transistors, and finite impedance of the current source in the emitters.

It can be shown that, considering this effect only,

$$\delta \simeq (R_5/h_{ib}^*)$$

so:

$$\gamma \simeq 1 - (h_{ib}^*/R_5) \quad \ldots\ldots\ldots\ldots\ldots\ldots (9)$$

$Z_{in(2)}$, the input impedance (with feedback) at the base of VT_2 is approximately:

$$Z_{in(2)} \simeq \alpha^{**}\,\alpha'\,R_2 \quad \ldots\ldots\ldots\ldots\ldots (10)$$

providing the external load is $\gg R_1 + R_2$. Allowing for R_3 shunting R_2, as explained above:

$$\beta = \frac{\gamma(R_2 \,\|\, Z_{in(2)})}{R_1 + (R_2 \,\|\, Z_{in(2)})}$$

The temperature coefficient of β is obtained by differentiating this expression after substituting equations (9) and (10), for γ and $Z_{in(2)}$.

$$1/\beta\,(\partial\beta/\partial T) \simeq \frac{1-\beta}{\alpha^*\alpha'}(1/\alpha'\,(\partial\alpha'/\partial T) + (1/\alpha^*)\,(\partial\alpha^*/\partial T)) - (h_{ib}^*/R_5)\,((1/h_{ib}^*)\,(\partial h_{ib}^*/\partial T)) \quad \ldots\ldots\ldots (11)$$

There is also to be included in the expression for the temperature coefficient of β that due to the resistors forming the feedback network if their variation with temperature is not negligible.

One now has expressions for the temperature coefficients of M_o and β and it is possible to write down the overall temperature coefficient of the amplifier.

The usual expression for the variation of M_f (the feedback gain) with M_o and β applies here, i.e.:

$$1/M_f\,(\partial M_f/\partial T) \simeq (1/\beta M_o)\,.\,(1/M_o)\,(\partial M_o/\partial T) - (1/\beta)\,(\partial\beta/\partial T) \quad \ldots\ldots\ldots\ldots (12)$$

if $\beta M_o \gg 1$.

Combining equations (7), (11) and (12):

$$1/M_f\,(\partial M_f/\partial T) \simeq (2h_{ib}^*/\alpha' R_2)\,\{1/\alpha'\,(\partial\alpha'/\partial T)\,(1 - h_{22}'Z_{L3} - (h_{11}'/R_4)) - 1/h_{ib}^*\ (\partial h_{ib}^*/\partial T) - (h_{11}'/R_4)\,.\,(1/h_{ib}')\ (\partial h_{ib}'/\partial T) - (h_{22}'Z_{23}/h_{ob}')\ (\partial h_{ob}'/\partial T)\} - \frac{(1-\beta)}{\alpha^*\alpha'}\{1/\alpha'\,(\partial\alpha'/\partial T) + (1/\alpha^*)\,(\alpha\partial^*/\partial T)\} + (h_{ib}^{**}/R_5)\,\{(1/h_{ib}^*)\,(\partial h_{ib}^*/\partial T)\} \quad \ldots\ldots\ldots\ldots (13)$$

The first term is normally by far the largest, and given the nominal temperature coefficients and parameters one may compute the expected temperature coefficient of M_f neglecting the 2nd and 3rd terms, which are due to change of β. These terms are small and approximately equal in magnitude and of opposite sign.

Other Effects

Two more causes of variations in output voltage will be considered, although strictly they are not variations in gain.

If the source impedance of the device from which the amplifier is fed is not zero (i.e. not a true voltage source) it will form a potential divider with the input-impedance of the amplifier, i.e. the impedance seen at the base of VT_1. Any bias resistor put in at this base must be con-

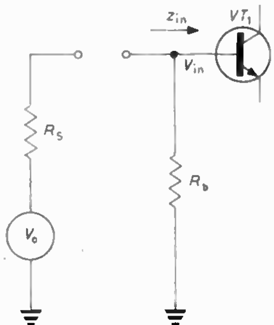

Fig. 4. The input circuit

sidered as in parallel with Z_{in}. See Fig. 4. Clearly:

$$v_{in}/v_o = \frac{R_b \| Z_{in}}{R_s + (R_b \| Z_{in})}$$

so:

$$\frac{1}{(v_{in}/v_o)} \cdot \frac{\partial(v_{in}/v_o)}{\partial T} \simeq \frac{R_b R_s}{R_b + R_s} \cdot (1/Z_{in}) \cdot (1/Z_{in})(\partial Z_{in}/\partial T)$$

R_b then behaves as if it were in parallel with R_s.

Z_{in} increases with temperature ($Z_{in} \simeq \alpha'\alpha^* R_2$) giving an apparent increase in gain.

In practice R_s or R_b is made small, e.g. R_b may be a 100Ω matching resistor. In that case this effect is also very small, typically $\sim 1/5$ that due to changes of M_o.

Finally, consider changes in the output impedance, Z_{out}, of the amplifier. Without feedback it is approximately $R_1 + R_2$, and with feedback it is easily seen to be $1/(1+\beta M_o)$ times this.

So if $\beta M_o \gg 1$

$$Z_{out} = \frac{R_1 + R_2}{\beta M_o}$$

This decreases with temperature. If it is feeding a low impedance, i.e. one which is not much greater than Z_{out} it will therefore give rise to another apparent increase in gain with temperature. Typically, however Z_{out} is less than 100Ω, so another similar stage may be fed without any great drift in gain from this cause, provided the next stage has a large bias resistor, say 10kΩ.

Results

Several of the circuit parameters derived were confirmed experimentally and some of the results are presented below.

TABLE 1

VT_1 and VT_2 are transistor types 2S103 (Texas Instruments)

	CALCULATED GAIN	EXPERIMENTAL GAIN
Sample 1	19·8	23
Sample 2	20·8	22·8
Sample 3	21	24·8
Sample 4	22	23

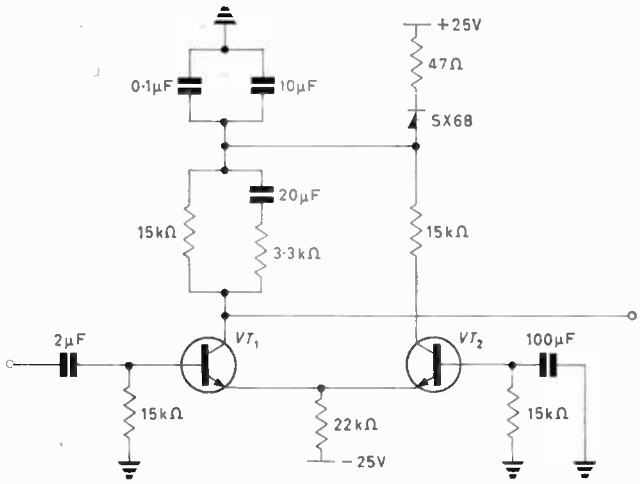

Fig. 5. Difference amplifier test circuit

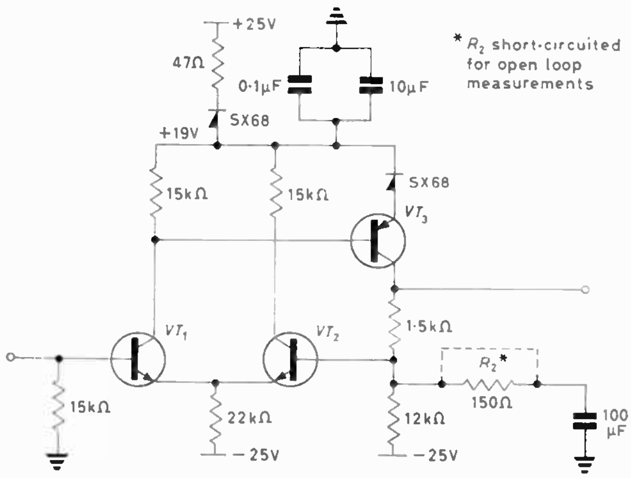

Fig. 6. Complete loop test circuit

Open Loop Gain

The calculated and experimental gain of the difference amplifier shown in Fig. 5 are compared in Table 1. Several samples of the transistors used were tested.

In Tables 2 and 3 the experimental and calculated open loop gain of the circuits shown in Figs. 6 and 7 are compared for three samples of each type of transistor.

TABLE 2

VT_1 and VT_2 are transistor types 2S103 (Texas Instruments)
VT_3 are transistor types AFZ11 (Mullard)

	CALCULATED GAIN	EXPERIMENTAL GAIN
Sample 1	320	360
Sample 2	700	750
Sample 3	1050	940

TABLE 3

VT_1 and VT_2 are transistor types 2S103 (Texas Instruments)
VT_3 are transistor types 2N726 (Texas Instruments)

	CALCULATED GAIN	EXPERIMENTAL GAIN
Sample 1	370	365
Sample 2	525	525
Sample 3	725	720

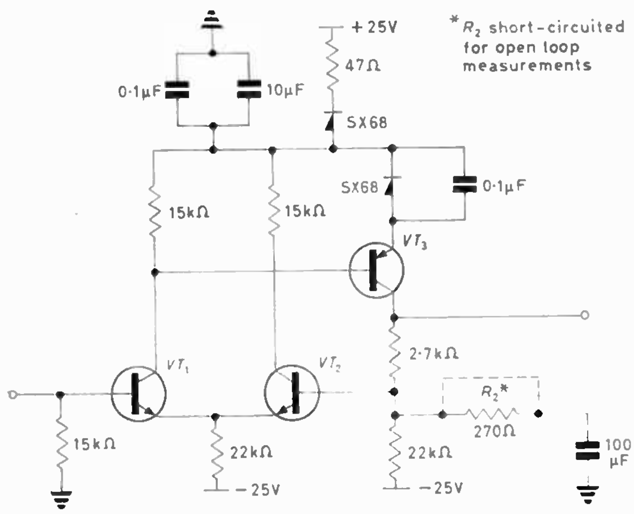

Fig. 7. Low temperature coefficient amplifier stage

Temperature Coefficient of Voltage Gain with Feedback

The temperature coefficient of the circuit of Fig. 7 with feedback was measured. VT_1 and VT_2 were transistor types 2N911 (SGS Fairchild) and VT_3 was type 2N1132 (SGS Fairchild). Six sets of transistors were used, with the same resistors, and gave a mean temperature coefficient of ·0048 per cent/°C with a standard deviation of ·0007 per cent/°C.

The temperature coefficient was calculated to be ·0030 ± ·0015 per cent/°C. In calculating this figure, the effect of the temperature coefficient of the high stability carbon resistors inside the loop on the operating point must be considered. The feedback resistors (Metallux) were expected to contribute a temperature coefficient of ·001 per cent/°C, and this is also included in the figure of ·0030 per cent/°C.

Conclusion

It has been shown possible to calculate some of the important properties of the amplifier and obtain results in good agreement with experiment. An amplifier stage with a gain of 10 has been designed and constructed on the basis of this analysis and showed the very low temperature coefficient of gain of 50 × 10^{-6}/°C.

Acknowledgments

The author is indebted to Mr. K. Aitchison and Mr. Monir Mohammed of the Nuclear Physics Research Laboratories, University of Liverpool, for their practical assistance.

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Periscopes and Television for 'Dragon' Project

A remote viewing system for use in the 'Dragon' reactor has recently been devised. 'Dragon' is a high temperature nuclear reactor project sponsored by the European Nuclear Energy Agency of the Organization for Economic Co-operation and Development.

The system devised by Barr and Stroud Ltd to meet the special conditions of viewing the interior of the reactor with its core temperature of 750°C comprises a combination of optical periscopes and lighting able to withstand the heat and radiation, and also capable of sending back images by way of a fully transistorized image-orthicon television camera placed far enough from the reactor to be protected from the temperature and radiation effects.

The whole of the system will be controlled from a desk-type console fitted with monitoring viewing screens and based in the reactor control room 80ft away.

Although Barr and Stroud Ltd have supplied reactor core viewing equipment at other atomic establishments, 'Dragon' presents new problems, many of which are due to the requirement for the equipment to remain permanently in the environment of the inside of the reactor vessel.

The object of 'Dragon' is to conduct experiments leading up to the construction of a land-based reactor of improved efficiency. The gas-cooled, carbon moderated principle was chosen. The gas and its working temperature differ from that of the conventional reactors being built in the U.K. at present to generate power. The use of helium as the gas in place of carbon dioxide posed special problems of diffusion through joints and even through some materials; the temperature above the core, 750°C, is more than double that of the current power generating stations, such as Berkeley, for which Barr and Stroud Ltd also supplied reactor core viewing equipment.

The requirement is to view inside the pressurized core or plenum chamber and also inside the connecting transfer chamber, immediately above the plenum chamber. Viewing is required of the surfaces inside these chambers, of the fuel handling machine in operation and in alignment, and of the fuel elements. Remote presentation is required in the control room 80ft away from the reactor pressure vessel. The equipment must, as it penetrates into the reactor, be sealed to the extremely high standard of leaktightness demanded of the reactor containment. It must also be capable of being serviced without depressurizing the reactor, an event which it is intended should not occur in under an interval of four years.

A contract for the design and manufacture of the viewing equipment was put out to European tender and placed with Barr and Stroud Ltd in 1962 by the U.K. Atomic Energy Authority, acting on behalf of the 'Dragon' project signatories.

The system chosen by the firm to meet the special conditions is an optical system and special lighting system protruding into the reactor and capable of withstanding the temperature and radiation. The materials of construction have to withstand high radiation levels and in the case of moving parts to be immune from seizure under the non-oxidizing conditions of a helium environment.

The optical system forms an image on an image-orthicon television camera placed at the outer end and thereby protected from unduly high levels of both temperature and radiation.

This type of camera tube was chosen because of the probable low light level of the image, namely less than a 1/1000 foot candle, owing to the low reflectivity of the carbon dust covering the objects to be viewed.

The apparatus is contained on the outside within pressure vessels forming part of the primary containment, and on the inside or reactor side within a pressure vessel fitted with a specially sealed window, which forms a secondary containment for use when servicing the equipment. All moving parts within the apparatus are electrically operated by remote control, as is the television camera; the electrical leads to the apparatus penetrate the primary containment pressure vessel through special seals. The permissible leakage rate from the equipment must not exceed 1ft³ of gas in 90 years and can only be measured by the most sensitve laboratory mass spectrometer.

The restricted available space gave rise to special problems of design. Possibly the biggest of these was the requirement to include a complete image-orthicon television camera and its circuits within the pressure vessels containing the periscope system. The problem was overcome by building the circuits round the camera tube, and fully transitorizing it.

The available aperture for the periscope to protrude through the reactor wall is 4½in diameter which affords a 2in diameter viewing window through which an area of 14ft diameter can be viewed at the core face 9ft away. The small aperture involved the need for a very compact but powerful illumination system and was met by the use of the recently developed quartz iodine lamps which are also able to withstand the temperature conditions.

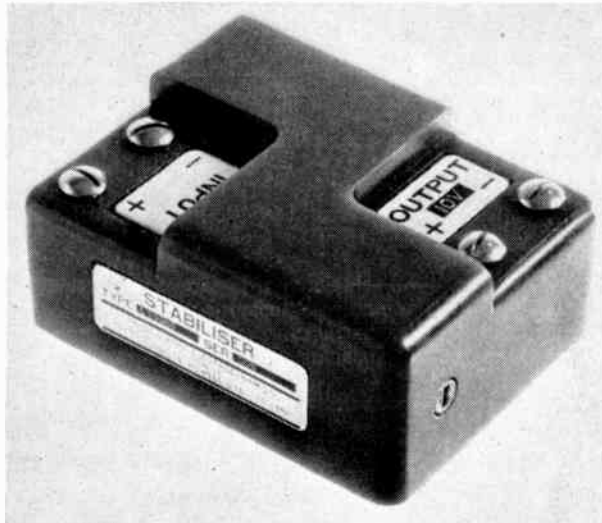
The control of the equipment is performed at a desk type console fitted with monitor viewing screens and is situated in the reactor control room.

A Simple, High-Performance Series Stabilizer

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

When stable d.c. voltages are required, the circuit designer has a large variety of circuit configurations from which to choose. His choice is dictated by the basic power parameters, the quality of performance and the allowable cost and whereas the majority of requirements are easily satisfied there occasionally arises a particular problem which requires a special solution. The article describes a circuit which was specifically designed to provide an inexpensive, compact, stable voltage source for exciting unbonded strain gauge pressure transducers.

(Voir page 353 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)



THIS note describes a simple series stabilizer circuit able to deal with a wide range of input voltage and particularly suitable for the condition when one extreme of the range approaches the minimal operating voltage of the series element.

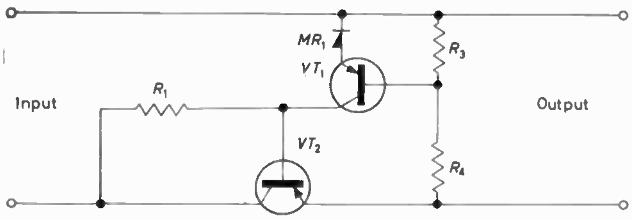


Fig. 1. Conventional series stabilizer

The deficiencies of the normal stabilizer circuit are examined and a number of modifications which might improve its performance are considered.

The circuit design for a particular application is carried out.

The Conventional Series Stabilizer

The conventional series stabilizer shown in Fig. 1 is well known and is derived from the equivalent valve/neon circuit. If one considers the operation of such a circuit under constant load but varying input voltage conditions one may encounter errors in output voltage for the following reasons:

(1) A wide change in input voltage produces a wide current change in VT_1. This current change produces a change of base-emitter voltage of the transistor.

(2) Almost the same current change occurs in MR_1, and since MR_1 has a finite dynamic resistance its terminal voltage changes.

(3) The base current of the transistor VT_1 changes and can produce a voltage error in the feedback network R_3 and R_4.

The second error can be reduced by suitable selection of type for MR_1 and by lowering its dynamic resistance with constant current fed from the output. The third error can be reduced by lowering the resistance of the R_3, R_4 combination.

The first error is, however, not easy to deal with, simply because the natural tendency is to design with a minimum current in VT_1 at minimum operating voltage. The error is intrinsic in the transistor and is produced

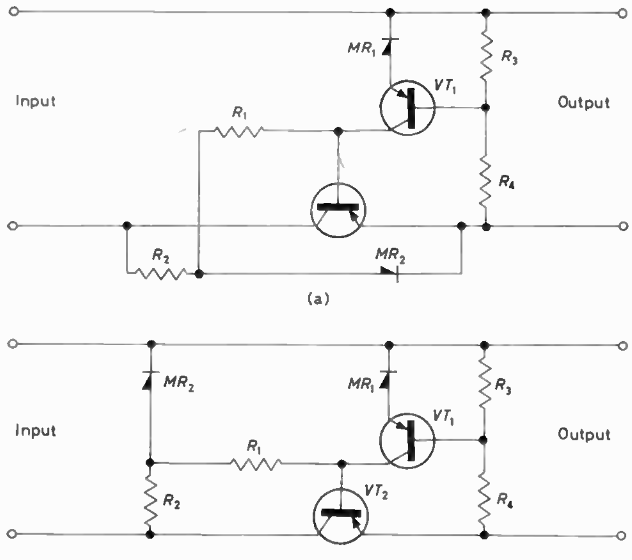


Fig. 2. Arrangements for returning collector load of VT_1 to stable point

largely by the current change in the emitter resistance of the transistor ($r_e \equiv 25/I_e$). In general VT_1 is a small signal transistor and when run at low current the emitter resistance will be high.

Possible Circuit Improvements

Returning to the basic cause of error, namely a wide range of input voltage and the fact that the collector load of VT_1 is returned to the input voltage point, if the collector load is returned to some stable voltage the error will be eliminated[1]. Such arrangements are shown in Fig. 2 (a) and (b). If one has to deal with a situation where the input voltage lower limit is say only one volt greater than the output voltage, stabilization of this type becomes impossible. The reason is obvious, the current in R_1 produces the voltage difference between the stable point and the base of VT_2 and this current must be greater than the base current of VT_2. One must also bear in mind the 'loop gain' concept which demands a high value for R_1.

* *Bell & Howell Ltd.*

Compounding VT_3 reduces the base current requirement in R_1 but the VT_1 component has a finite minimum and again one accentuates the emitter resistance error.

Compounding VT_1 does nothing to improve the situation because the base-emitter voltage change of the combination is at least as great as the single transistor.

Replacing VT_1 by a long tailed pair (a normal process to reduce thermal drifts) worsens the situation by doubling the error.

These attempts to improve the situation are limited to the addition of one semiconductor component.

The Improved Stabilizer

Accepting that as a result of having a wide input voltage change one must have a wide current change in R_1, the philosophy must be to prevent this large change getting to the input transistor. In other words an intermediate stage of current gain is required, but this stage must not introduce a phase reversal. A common collector is a natural choice but the circuit configuration prohibits a pnp transistor and one is forced to consider an npn device. The resultant circuit is shown in Fig. 3.

In this circuit due to the presence of VT_3 the current change occurring in VT_1 collector circuit is small and

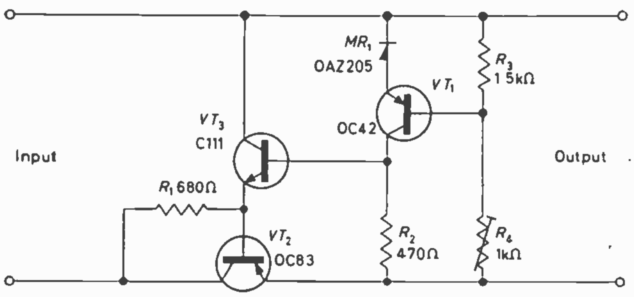


Fig. 3. Improved series stabilizer

consists of two components:

(a) The base current change of VT_3.

(b) The change due to variations in the difference of the base-emitter voltage of VT_2 and VT_3.

It is of advantage to keep the VT_1-MR_1 quiescent current high and still have a large value for R_2, in order to maximize the loop gain. In other words as large a voltage across R_2 as possible is required. Unfortunately the basic circuit limits the voltage drop to the difference between base-emitter voltages of VT_2 and VT_3. For this reason it is essential that VT_2 should be germanium and VT_3 silicon.

The inclusion of a silicon diode in VT_3 emitter circuit increases this difference but unfortunately the diode adds in an error when the VT_3 emitter current changes.

The Practical Circuit

The circuit to be described is intended for exciting a strain-gauge transducer from a vehicle borne battery. The transducer requires a stable 10V at 32mA and the battery varies between 11 and 15V.

An OC83 was chosen as the series element, its voltage and current ratings being adequate and a power dissipation of 180mW is acceptable up to 40°C ambient. In passing one must consider the use of a 'bleed by' or bypass resistor to reduce the series element dissipation but in this case the use of such a resistor is liable to degenerate the stabilizer performance and so the OC83 will require a heat sink if it has to operate in higher ambients.

Allowing some 6mA for the stabilizer circuit the OC83 emitter current is 38mA and assuming a minimum current gain of 50 the base current is about 0·7mA, the emitter base voltage being about 0·3V.

Under a minimum supply voltage condition (11V) R_1 has to be chosen such as to allow for VT_2 base current and a minimum current for VT_3 and in this case a value of 680Ω is acceptable.

Assuming the load current to be substantially constant and considering maximum supply voltage, the current in R_1 rises from 1mA to 6·9mA and the current change in VT_3 is from 0·3mA to 6·2mA.

Fortunately as this current increases in VT_3 its current gain increases and so its base current changes from about 7μA to about 80μA.

The stabilizer reference Zener diode is selected for a highest voltage/low dynamic resistance/low temperature coefficient compromise. A high reference voltage keeps loop gain loss in the feedback network to a minimum. The OAZ205 having a nominal voltage of 7·5V, dynamic resistance of 8·6Ω, and a temperature coefficient of 4mV/°C all at 1mA was chosen. The input transistor operates at low power and the only significant factors affecting its choice are that it should have a high f_1 and its cost. The f_1 should be substantially higher than the series element cut-off frequency in order to avoid a loop instability problem.

The OC42 was chosen for VT_1, and for VT_3 a C111 (npn silicon) was selected. The added advantage of the C111 is that it is of planar construction and so the minimum current problem is eased.

The OC42 and OAZ205 are run at 1mA by a suitable choice for R_2 (470Ω). The base current change of the OC42 will be of the order of 1μA, and the R_3, R_4 resistance combination is chosen for relatively high current in order to prevent significant output voltage change.

Measurements made on stabilizers using this circuit indicate an output voltage change of the order of 10mV (0·1 per cent) for an input voltage change of 11 to 15V. The output resistance is typically 0·2Ω.

A photograph of a completed encapsulated stabilizer is shown on page 330.

Conclusion

This simple circuit has an extremely good performance and its simplicity leads to simple packaging, such as encapsulation, using conventional components. In viewing its simplicity one notes the possibility of a solid circuit version for the following reasons:

(a) With solid circuits it is as cheap to make semiconductor components as it is to make resistors, and in this circuit there are as many semiconductors as resistors.

(b) Capacitors are uneconomic in solid circuits and providing that the f_1 frequencies of the three transistors are adequately staggered instability is unlikely and no capacitors are required.

(c) Resistors R_1 and R_2 are not critical. R_3 and R_4 are only critical as a ratio and in solid circuits ratios are much more easy to control than absolute values.

Acknowledgments

The author wishes to record his thanks to the Directors of Bell & Howell Ltd for permission to publish this article.

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Short News Items

G.E.C. (Telecommunications) Ltd is to supply microwave equipment to the Quebec Telephone Company of Canada for a new u.h.f. broad-band radio relay system operating in the frequency band 1700Mc/s to 2300Mc/s.

The system will provide radio communications over the St. Lawrence river between Riviere Blanche and Baie Comeau using dual spaced diversity reception.

The equipment will be suitable for the transmission of either monochrome or colour television or up to 600 telephone channels. The system will be arranged initially to provide a working (regular) and a standby radio channel for each direction of transmission, but can easily be expanded to provide one standby radio channel for a group of up to three working radio channels.

The equipment will be manufactured at G.E.C.'s Telephone Works, Coventry.

Closed circuit television cameras are being installed on a section of the M6 motorway as part of a four-month experiment by the Home Office Police Research and Planning Branch in unified motorway policing methods. In the first instance, four cameras will be installed and they will relay pictures to the control room of the Knutsford (Cheshire) Service Area Police Post. The equipment is being supplied by the Rank Organization; the cameras are Murphy type MR765 with 6in telephoto lenses and the monitors Murphy type MR763 with 23in 625-line displays.

The Road Research Laboratory (D.S.I.R.) at Harmondsworth, Middlesex, has bought a Redifon 10/20 analogue computer.

It will be used to simulate various road problems such as the behaviour of different types of vehicles on varying road surfaces. The simulation of vehicle suspensions will be studied, as also will be the problem of finding the optimum road profile to provide a smooth transition for all vehicles from one road section to another.

Among the other subjects for which the computer will be used are investigation into heated roads where the automatic heating is switched on when icy conditions are threatening: problems of heat flow and depth of heating grids in such roads and the effect of thermal shock when iced-up roads are salted.

The future programme includes the feeding of road profile constants to the computer to discover a road profile which will give the best riding properties, and also the study of road stresses in relation to wheel loading.

There is also the optimum shape of a snow-plough blade to be investigated and the computer will be used to take a closer look at the behaviour and flight of a snow particle from the blade.

A laser-ophthalmoscope has been developed by surgeons working at the Royal Victoria Infirmary, Newcastle upon Tyne, and scientists of the International Research and Development Co. Ltd. (IRD). This consists of a specially designed miniature ruby laser built into the handle of an ophthalmoscope.

The International Research and Development Co. Ltd designed the miniature laser following the treatment of animals with an experimental laser, also supplied by IRD, at the Royal Victoria Infirmary. Prototype instruments for the treatment of human conditions have been constructed and are undergoing clinical trials. It is understood that this is the first time that this has been done in the U.K. or Europe.

The principal use of this instrument consists in the prophylaxis or normal treatment of retinal detachments and it will also find applications in the treatment of various vascular lesions of the fundus and tumours of the eye. It has several major advantages over instruments at present available for photocoagulation. It enables the normal examination of the eye, treatment and observatión to be carried out with one instrument. No pain or discomfort is caused by the application of the laser beam, because this takes place in one-thousandth of a second, and the patient has no time to react to the bright light and move his eye. The instrument is small, portable and can be operated from any mains supply.

A new electronic sorting system capable of handling over 200 000 letters an hour is being developed by Telephone and Electrical Industries Pty. Ltd, an Australian telecommunications company operating within the Plessey Group.

The equipment has been designed and is being built by TEI at Meadowbank, New South Wales, one of the factories operated in Australia by Plessey Overseas Ltd. The system incorporates a magnetic memory drum store and is claimed to be one of the most advanced of its kind in the world. It will be installed in the Australian General Post Office's new Mail Exchange at Redfern, near Sydney, which is at present under construction.

S.E. Laboratories (Engineering) Ltd has supplied two consoles containing rack mounting electronic equipment and recorders to assist in the static tests being carried out on the Hawker Siddeley 'Argosy' freighter aircraft to measure various phenomena. The total cost of the two cabinets is over £6 000.

One console contains a multi-channel ultra violet recorder type SE.2800 together with 12 channels of carrier amplifiers and associated double integrator networks. The double integrator networks allow measurements to be made from accelerometers and other seismic pick-off devices without need for a reference surface. They will be used to measure displacement of wings under vibratory conditions, and of other control surfaces when subjected to controlled shock loadings.

The second console contains a further SE.2800 recorder and demodulator units. The electronic equipment is designed for operation with inductive transducers for the measurement of pressures, displacement acceleration etc. These again will be used under simulated test conditions on the ground for recording engines and hydraulic pressures, valve movements etc., of the equipment fitted to the 'Argosy' freighter.

It would be possible at a later date to add even further electronic units to these consoles for other measurements associated with this aircraft.

The Ministry of Transport has awarded a contract to Decca Radar Ltd for the installation of equipment which will provide an electronic 'bird's eye view' of the area around Vauxhall Cross, a key junction for South West London and a notorious trouble spot.

The system, which will comprise eleven 'trafficometer' points connected to an information room in the Ministry of Transport, Southwark, will come into operation early in the Summer. Information on the speed (in miles per hour) and flow (in vehicles per hour) of traffic passing each point will be continuously available for display on meters and pen recorders.

A 'Colormatic' display map of the area will form an integral part of the system. This map, giving a dynamic presentation of the traffic situation at a glance, shows the speed and flow of traffic at each trafficometer point by means of coloured lights. These lights represent traffic speed according to a pre-determined code e.g. steady green, no traffic; flashing green—speed above 10 mile/h; amber—speed below 10 mile/h; red—congestion.

The British Association for the Advancement of Science will hold its 126th annual meeting at Southampton from 26 August to 2 September. The meeting will be inaugurated by Lord Brain, F.R.S., Consultant Physician, London Hospital, and Maida Vale Hospital for Nervous Diseases, when he delivers his presidental address entitled 'Science and Behaviour' in the Guildhall, Civic Centre, Southampton, on the evening of 26 August.

The BBC has placed another order with The Marconi Co. Ltd for mark IV 4½in image-orthicon television cameras for use in a new studio at the Television Centre which is due to come into operation during 1965. This new order is for six cameras incorporating circuits for both 405 and 625 line operation. The camera channels are engineered so that a single external switch will effect an instant change of line standard without any need for internal changes to the equipment.

A closed circuit television system has been installed in the Mersey Tunnel, by EMI Electronics Ltd, to help speed the flow of traffic.

Some of the cameras command views of the approach roads feeding the four entrances to the tunnel. Others observe conditions at the two underground junctions and in the central section. This initial siting of the cameras will be subject to amendment and additions in the light of experience. A battery of television receivers is being housed in the traffic inspector's office at the Kingsway end of the tunnel, until a permanent traffic control building is constructed.

All cameras used are EMI type 6 minicameras, to facilitate unobstrusive siting. The cameras in outdoor positions are fitted with remotely-controlled zoom lenses and pan and tilt facilities. They are enclosed in all-weather housings equipped with windscreen wipers and thermostatically-controlled heating. So the traffic controller can see long shots or close-up views in all directions, whatever the weather.

Preferred number series for resistors and capacitors, IEC Publication 63, the 2nd edition of which has just been issued, gives a series of preferred values for fixed resistors and capacitors for both ordinary and close tolerances, of those types of resistor and capacitor intended for use in electronic equipment.

This second edition now contains a preferred number series for close tolerance resistors and capacitors.

Copies of IEC Publication 63 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 10s. 2d. each

The Independent Television Authority has placed a contract with EMI Electronics Ltd for a television aerial, mast and feeder system to be installed at a new ITA transmitting station at Sandy Heath, Bedfordshire. This aerial is intended to improve reception of ITV programmes in the area between those served by the London, Lichfield and Mendlesham ITA aerials. ITA and BBC future u.h.f. aerial requirements in this area in Bands IV and V will also be co-sited on this mast.

Sub-contractor for the design, supply and erection of the 750ft high triangular lattice mast is British Insulated Callender's Construction Co. Ltd.

Mullard Equipment Ltd and Research and Control Instruments Ltd have now been formed into one company known as M.E.L. Equipment Ltd.

Both Companies have for some time been operating under the same management. Their combination is expected to lead to increased efficiency in common services. Continuity of supply of the products of both companies, and existing links with customers, will be preserved.

In dealing with products handled formerly by Mullard Equipment Ltd the new company will operate from the existing Mullard Equipment headquarters at Manor Royal, Crawley, Sussex. For products formerly in the range of Research and Control Instruments Ltd it will operate for the time being from Instrument House, 207, Kings Cross Road, London, W.C.1.

Compania Telefonica Nacional de Espana (CTNE) has placed a £2·5M order with Standard Telephones and Cables Ltd for the first ever telephone cable link between the mainland and the Canary Islands.

The new Spanish cable will be capable of carrying 160 high quality two-way telephone circuits. It will come ashore near San Fernando, which is almost 10 miles south of Cadiz. Through circuits to Madrid and beyond will be provided on the CTNE network. In the Canaries the cable landing point is near Santa Cruz de Tenerife.

Besides supplying and laying the complete 750 nautical miles undersea cable with its repeaters (amplifiers) STC and its Spanish associated company, Standard Electrica, S.A., are also providing equipment for a 50 miles microwave telephone link between Santa Cruz and Las Palmas.

Advance Components Ltd has now changed its name to Advance Electronics Ltd.

The reason given for the change is the fact that for many years Advance has been manufacturing complete electronic equipments and not merely components as the previous name implied. It therefore seemed appropriate that the name should be changed to represent the Company's practice.

Pye Telecommunications Ltd of Cambridge has been awarded a contract valued at £½M for the supply of a fixed and mobile radio-telephone network to cover the whole of Denmark. The radio-telephone equipment, which is to be fitted at police headquarters throughout the country and in police cars and on motorcycles, employs fully transistorized receivers and instant heating transmitter valves. Transmissions from mobile vehicles may be picked up at any of a number of reception points and at the Central Headquarters. Automatic equipment selects the best signal at all times.

A new ARQ centre for the automatic error detection and correction in international radio-telegraph circuits has recently been opened at Fleet Building, London and is expected to be fully operational by August this year.

About 66 per cent of International Radio-Telegraph circuits terminating in London are 'protected' from errors arising from interference or fading on the radio path between the terminal stations, and which are automatically detected and corrected by protective equipment known as ARQ.

The equipment includes 20, 4-channel transistor-type ARQ terminals with auxiliaries and a further 30, 4-channel terminals are now being put in. It is necessary to provide equipment to store traffic during poor radio conditions when Telex calls are sent over protected radio links. This permits the Telex customer to send a message although the ARQ radio path may be interrupted temporarily. When the radio path is restored the characters are then transmitted from the store automatically. The storage equipment provided at the Centre uses magnetic drums with transistor control and a maximum storage capacity of 4000 characters per channel, at six channels per drum.

When completed the whole installation will comprise 200 channels with auxiliary equipment, 90 channel stores and associated testing equipment.

The Radio Marine Associated Companies (RAMAC) of Ingersoll House, Kingsway, London W.C.2, through its twenty-two member companies, has established some four hundred expertly manned shore based depots in more than seventy different countries and island ports, of which more than fifty are based in the British Isles.

The object is to provide a round-the-world and round-the-clock repair and maintenance service for the vital radio communications equipment, navigational aids and allied installations carried by ships at sea.

BOOK REVIEWS

Principles of Coding, Filtering and Information Theory

By L. S. Schwartz. 255 pp. Med. 8vo. Cleaver Hume Press. 1963. Price 72s.

THERE is a need for a short book which surveys the closely related fields of coding, filtering and information theory and would serve as the first introduction to these subjects. In this vein, the author covers a wide range of topics including information transmission on discrete and continuous noisy channels, error correcting codes, power spectra and correlation functions, optimum filtering, detection of signals in noise applying results from statistical decision theory, and communication systems with various kinds of feedback. The mathematics which is introduced is heuristically rather than rigourously justified. It is unfortunate, therefore, that the book lacks a unified presentation of the topics covered.

It also seems unfair to state, as the author does in his preface, that readers without an adequate background in the calculus and the theory of probability will be able to follow essentials. This is evidenced by the inclusion of calculations which use Lagrangian multipliers and the mention, at least, of the Weiner-Hopf integral equation. It would seem that a prior knowledge of the elements of probability theory and the calculus are a prerequisite for the book. One might then envision the book as being suitable for second year university students. Even then, the presentation is jumpy and more than once the author is forced to refer forward to results which are to be introduced in later chapters.

The strength of the book lies in its wide coverage of the important topics of information theory. It would be helpful to the practising engineer who, having the prerequisites, wishes an introduction to information theory. Since the references at the end of each chapter are ample, a more detailed picture can be gotten from them if necessary.

The best chapter of the book is the one dealing with Professor Schwartz' special field of feedback communication systems. While here the presentation is unified, the many types of feedback which are introduced are sometimes inadequately explained. The three appendices which derive the performance relations for these systems should clarify the concepts for the reader.

S. M. Harris.

Informationstheorie (Information Theory)

By Dr.-Ing. Peter Fey. 214 pp. Royal 8vo. Akademie-Verlag Berlin. 1963. Price DM 27.

THIS is the third volume of a series on electronic computing and control. The first two volumes, dealing with the theory and practice of the processing of digital and analogue information in computers, are augmented by the theory of the transmission of discrete and continuous information. In contrast to the determining methods, information is treated according to Shannon as a statistical category during processing. German literature has only recently included methods of the probability theory into the analysis of data processing and transmission. This book deals with the transmission of information via channels subject to interference as a whole, commencing with a statistical description of the source of information and continuing with the problems arising in connexion with its error-free transmission by suitable coding.

Numerous examples facilitate the understanding of the statistical treatment and lead up to its practical application.

E. R. Friedlaender.

Principles of Radio and Electronics

By E. H. Jones. 342 pp. Demy 8vo. Cleaver Hume Press Ltd. 1963. Price 45s.

THIS book is intended to cover the requirements of students taking Part A of the graduate examination of the Institution of Electronic and Radio Engineers. The book assumes no previous knowledge of electricity and the first two chapters, representing about a quarter of the text, are concerned with electricity and magnetism and with a.c. theory. Although these chapters are well written, they are inevitably somewhat abbreviated and a student new to the subject is likely to find them difficult to understand.

The next two chapters deal with network theorems, oscillatory circuits and transmission lines. The development follows a logical pattern and the application of theory is illustrated by worked examples. Further examples, mainly drawn from the examinations of the British Institution of Radio Engineers, are given at the end of each chapter. These examples are rather few in number; for instance only four are given with the chapter on oscillatory circuits. After a short chapter on components, and a brief description of thermionic devices, the various types of valve amplifiers are considered. This is followed by an introduction to transistors; the physics of the device, its parameters and the circuit applications are all dealt with in the space of twenty-five pages. The remainder of the book covers valve oscillators, power supplies, radio transmitters and radio receivers. The final chapter deals with electrical machines.

This book could prove useful to a good student already following a regular course in electronics at a technical college. However, the subject matter is too condensed for the volume to be really suitable for individual study.

V. H. Attree.

Electronic Information Display Systems

Edited by J. H. Howard. 309 pp. Med. 8vo. Cleaver Hume Press. 1963. Price 78s.

THIS book has been produced by bringing together a number of papers by specialist authors—a procedure which is being increasingly adopted in the United States for technological subjects of this kind. There are dangers in such an approach, particulary when, as in this case, the material is largely "adapted from oral presentations." *

Clearly lack of uniformity tends to develop, and in this instance it occurs in quality, treatment and style, virtually with each change of writer, i.e. for each of the 20 chapters. This effect is made worse by the way in which many of the authors have left in so much of their verbal material when adapting it for the book (or give the impression of having done so). Thus in some of the chapters the final result is so oblique that the technical content is obscured. Even in one of the best chapters in the book—by Harvey G. Talmadge, Jr. of the U.S. Naval Research Laboratory, on Classification of Displays—the writing has been weakened by the evident attempt to conform to the chatty style.

The uneven character of the book is further accentuated by its overall structure. No attempt seems to have been made to achieve uniformity in, for example, the presentation of illustrations. Many of them carry no titles—in one chapter there are 21 untitled figures, most of these, however, bearing the acknowledgement (Courtesy—manufacturer X). Furthermore the unbalanced nature of the book is shown in the spread in the length of the chapters which vary from four to a maximum of some 27 pages. At the same time only four chapters include a list of references, the greatest number being 16 for that by Talmadge.

In the light of these comments it may seem inconsistent to state that 'Electronic Information Display Systems' should be read by all those interested in the visual presentation of data and intelligence generally. It contains much that is new, at least on this side of the Atlantic,

* *Made at the Center for Technology and Administration of The American University, Washington, D.C.*

certainly in this consolidated form. Notable in this main context are 'Alphanumeric Symbology Generation for TV Scan—Converted Displays'—Chapter 11 and 'Applications of Display Systems at Goddard Space Flight Center'—Chapter 13.

The cricitisms given above have been assembled in order to explain the rider that reading in this case is an extremely difficult task, especially if one is attempting to link up and correlate the material. Also in view of the declaration in the Preface that "The book . . . is designed to help the manager," it is surprising to find all these obstacles to quick reading —considered to be part of managerial life.

R. E. YOUNG.

Optimum Design of Digital Control Systems, Vol. 10

By J. T. Tou. 186 pp. Med. 8vo. Academic Press. 1963. Price $7

FOR several years now the digital computer, in both large and small forms, has been used in data processing, solution of mathematical problems, simulation studies and many similar applications. For a good many years longer the analogue computer, in some form or another, has been used in system control engineering, a job which it has performed extremely well.

Over the past few years, however, the demands being made on the conventional analogue control systems have shown that they have reached the limit of complexity and accuracy particularly in fields where non-linear control laws are predominant. Over the same period of time the factors which had previously told against the digital computer, very large physical size, slow speed, the restriction to minimal input/output facilities and low order of reliability have been gradually overcome and a new generation of digital computer is available ideally suited to on-line applications.

Mr. Tou's book concerns such application. He states that in control system engineering the most common parameters are bandwidth, gain, rise-time and phase shift, in essence functions continuous in time. A digital computer, of course, is inherently discrete in nature, thus one of the main problems to be solved in digital control is the handling of these continuous parameters by a quantized process.

The author's approach to the problem is by way of dynamic programming and optimization techniques. The first chapter discusses the overall nature of the digital control system, the second discusses dynamic plant characteristics. Chapter three defines the optimum control problem and the relevant laws. Thereafter up to chapter nine, the last, each aspect is considered in greater detail. The author concludes with a set of examples to be worked.

Reference is made throughout the book to the techniques of dynamic programming and large sections of each chapter are entirely mathematical making reference to advanced vector treatment, matrix algebra and algorithms. As such it is a book to be studied by the mathematician who is entering this particular field although anyone concerned with on-line digital computer application could read the book to advantage.

T. R. H. SIZER.

Impulsschaltungen (Pulse Circuits)

By A. P. Speiser. 288 pp. Med. 8vo. Springer Verlag. 1963. Price DM 36.

THE sub-title of this excellent book is "The Production and Handling of Pulses by Means of Valves and Transistors". The treatment of the subject matter is a balanced combination of essential theory with down-to-earth practical considerations as one might expect from an author like Dr. Speiser who is both a teacher and the head of a commercial development laboratory. Theory and practice go hand in hand throughout the book and although the early chapters are devoted to basic theory, even they are presented in an essentially practical way giving useful approximations and rules of thumb for immediate application. The scope of the book, to provide basic knowledge and design information to students and practising engineers about pulse techniques, mainly with pulses of rise times under 10nsec is defined in the first chapter and rigidly adhered to in spite of the fact that about 20 pages at the end of the book are devoted to nano-second pulses. The organization of the book is logical and disciplined and it is surprising that the author should have had doubts, somewhat self-consciously expressed in chapter 2, on this point. The 3 pages of this chapter could easily be missed, even the synoptic table is not of much use. Another excellent feature is the careful comparisons based on technical and economic considerations made between the relative merits of valves and transistors for specific applications. In short, although nothing basically new is revealed, the book contains all that even an advanced worker in the field will require. A book of this type might be welcomed in the English language, but would a mere translation do justice to the easy, readable style?

K. L. SELIG.

USSR Direct Current Research (Reports of the Leningrad Direct Current Research Institute)

Translated from Russian by L. A. Fenn. 287 pp. Med. 8vo. Pergamon Press. 1964. Price £5

This book presents the results of recent research in direct current study, and includes details of research into the latest techniques in measuring gas pressure in apparatus containing mercury, gas-separation control informing a high-voltage valve, and the effects of fouling on the insulators of h.v.d.c. overhead lines. Other subjects include the future prospects of the high voltage direct current transmission of electric power in the U.S.S.R. and an analysis of the stability of the direct current transmission line from Kashira to Moscow.

Optical Processing of Information

Edited by D. K. Pollack, C. J. Koester and J. T. Tippett. 286 pp. Med. 8vo. Cleaver Hume Press. 1963. Price 60s.

This volume comprises the Proceedings of the Symposium on Optical Processing of Information, co-sponsored by the Information Systems Branch of the Office of Naval Research and the American Optical Company, held on October 23-24, 1962, in Washington, D.C. The purposes of the symposium were to bring together research workers in the fields of optics and information processing.

Telemechanics (International Series of Monographs on Automations and Automatic Control)

By V. S. Malov. Translated by F. Immirzi. 100 pp. Demy 8vo. Pergamon Press. 1964. Price 35s.

An attempt has been made in this small book to give a short description of the problems of telemechanics and of the principles and design of such systems.

Optical Masers

By O. S. Heavens. 103 pp. Crown 8vo. Methuen & Co. Ltd. 1964. Price 16s.

A description of optical masers is given at a level suitable for those equipped with the background of an ordinary university course in physics.

An Introduction to Industrial Radiology

By J. C. Rockley. 237 pp. Med. 8vo. Butterworth & Co. Ltd. 1964. Price 55s.

This book will be of use to inspection engineers who are concerned with the application and evaluation of radiological techniques and the interpretation of radiographic evidence, and to industrial radiologists who are preparing for the City and Guilds of London Institute examination in industrial radiology.

Theory of Psychological Measurement

By E. E. Ghiselli. 408 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price 69s. 6d.

This book offers a comprehensive treatment of the problems, statistical techniques, and theoretical concepts basic to psychological testing and the measurement of mental traits. Mathematical models and their use in dealing with problems in psychological measurement are discussed.

After discussing the concepts fundamental to an understanding of measurement consideration is given to norms, correlation, composites of scores, and reliability and validity of measurement.

Automatic Voltage Regulators and Stabilizers

By G. N. Patchett. 468 pp. Demy 8vo. Sir Isaac Pitman. 1964. Price 65s.

The book has been heavily revised to bring it up to date and much new material has been added on tap change equipment. Since the first edition, transistors have become important and a large new section has been added on transistor stabilizers, including Zener diodes.

All types of automatic voltage regulators and stabilizers are dealt with, ranging from those designed for use with a few watts in electronic equipment to those for use in the largest alternators of 100 000 megawatts and more. Both the electromechanical types of regulator and the newer electronic regulators and stabilizers are covered

Electrohydraulic Servomechanisms

By A. C. Morse. 238 pp. Med. 8vo. McGraw-Hill Book Co. 1963. Price 89s.

This book has been written for the systems design engineer and deals with the components, operation and maintenance of electrohydraulic servomechanism systems and their design.

The French Component Exhibition

A description, compiled from information supplied by the manufacturers, of a few of the French exhibits at the International Exhibition of Electronic Components in Paris from 7 to 12 February 1964

(Voir page 345 pour la traduction en français; Deutsche Übersetzung Seite 354)

COGIE

3 et 5, boulevard Anatole-France, Aubervilliers (Seine)

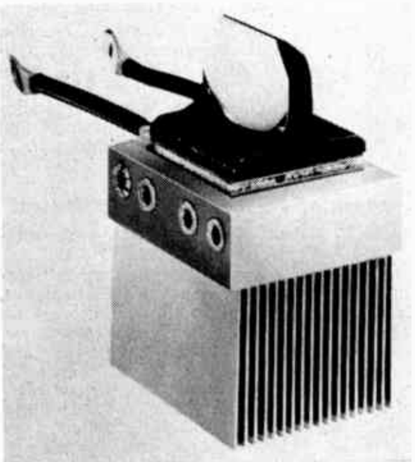
THERMOELECTRIC COOLING MODULES

(*Illustrated below*)

These 'Fricogie' thermoelectric (Peltier effect) modules type FC8-7A consist of eight bismuth-telluride couples assembled with thermal insulation and electrically connected in series. The maximum heat extraction is 32W.

By changing the polarity of the d.c. supply the hot and cold junctions are inverted which provides an excellent means of temperature regulation. The maximum difference between the hot and cold junctions is 75°C with the hot junction at 25°C.

Smaller versions of these modules are also available.



EE 69 751 *for further details*

C.O.P.R.I.M.

(Compagnie des Produits Élementaines pour Industries Modernes)

7, Passage Charles-Dallery, Paris 11e

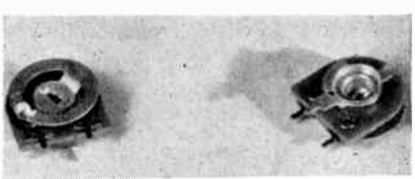
MINIATURE POTENTIOMETERS

(*Illustrated below*)

This new model (E 086) forms part of the range of carbon film potentiometers; it is intended for use in printed circuits, pitch 0·1in.

It is smaller than the model 4 E 097, although of a similar aspect and construction, model E 086 is rated at 0·1W at 40°C.

These components are suitable



wherever space is at a premium and for lower powers.

They are available in a range from 100Ω to 20kΩ, and or up to 500kΩ, upon request.

EE 69 752 *for further details*

ELNO

(Etabl Lailler-Pecquet et Cie)

18 et 20, rue du Val-Notre-Dame, Argenteuil (Seine-et-Oise)

LIGHTWEIGHT HEADSET-MICROPHONE

This headset-microphone combination weighs less than 150g and consists of a steel wire headband, wrapped in plastic sheathed polyester foam, at one end of which is attached a lightweight earphone. The front of the earphone is padded while the microphone boom is attached to the back. The microphone arm permits both vertical and lateral movement. The leads from the earphone and the microphone are taken to a junction box attached to the clothing of the wearer.

The frequency range of the earphone is 20c/s to 4 kc/s ±3dB. It is available in six standard impedances from 15Ω to 2kΩ.

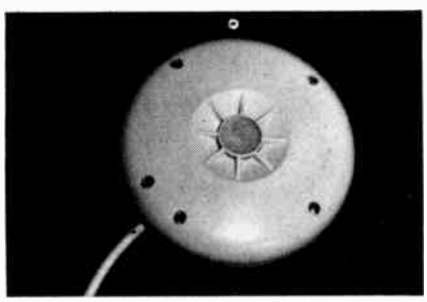
The sensitivity of the microphone is 0·1mV/μB. The frequency response rises at 6dB/octave up to 1·5kc/s and is then flat to 4kc/s. It is available in six impedances from 15Ω to 2kΩ.

EE 69 753 *for further details*

PILLOW LOUDSPEAKER

(*Illustrated below*)

This loudspeaker is intended for persons lying in bed in hospitals etc. It consists of a circular case of ellipsoidal section with a vent on both faces to allow



of sound diffusion. An electromagnetic earphone receiver is mounted inside the case. It is available in ten standard impedances from 15Ω to 18kΩ.

EE 69 754 *for further details*

FÉRISOL

18, avenue Paul-Valliant-Couturier, Trappes (Seine et Oise)

COAXIAL TEST LINES

Test lines type UB 200 and UA 200 are slotted impedance coaxial lines—50Ω for type UB 200 or 100Ω for type UA 200.

They are fitted with an identical mobile carriage comprising a plunger that is adjustable in depth and serves for h.f. voltage pick-up. This is aperiodic and does not include a tuning facility. The h.f. voltage is available on an 'N' plug. The mobile carriage also contains a coaxial crystal detector for modulated or non-modulated h.f. voltage. The v.s.w.r. of the impedance studied can thus be directly measured by connecting type AG 101 v.s.w.r. indicator with the test line.

The mechanical displacement of the carriage on the test line is accurately marked on a scale graduated in centimetres and millimetres and on a vernier graduated in tenths of a millimetre. The resulting precision is of the order of 0·1mm. Greater precision (0·01mm) can be obtained for small displacements (maximum travel 50mm) by making use of a special fixing device for a mechanical comparator.

EE 69 755 *for further details*

ELECTRONIC MEGOHMMETER

Megohmmeter type RM 200 is designed for measuring resistance from 0·2MΩ to 10^8MΩ.

The unknown resistance is directly displayed on a galvanometer dial by a new technique for the measurement of very low d.c. voltages. In contrast to conventional devices of the Wheatstone bridge type where the operator has to restore the bridge balance with every variation of the unknown, the new type RM 200 displays this value without any external intervention. This makes it possible not only to measure the instantaneous value of the unknown but also to make a study of its behaviour on exposure to other parameters such as temperature, humidity, etc. For this purpose, a 'recorder' output supplies a d.c. component whose value is linked through a simple function to the value of the unknown.

The numerous test voltages supplied by the unit itself make it a universal instrument suitable for the most modern techniques. For example, the 10V test voltage permits the study of transistor components.

In addition, the unit is so designed as to permit the use of a test voltage of any value from an outside source. A simple calculation establishes the true value of the unknown, taking into account the value shown on the test galvanometer dial and the test voltage applied.

EE 69 756 *for further details*

FERRANTI LTD

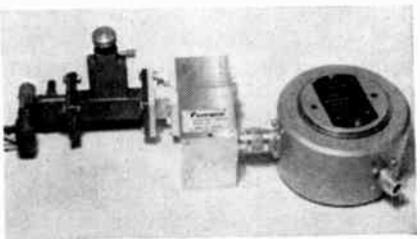
Hollinwood, Lancashire

Parametric Amplifiers

(Illustrated below)

The 570Mc/s parametric amplifier (type VCA/L10) achieves a 16dB gain at a 6Mc/s bandwidth (relative to the 1dB points) and a total noise figure of less than 1dB. It is interesting to note that this noise figure is much less than the received sky noise at this frequency.

The frequency 570Mc/s is the transmission frequency of the new British Broadcasting Corporation's 625 line television programme, and the new amplifier has been primarily developed for purposes associated with this transmission. An order for these amplifiers has been received from Rediffusion Research Ltd who will use them to enable their television subscribers to receive the new programme on a wired circuit at Southampton.



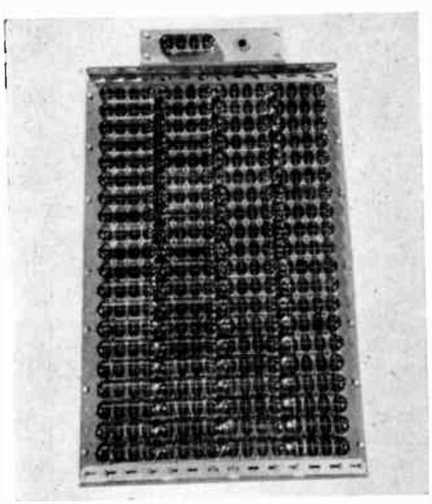
Another parametric amplifier on view was the type VCA/S30. This amplifier is designed for operation at a centre frequency of 3 000Mc/s and it achieves a 20dB gain at a bandwidth of 70Mc/s (relative to the 3dB points), and a noise figure of 2·7dB. Main applications of parametric amplifiers designed for operation at S-band frequencies, are in radar receivers where they increase the effective radar range by a minimum factor of three.

EE 69 757 *for further details*

Photovoltaic Cells

(Illustrated below)

A new large-area silicon photovoltaic cell unit known as type MS40 was exhibited which is primarily intended as a very efficient and economic device for converting solar energy into electric current suitable for charging battery buffer storage systems. The MS40 is composed of four individual cells parallel connected to a common heat sink. An integral moulded cylindrical lens encapsulation increases the effective solar radiation collection area and this, together with other design features, results in a solar spectrum conversion efficiency of about 8 per cent at ground level.



With a tungsten source at 2854°K (normal incidence) and an illumination density of 3 000 lm/ft², the new cell will deliver (at 25°C ambient) a minimum current of 500mA into a 0·6Ω load. A great deal of design attention has been given to the severe corrosion and mechanical environmental stresses to which solar panels constructed from MS40 modules will be subjected. Mounting and electrical connecting posts are constructed of Monel 400 alloy which has a corrosion rating sufficient for direct marine environments, and a rugged epoxy plastic encapsulation is employed which withstands the operation and storage ambient temperature range of from −20°C to +85°C.

A trial assembly of MS40 modules has been shipped to the Persian Gulf where it is to be tested on fixed maritime structures in order that further data may be obtained to complete the field trials. This final field trial is being made to determine the optimum design of large solar-electric power supplies for use in remote locations.

EE 69 758 *for further details*

HEWLETT-PACKARD S.A.

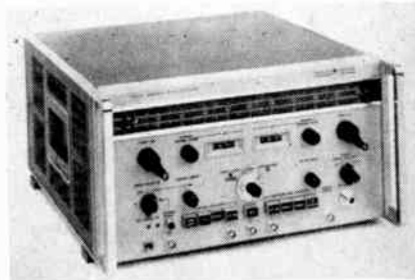
54 Rte des Acasias, Geneva, Switzerland

Microwave Sweep Generators

(Illustrated above right)

The sweep generators types 691A/B (1 to 2Gc/s) and 692A/B (2 to 4Gc/s) feature three modes of sweep operation. In the first mode called 'normal' the output frequency is swept in either direction between two continuously adjustable limits. In the 'F' mode, frequency is swept about an adjustable centre frequency. In addition these sweepers have two independently adjustable frequency markers which may be digitally adjusted with a resolution of ±0·05 per cent of the output frequency. These markers amplitude modulate the output in both the 'normal' and 'F' modes—a very important feature when making x-y-recordings or observing the sweep on an oscilloscope. As a third mode of sweep operation, these same frequency markers may be used to establish a sweep mode similar to 'normal' called 'marker sweep' in which the markers determine the sweep range. Push-button mode selection makes it possible (for example) to first sweep the entire pass-band of a filter in 'normal' and sweep the rejection region of the same filter by switching to 'marker sweep' without readjusting sweep limits.

In the 691B and 692B diode strip-line sections are used for modulation



and attenuation. The amplitude modulation which these pin diodes perform is completely independent of the b.w.o. tube so frequency pulling is virtually eliminated. These pin diode networks are also utilized to produce an output signal levelling of 0·3dB.

EE 69 759 *for further details*

Standard Frequency Receiver

(Illustrated above)

The receiver type 5090A is intended for reception of the 200kc/s transmissions from Droitwich. Since the stability of the Droitwich carrier frequency is within 5×10^{-10} per day this provides a simple and inexpensive method of obtaining a highly accurate frequency standard for laboratory use.

The 5090A has output frequencies of 100kc/s and 1Mc/s.

EE 69 760 *for further details*

I.E.R.

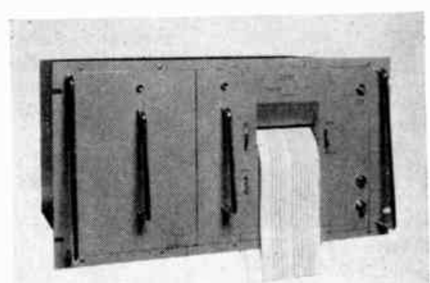
(Impression Enregistrement des Résultats)

6, rue Blondel, Courbevoie (Seine)

Strip Printer

(Illustrated below)

The high speed parallel input digital



strip printer type 205 can accept an input in binary coded decimal and decode and print at a maximum rate of ten lines per second. The maximum width is 13 columns and thus the maximum speed is 130 characters/second.

It was developed from the type 203 which works at a speed of six lines per second.

EE 69 761 *for further details*

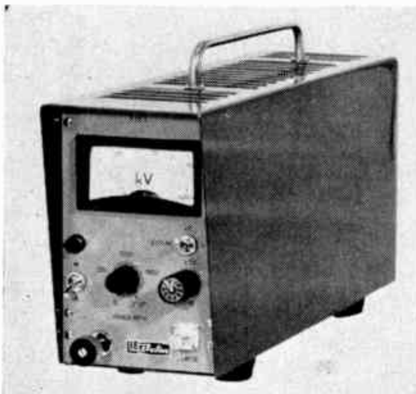
LIE-BELIN

296, avenue Napoléon-Bonaparte, Roeil-Malmaison, (Seine-et-Oise)

E.H.T. Power Supply

(Illustrated below)

This fully transistorized e.h.t. supply unit is available as a standard Esone 2 M component or an enclosed version with carrying handle.



Its main characteristics are as follows:

Output: 2kV, 2mA.

Stability: 5×10^{-5} for ± 10 per cent mains voltage variation.

Internal resistance: 50kΩ.

Voltage control: —by 80V steps
—fine by means of 10-turn potentiometer.

Control: by voltmeter on front panel.

Supply: 127 or 220V 50c/s mains supply or 24V d.c.

EE 69 762 *for further details*

Radioactive Filter Analyser

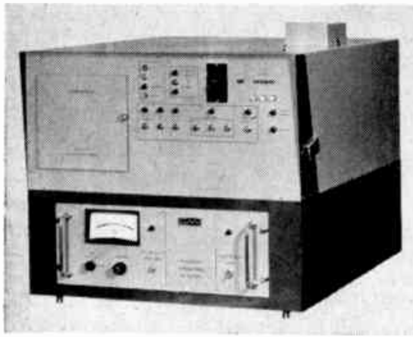
(Illustrated above right)

This instrument is designed for automatic or manual measurement of the α-β-activity of dust-sampling filters.

The filter to be tested is brought before a photo-multiplier followed by an electronic counting unit complete with display and printing facilities.

For α-detection, a separate scintillator (ZnS) is used for each filter and placed against it in the filter-holder. For β-detection, use is made of a single scintillator which is integral with the photomultiplier.

The unit consists of a case made up of two sections: The upper section houses the detector (54 AVP) with preamplifier amplifier and shaper and, where necessary, a lead turret (β-analysis); filter-holder magazine, magazine feed, devices for the selection of filters



from the magazine and for their feed and return, and for printing the results on each filter-holder and on a continuous paper strip; electromechanical parts with programmer, relaying and safety devices.

The lower section comprises a standard 4U3 chassis housing the electronic and supply systems with; 5 counting decades with display and printing by mechanical monodecades.

The loader capacity is 60 filters of 160mm × 160mm.

EE 69 763 *for further details*

METRIX

(Compagnie Générale de Métrologie)

Chemin de la Croix-Rouge, Annecy, (Haute-Savoie)

Transistor Meter

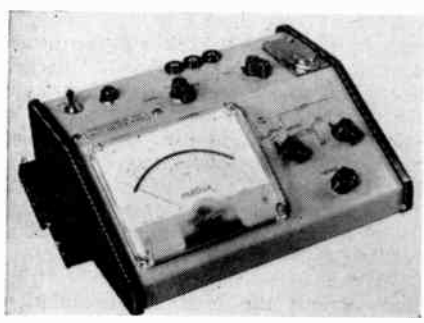
(Illustrated below)

The transistor meter type 302A has been developed for the testing of diode detectors and rectifiers, Zener diodes and transistors ranging from those of a few milliwatts to power transistors.

Diodes are tested by checking the inverse current which is measured by means of a high-sensitivity galvanometer. A series resistor in the circuit protects the galvanometer against a faulty connexion. Throughout the test, the diode is subjected to a voltage not exceeding 4V.

When testing Zener diodes the diode is series-connected with an 18V source and a current-limiting resistor. The meter, whose power consumption is negligible, is connected to the diode terminals and measures the voltage regulated by the diode. The power dissipated in the diode is extremely low (approximately 30mW).

In the case of transistors, measurement of I_{CBO} is carried out in the same way as measurement of the diode inverse current.



For the measurement of static gain h_{21E} or h_{FE} the circuits of the instrument are designed so as to enable the collector current to be adjusted to 1, 10, 100mA or 1A.

When this operation is carried out, the collector-emitter voltage is 2V. Power dissipation in the transistor is then 2mW, 20mW, 200mW or 2W, depending on the collector current selected. Power is at a maximum at the time of measurement. There is thus no danger of overload while the collector current is being calibrated.

Once this preliminary operation has been completed, all that is necessary is to measure the base current so as to obtain the static gain as expressed by the ratio $\frac{I_C}{I_B}$. The galvanometer used for measuring I_B is graduated directly in terms of gain.

EE 69 764 *for further details*

ORÉGA

(Societe Electronique et Mecanique)

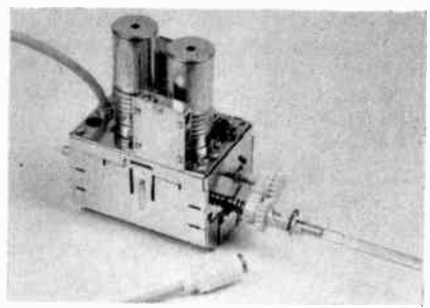
106, rue de la Jarry, Vincennes (Seine)

Television Tuner

(Illustrated below)

Known as the 'Rotomatic' this is a v.h.f. tuner-amplifier of very small size for use in portable television receivers and is available for either mechanical or electrical tuning.

The valve version is claimed to be the smallest on the market with the technical specification laid down for the 819 line system. It has a high gain and a



low noise figure of 5 to 7dB. It employs valve types ECC189 and ECF801.

A transistor version is also available which provides high gain and an even lower noise figure. The power consumption is 17mA at 12V.

Both types are available for either French or foreign standards.

EE 69 765 *for further details*

QUENTIN ET CIE

2, rue Hoche, Ermont (Seine-et-Oise)

Transistorized Power Supplies

(Illustrated on page 339)

These stabilized power supply units are intended for use from a 110 to 127V or 220 to 240V, 48 to 65c/s supply. They provide a d.c. output of 0 to 150V: three models are available known as the AS280, AS281 and AS282 which have output current ratings of 0 to 1A, 0 to 2A and 0 to 5A respectively.

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The output voltage regulation is better than 0·05 per cent for ±15 per cent change in input voltage. The load regulation is better than 0·05 per cent change in output voltage from no load to full load.

The stability is such that the output varies less than 0·1 per cent over a period of 8 hours after a 30min warm-up period. The output voltage varies less than 0·05 per cent/°C for ambient temperature changes in the range −10°C to +50°C. The output ripple is less than 3mV peak-to-peak.

A current limiting circuit is provided which is adjustable between 20 per cent and 110 per cent of rated output.

EE 69 766 *for further details*

RIBET-DESJARDINS

13-17, rue Périer, Montrouge (Seine)

Transistorized Oscilloscope

The oscilloscope type 349A is fully transistorized and combines the advantages of small dimensions with the properties of a reliable instrument for high performance measurement. It can be operated from the mains or from a rechargeable battery.

The vertical deflexion amplifier has a pass-band of 0 to 1·5Mc/s at a sensitivity of 50mV/div and 8c/s to 1·5Mc/s at 10mV/div. It has a calibrated attenuator with steps of 10mV/div to 20V/div.

Synchronization can be automatic internal or external. Minimum voltage required is 10V peak-to-peak and the input is protected against overvoltages of up to 300V.

A built-in calibrator provides square waves of 1V peak-to-peak.

EE 69 767 *for further details*

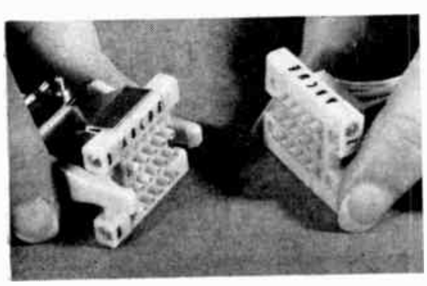
SOCAPEX

9, rue Edouard-Nieuport, Suresnes (Seine)

Multi-Pin Connectors

(Illustrated below)

The connectors of series 67—male and female stackable 6-contact (3 male + 3 female contacts) connectors for high operating voltages—have been considerably improved in the course of the year.



The capped connectors now have one of their end fittings and the corresponding locking catch in different colours in order to prevent errors.

The marking system employed enables any contact to be immediately identified, whether it is on a plug or base.

EE 69 768 *for further details*

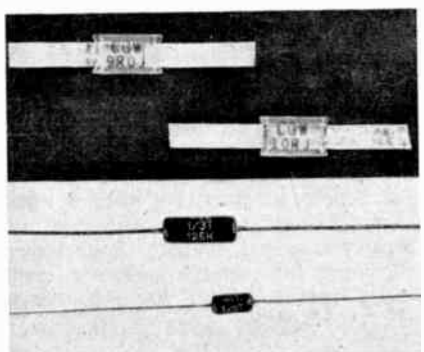
SOVIREL

27, rue de la Michadière, Paris 2e

Miniature Resistors and Capacitors

(Illustrated below)

The ranges of metal oxide film resistors in the high and very high stability ranges (from 1/8Ω to 6kΩ) and glass dielectric capacitors, ranging from 0·5pF to 0·01μF with a loss angle of less than 10×10^{-4}, provide an answer to the quality and space problems presented by modern circuits.



Resistors in types CR, A and HR are available having a proven fault rate of only 0·00069 per cent/1 000h (7 000 resistors tested over a period of 2·000h at 2½ times rated dissipation).

Miniature glass dielectric capacitors from 0·1 to 30pF specially designed for tuners are also available.

EE 69 769 *for further details*

TACUSSEL SOLEA

2 et 4, rue Carry, Lyon, 3e (Rhone)

Electronic Multi-Meter

(Illustrated below)

The electronic multi-meter type VE7 is characterized by a very high input impedance when used as a voltmeter, the instrument includes an amplifier

fitted with an electrometer valve and ten transistors (mostly silicon). It is powered by internal batteries.

The principal measurements it permits are:

- Voltages from 1mV to 1kV d.c. 1 per cent f.s.d. accuracy, infinite input impedance up to 3V;
- Currents from 1pA (10^{-12}A) to 1mA d.c., 1 per cent f.s.d. accuracy;
- Resistances from 2 to 1 000 000MΩ, 2 to 3 per cent accuracy.

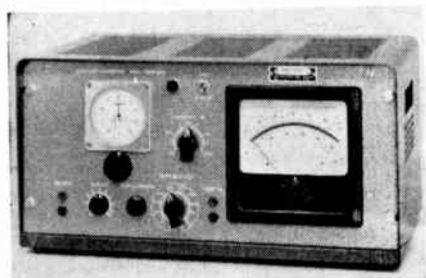
Several accessories are available for the following measurements integration of quantities of electricity; capacitances; small voltage fluctuations; conductances; pH values.

EE 69 770 *for further details*

Distortion Meter

(Illustrated below)

The harmonic distortion meter type DH 2 is a broad band instrument designed to measure harmonic distortion values between 0·05 and 100 per cent,



in signals of which fundamental frequencies may range from 3c/s to 300kc/s. It includes active electronic (Wien bridge) filter and a broad band a.c. millivoltmeter. It has a low impedance output for oscilloscope examination of residual waveforms.

This instrument is also intended to be used as an a.c. millivoltmeter for signals from 0·5mV to 50V (3c/s to 1Mc/s).

EE 69 771 *for further details*

TELEFUNKEN AG

Ernst-Reuter-Platz 7, 1 Berlin 10, Germany

U.H.F. Television Tuner

This tuner is designed for the television band 470 to 860Mc/s. Its robust flat case is divided into four almost equally large spaces by means of intermediate walls, which act as tank circuits in each case and which may be tuned to the desired frequency by means of a triple rotary capacitor assembly. The tuner type 132, designed for manual tuning, contains valves type EC88 and EC86. The u.h.f. tuner type 140 is housed in the same case as type 132. It is fitted with two transistors AF139, the first operating in the input stage and the second in the self-excited mixer stage. Both tuners are designed in half-wave technique. However, as from the middle of 1964 a quarter-wave tuner will be supplied for the French market which will likewise be fitted with two transistors AF139.

EE 69 772 *for further details*

ELECTRONIC EQUIPMENT

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

(Voir page 348 pour la traduction en français; Deutsche Übersetzung Seite 357)

MAGNETIC AMPLIFIER

Distributed by: Auto-Electronics Ltd, Peel Grove, London, E.2

(Illustrated below)

Manufactured by Serelec, of Paris, the type AM-TH-PP is a versatile high performance push-pull magnetic amplifier for the measurement of low level d.c. signals from thermocouples, strain gauges, various bridges, Hall effect devices, etc. It is also suitable for use as a stable pre-amplifier in closed loop regulation systems.

It features good linearity, stability and ruggedness combined with small dimensions, light weight and low cost. There are no moving parts and if used within their design limits they are not subject to loss of efficiency, and they have a

limitless life. The 'dead zone' (a common shortcoming of this class of amplifier) has been eliminated and signals of 0 to 100μV are amplified as faithfully as those of 0 to 100mV. It is also insensitive to any external magnetic fields.

It has high dynamic input and output impedances and variations in the resistances of the source and of the load have negligible effect on the gain and accuracy of the amplifier. This eliminates problems arising from variation in the resistance of leads or contacts. It is also suitable for use with coaxial thermocouples.

EE 69 773 *for further details*

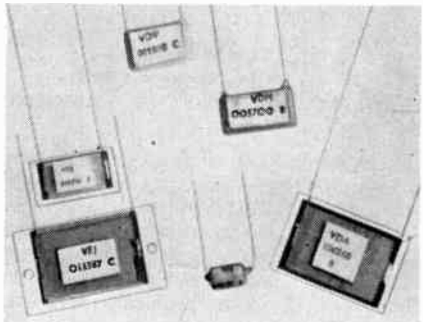
SILVERED MICA CAPACITORS

Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(Illustrated above right)

Two alternative standard versions of silvered mica capacitor are now available from Plessey-UK Ltd. The wax-dipped type utilizes a special non-cracking wax and lacquer coating and is suitable for the temperature range $-25°C$ to $+70°C$. The epoxy resin encapsulated types provide humidity protection to H.2 standards and extend the temperature range from $-40°C$ to $+85°C$.

The capacitors can be supplied in any



value up to 1·0μF at 200V d.c. working; 0·75μF at 350V; 0·5μF at 500V; and 0·05μF at 750V. Four sizes of plates are used in the standard range, from 7/16in × ¼in (11·1mm × 6·4mm) to 1in × ½in (25mm × 13mm) depending on the capacitance. Terminations are tinned copper wire (normally radial, but axial on request) attached in a manner which ensures good electrical connexion and great mechanical strength. Special mica capacitors for use at temperatures up to 250°C are available if required.

EE 69 774 *for further details*

DYNAMIC BALANCING MACHINE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated on right)

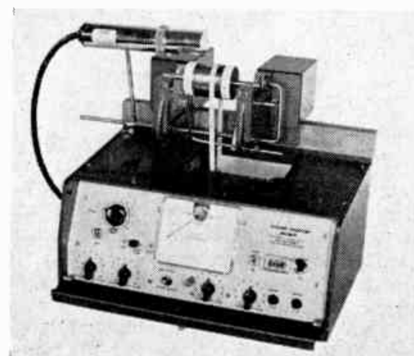
The new type 1250C dynamic balancing machine has been designed to provide a simple yet accurate means for balancing rotating components of all kinds on the production line.

The machine is a compact unit (base 15 × 12½in; height 15in; weight 55 lb) intended for use on a bench. Mounted on a console containing all circuits is a robust rail, on which two open cradles can be adjusted laterally to accept workpieces from 2in to 13in long, weighing from 1oz to 10 lb. The workpiece under test is driven by a light belt from a built-in electric motor, at speeds from 600 to 3 000rev/min.

The cradles incorporate vee-blocks so mounted that they have full freedom in the transverse direction. Vibration due to unbalance in the workpieces under test is transmitted by the vee-blocks to pick-up coils in a strong magnetic field. An alternating voltage is generated in the coils proportional to the amplitude of vibration of the workpiece and is fed, after amplification, to an indicating meter. It is also used to trigger a white-light stroboscope mounted above the workpiece. A numbered tape or collar applied to the workpiece is thus 'frozen' and shows, against an adjustable pointer, where weights must be added or removed (depending on the position of a selector switch) to restore balance.

For dynamic balance, weight adjustment must be made in two transverse planes. A further switch enables the right-hand or left-hand cradle to be selected for independent indication.

The type 1250C machine incorporates a variable filter circuit to enable extraneous vibration to be filtered out.



Each cradle also has associated calibration and sensitivity-adjusting circuits. These are preset for production-line work, so that even an unskilled operator can determine dynamic unbalance in a workpiece in about 20sec after brief instruction.

The white-light high-intensity stroboscope enables the machine to be operated in daylight. The open vee-bearings simplify loading and unloading. The provision of two bearings and a selector switch obviate the need for reversing the workpiece, thus speeding up the operation.

EE 69 775 *for further details*

MULTI-PURPOSE POTENTIOMETER

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustrated on page 341)

This new instrument, which supersedes the Cambridge workshop potentiometer type 44226, is a compact and completely self-contained instrument designed primarily for 'on-site' testing and calibration of thermocouples and associated indicators and recorders. It is also suitable for many routine current and voltage measurements in the laboratory.

The potentiometer has a single range

ELECTRONIC ENGINEERING will occupy stand number N.163 at the I.E.A. Exhibition, Olympia, from 25 to 30 May and visitors will be welcome

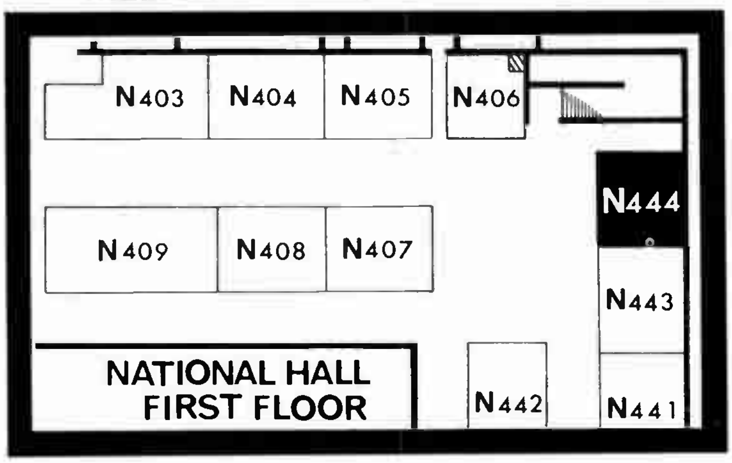
SEE OUR WORKING DISPLAYS ON STAND No. N444 I.E.A. MAY 25th MAY 30th

SEE Silicon Controlled Rectifiers as opposed to Saturable Reactors

SEE Differential Amplifiers on marginal error detection

SEE Trip Amplifiers in typical working conditions

CORREX



MODULAR EQUIPMENT

FOR STEPLESS PROPORTIONAL CONTROL OF VOLTAGE, SPEED, POWER, HEAT, LIGHT, ETC.

This range is readily available as individual units or as packaged systems designed to enable you to construct control schemes using the CORREX modular "add on" system.

Included, for use with and in this range are:-

PROPORTIONAL CONTROLLERS
MAGNETIC INSTRUMENT AMPLIFIERS
MAGNETIC AMPLIFIER DRIVER STAGES
TRIP AMPLIFIERS (MAGNETIC)
SATURABLE REACTORS in a wide range UP TO 180 KVA.
TRANSFORMERS

Other specialised control equipment for all purposes.

PHOENIX TELEPHONES LIMITED
GROVE PARK, LONDON, N.W.9
Telephone COLindale 7243 Telegrams PHONOFENIX LONDON NW9

of 0·01 to 101mV, selected by a multi-point switch and a calibrated slidewire, the accuracy of adjustment being ±0·1 per cent or ½ slidewire division, whichever is the greater. The main potential dial controlling the multi-point switch has 100 steps of 1mV, and the fine potential dial controlling the slidewire has a range of 1mV and is divided at intervals of 0·1mV, allowing estimation to 0·002mV.

A five-position function switch selects one or other of the two test circuits, which are primarily used together for comparing test and standard thermocouples. A third position connects a continuously adjustable millivolt supply to the test terminals for deflecting moving-coil instruments, and a further position allows the instrument to be used as a calibrated potential source for testing potentiometric instruments.

The instrument is completely self-contained with built-in galvanometer, standard cell and dry batteries. External terminals allow a more sensitive galvanometer to be connected without removing the built-in galvanometer, and the range of the instrument can be extended to 500V d.c. and 10A d.c. with the aid of a few simple accessories.

The potentiometer can be standardized at any time during tests, without alteration of the dial or function switch settings, by pressing the standardize key and adjusting a single rheostat.

EE 69 776 *for further details*

PRECISION RESISTORS

Electrothermal Engineering Ltd, 270 Neville Road, London, E.7

(Illustrated below)

Electrothermal Engineering Ltd has extended its series G 101 range of all moulded wire-wound Precistors to include two new tag versions.

These are the type 3505T ¼W and type 3507T ½W which employ the same moulded-in anchoring features used in all series G 101 Precistors. This ensures robust construction, with a high reliability factor as well as 100 per cent positive humidity sealing.

The values available are from 0·1Ω to 1MΩ in the ¼W type and 0·1Ω to 2MΩ in the ½W type. Tolerances are from 0·1 per cent to 0·01 per cent dependent on value, while the stability is better than ±0·02 per cent. Maximum working voltages are 550V and 750V respectively.

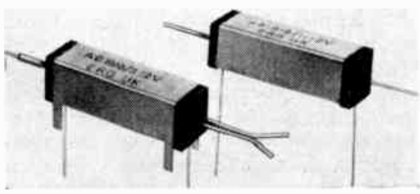
EE 69 777 *for further details*

DRY REED RELAY

Erg Industrial Corporation, Luton Road Works, Dunstable, Bedfordshire

(Illustrated below)

A new compact dry reed relay, known as style AB.1618, specifically designed to meet the requirements of the industrial process control, computer and allied equipment fields, has recently been introduced by Erg Industrial Corporation.



Main features of this new relay include a fast switching time—down to one millisecond—coupled with a working life of many millions of operations.

The gold-plated contacts are hermetically sealed in an inert atmosphere, and are available either as single change-over (s.p.d.t.) or one normally open contact (s.p.s.t.). The mounting of the dry reed switch enables them to be interchanged and magnetic screening of the relay minimizes effects of extraneous magnetic fields.

The bobbin of the relay is moulded from fibre glass reinforced epoxy resin, and the integrally moulded coil leads are specially designed to suit the latest printed circuit mounting technique.

Fixing lugs, incorporated in the metal screen—style AB.1652—are available at no extra cost.

EE 69 778 *for further details*

IMPEDANCE METER

Distributed by: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Illustrated above right)

The type GB11 impedance meter manufactured by Radiometer of Denmark is available from Livingston Laboratories Ltd.

The instrument will measure impedance in terms of magnitude and phase-angle from 1Ω to 1·1MΩ and from 0° to ± 90°. The basic accuracy is ± 1 per cent for magnitude and ± 0·5° for phase-angle. Twelve spot test frequencies are provided by the internal genera-

tor and external sources can be used up to 1Mc/s. The test currents can be varied from 3·2μA to 1A. Measurements can be made on impedances which are floating, grounded or balanced to ground, and there is provision for d.c. polarization. Other features include transistorized circuits, in-line read-out of magnitude and a meter providing a direct reading of phase-angle.

Applications include measurement of impedance of filter networks, coils, loudspeakers and transformers, incremental impedance of iron-cored chokes and diodes and the effective impedance of non-linear devices at varying test currents.

EE 69 779 *for further details*

V.L.F. GENERATOR

Distributed by: Claude Lyons Ltd, Hoddesdon, Hertfordshire

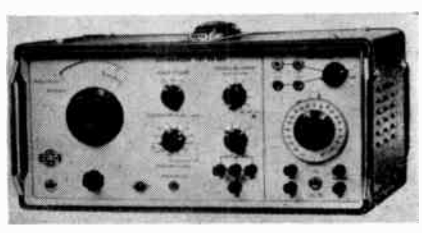
(Illustrated below)

Manufactured by Constructions Radioeléctriques et Électroniques du Centre the type GB 860 provides sine, triangular, or square waves in the very low frequency range from ·001c/s to 1kc/s, with an output power of the order of 0·5W. A variable-phase output is provided, making the instrument suitable for phase-shift measurement. It is especially intended for use in the study of servomechanisms and of mechanical deformations and strains, while other applications include transmission lines, transformers for subsonic frequencies, medical research, geophysics, and to provide precise signals for programming purposes.

The oscillator circuit employs a Schmitt trigger and a Miller integrator, and produces symmetrical triangular or square waves of continuously variable frequency and amplitude, the frequency range being covered in six linearly-calibrated ranges. A waveform-shaping circuit employing cascaded diode clippers produces the sine wave output from the triangular waveform.

A transistorized output power amplifier permits each of these three waveforms to be made available at low impedance, balanced or unbalanced.

A positive trigger pulse, obtained by



differentiation of the square-wave oscillator output, is provided for oscilloscope time-base synchronization. A second pulse output, variable in phase over 360°, makes possible the precise determination of phase shift. When this output is used with a dual-trace oscilloscope, such as the C.R.C. type OC 728, the phase-angle between the generator output and that of the system under test can be measured to better than one degree.

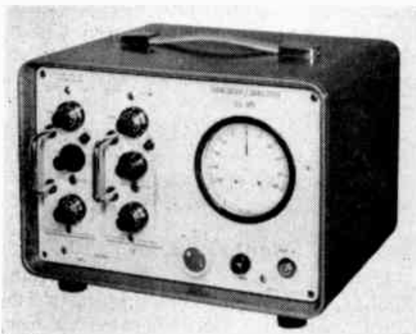
EE 69 780 *for further details*

TRANSDUCER-CONVERTOR

S.E. Laboratories (Engineering) Ltd, North Feltham, Middlesex

(*Illustrated below*)

S.E. Laboratories (Engineering) Ltd has introduced a low-cost portable and self-contained one or two-channel transducer/convertor system S.E 905. This unit is suited for indicating and/or



recording information from inductive or strain-gauge-type transducers.

Pressure, strain, acceleration, displacement measurements etc., may therefore be carried out on site.

A special feature of the S.E. 905, is a mains-operated power supply with a built-in rechargeable battery cell, which enables the unit to be used in the field without recourse to a mains supply. It contains a stable 3kc/s oscillator.

Two types of plug-in amplifiers are available: S.E. 423,—primarily intended for static and dynamic measurements from strain-gauges or strain-gauge type transducers, and S.E. 449—used for static and dynamic measurements from inductive-type transducers and differential transformers.

A built-in cirscale meter can be directly calibrated to customers' requirements. Outputs are also available for driving galvanometer and tape recorders, and oscilloscopes.

EE 69 781 *for further details*

LOW VOLTAGE POWER SUPPLIES

Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire

(*Illustrated above right*)

The L.P. 401 stabilized metered power supply is the first of a new range of high stability, low cost, precision units with a standard of performance comparable with instruments in a much higher price bracket.

0·5A and 1A, 10 to 30V versions are

available with single and twin outputs, the performance specification being similar in each case.

Voltage, current, and the fully adjustable constant current setting, are monitored on the large 3¼in clear scale panel meter without recourse to range switching.

The coarse and fine voltage controls facilitate precise vernier adjustment of the output, the fine voltage control having a total range of ± 0·75V at all output levels.

Particular attention has been given to the design of the constant current control circuits, which provide extremely sensitive stable and accurate control of current limiting over a very wide range, thus ensuring complete protection for sensitive equipment of all types.

All the units in the 400 series will withstand prolonged short-circuiting of the output, without damage.

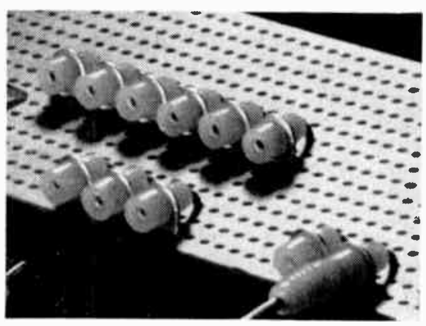
EE 69 782 *for further details*

PRINTED CIRCUIT SOCKET

Oxley Developments Co. Ltd, Priory Park, Ulverston, Lancashire

(*Illustrated below*)

The Oxley printed circuit socket type 50S/PCB has been designed for direct mounting on a 0·1in module printed circuit board with 0·050in minimum diameter holes. The mounting is arranged so that the plugs can be inserted in a plane parallel to the printed circuit board, thus permitting close stacking of boards in an equipment. The unmounted breakdown voltage is 5kV d.c. and the capacitance added to the circuit is less than 0·7pF per outlet. The overall height above the printed circuit board is



0·250in. This socket can be supplied from 1 to 12 ways and accepts the Oxley standard plug type 50P/156 and this permits close spacing of the test points. When a multiple-way type is required the socket outlets can be supplied colour coded—in which case the number of outlets should be given together with the arrangement of colours reading from left to right. Colours available—black, brown, red, orange, yellow, green, blue, violet, pink, grey and white.

EE 69 783 *for further details*

BATCH COUNTER

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(*Illustrated below*)

Advance Controls Ltd has introduced a new timer/batch counter, type TB42.

This instrument can be used to time a process, or to measure a batch continuously or singly. Timing speeds of 0·01sec are available and batch counting

may be undertaken at up to 1 000 a second. The TB42 has instantaneous reset which eliminates any loss of count or time on the repeat cycle, and repeat may be cycled automatically, by remote initiation, or by manual push-button.

Cold-cathode tubes are fitted to provide visual monitoring of correct counting, and the instrument has a pre-batch signalling facility of up to 99 counts or 0·99sec prior to the end of a pre-set count.

The output relay may be arranged to provide an automatic reset after a specified time. Alternatively, it can be reset by an external control.

The TB42 measures 9in wide by 6¼in high by 5¼in deep. A complete range of pick-ups, photo-electric, magnetic, reed-switch and proximity, is available for use with this instrument.

EE 69 784 *for further details*

ELECTRONIC TIMER

Pioneer Designs Ltd, Crown House, Walton-on-Thames, Surrey

(*Illustrated on page 343*)

This timer incorporates a pulse operated trigger circuit with variable sensitivity adjustable by an individual front panel control. Sensitivity can be controlled so that low d.c. and a.c. voltages applied from an external source to a high impedance input trigger circuit will start the timed cycle. Alternatively, the sensitivity can be increased to a point where automatic cycling takes place.

Two time ranges are incorporated in this model, 0·2 to 1·0sec and 1 to 5 secs. The timed period is infinitely variable and is set by a 5in diameter dial.

In addition, the timed cycle may be initiated by a push-button on the front panel, or by short-circuit, e.g., micro-switch, across two terminals.

Output is by relay, acting as a single-pole changeover switch, and make and break terminals are provided for an external circuit separately fused at 2A. By changing an internal connecting link the timer can be arranged to supply 230V, 50c/s, up to 500W at the output terminals. The timer itself consumes less than 8W.

EE 69 785 *for further details*

DISCHARGE-FREE CAPACITORS

F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(Illustrated below)

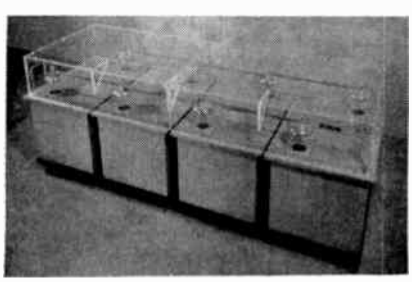
F. C. Robinson and Partners Ltd announce that a complete range of ionization discharge-free capacitors is now available.

When carrying out ionization discharge tests, with any detector, maximum sensitivity is achieved only when the blocking capacitor value is high compared with that of the specimen. With this in mind values from

1 000pF to 0·05μF at 50kV r.m.s.
2 000pF to 0·1μF at 25kV r.m.s.
5 000pF to 0·5μF at 10kV r.m.s.
and 0·01μF to 2μF at 5kV r.m.s.

are being produced as standard items. Higher working voltages or larger capacitance values may be attained by connecting certain units in series or in parallel. The photograph shows a parallel bank of 0·05μF, 50kV models.

Depending on the voltage rating and capacitance value, some types are of the familiar tubular construction, while others are built into oil-filled steel tanks or rigid p.v.c. containers. All are carefully checked at each stage of manufacture to ensure that they are, and remain, discharge-free.



EE 69 786 *for further details*

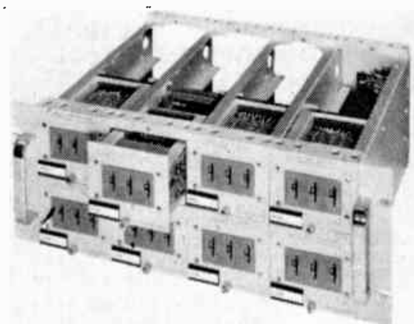
DIGITAL LIMIT SELECTOR

Digital Measurements Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated below)

Digital Measurements Ltd has announced a digital limit selector DM5080 and a logic unit DM5006 which together form an alarm setting system for use with digital voltmeters, digital data recording systems, counters, etc.

Up to three decimal decades can be accepted by the logic unit which will drive up to 16 limit selectors. 'High' limit' selectors produce an output when the input to the alarm setting system is equal to, or greater than, the selected level. 'Low limit' selectors produce an output when the input to the system is below the selected level. The selectors are easily converted from 'high limit'



to 'low limit' and vice versa, by altering link connexions.

The output can be used to initiate automatic remedial action or to inform a plant operator of off-limit conditions by means of lighted panels, audible alarms, etc. Alternatively, where inputs from various sources are being recorded by a logging system, the colour shift on the typewriter or printer can be operated automatically.

EE 69 787 *for further details*

ANALOGUE DIGITAL ELEMENTS

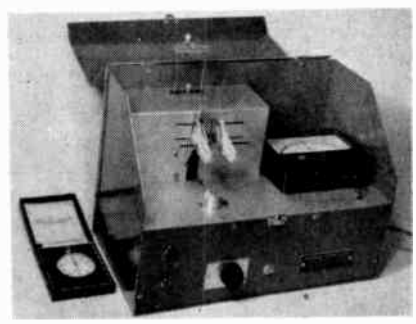
Trainer and Simulator Division, Elliott Brothers (London) Ltd, Chobham Road, Frimley, Nr. Aldershot, Hampshire

An analogue/digital computing or simulator system of any desired configuration can be built up from a new range of standard elements developed by the Trainer and Simulator Division of Elliott-Automation Ltd. These elements, which originated from work on a special simulator of advanced design, enable the systems designer to build to a known accuracy and performance at the minimum cost. The field of application is extremely wide, for example, in the 'breadboard' proving of analogue and digital systems, the teaching of control engineering theory, and for the construction of design or training simulators.

The basic elements, all of which are available separately, consist of a servo electronics board, comprising a combined modulator-preamplifier and power amplifier for driving standard servo-motors, a high-accuracy gearbox with seven output shafts which give gear ratios from 1 to 18 750 for the operation of servo components in the range size 08 to size 15, a demodulator board for conversion of 400c/s signals and general-purpose a.c. or d.c. computing amplifiers. The elements, together with their power supplies, can all be housed in a modular racking system.

These new Elliott analogue elements may be combined to form a wide range of servo systems—for example, a d.c. position servo or a velocity servo with d.c. or 400c/s a.c. inputs, and synchro follow-up or resolver servos. Outputs may be in the form of shaft rotations or signals derived from potentiometers, synchros, resolvers and shaft encoders. Shaft-angle indication can be displayed on the front panel of the system with the aid of couplings from the gearbox shafts.

EE 69 788 *for further details*



SOLDERABILITY TEST MACHINE

Multicore Solders Ltd, Maylands Avenue, Hemel Hempstead, Hertfordshire

(Illustrated above)

The demand for increased production rates in the assembly of electronic equipment has created a need for a simple, accurate method of assessing the solderability of component termination wires.

Multicore Solders Ltd, in collaboration with the Electronic Engineering Association, has developed a machine claimed to be the only equipment capable of testing wire ends for solderability.

In operation, the solderability test machine smoothly lowers a specimen wire, previously fluxed, on to a molten drop of solder. The time in seconds for the solder to flow around the wire and unite above it is a measure of the solderability of that wire. A built in power pack supplies the thermostat, which controls the heater, the temperature being indicated by a battery operated pyrometer with an accuracy of 1 per cent.

The solderability test machine is supplied complete with pyrometer, stopwatch calibrated to 1/10th of a second, instruction manual, samples of solder pellets and liquid flux, packed for export.

EE 69 789 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF CIVIL ENGINEERS

All meetings will be held at Great George Street, Westminster, S.W.1, at 5.30 p.m., unless otherwise stated.

Date: 7 May.
Discussion: Education for Management of Engineers in the U.S.A.
By: F. H. Barker-Benfield.
Date: 12 May.
Lecture: The International Hydrological Decade.
Date: 28 May.
Discussion: Safety Fences.
By: R. L. Moore and V. J. Jehu.

INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m., unless otherwise stated.

I.E.E./R.Ae.S. London Joint Group on the Applications of Electricity in Aircraft

Date: 26 May. Time: 6 p.m.
Held at: The Royal Aeronautical Society, 4 Hamilton Place, W.1.
Discussion: Designing and Engineering Aircraft Electrical Systems for Reliability.
Opened by: G. G. Wakefield.

Electronics Division

Date: 4 May.
Lecture: The Engineering and Scientific aspects of the Canadian Ionospheric Satellite.
By: E. D. R. Shearman and J. W. King.
Date: 6 May.
Lecture: Image Intensifiers.
By: R. L. Beurle.
Date: 11 May.
Discussion: Are Transistors Reliable Enough.
Opened by: J. M. Groocock, W. R. McLachlan, F. F. Roberts and M. Young.
Date: 12 May.
Lecture: U.H.F: Television Receiving Aerials.
By: C. F. Whitbread.
Date: 13 May.
Lecture: Second Annual Lecture.
By: A. F. Huxley.
Date: 25 May.
Lecture: Results of the Extraordinary Administrative Radio Conference, Geneva, October-November 1963.
By: C. F. Booth.
Date: 27 May.
Lecture: The Development of Secondary Surveillance Radar for Air Traffic Control.
By: D. G. Terrington.

Power Division

Date: 6 May.
Lecture: The Goodness of a Machine.
By: E. R. Laithwaite.
Date: 14 May.
Lecture: Control of Motor Voltage in a.c. Traction.
By: E. A. K. Jarvis.
Date: 20 May.
Lecture: The Dutch Delta project; shortening the line of defence against sea.
By: J. Volkers.

Science and General Division

Date: 5 May.
Lecture: Rock Magnetism and Continental Drift.
By P. M. S. Blackett.
Date: 7 May.
Lecture: Finite Expansions and Orthogonalisation.
By: J. K. Lubbock.
Date: 8 May. Time: 2.30 p.m. and 5.30 p.m.
Colloquium: Equipment for Standards Laboratories.
Date: 12 May.
Discussion: The new HNC.
Opened by: A. D. Collop.
(Joint meeting with the I.E.R.E. Education and Training Group.)

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, unless otherwise stated.

Date: 13 May. Time: 10.30 a.m.-5 p.m.
Held at: Birkbeck College, Malet Street, London, W.C.1.
Symposium: Modern Techniques for Recording and Processing Seismic Signals.
(Advance registration will be neccessary and application forms may be obtained from the Institution at 9 Bedford Square, W.C.1, on request.)
Date: 26 May. Time: 6 p.m.
Lecture: The Fifth Clerk Maxwell Memorial Lecture.
By: Sir W. G. Radley.
(Admission will be by ticket only.)

Radar Group

Date: 6 May. Time: 6 p.m.
Lecture: Long-term Air-traffic Control System Concepts.
By: H. Jessell.

Television Group

Date: 7 May. Time: 6 p.m.
Papers on: The New Test Cards D and E.
Date: 21 May. Time: 6 p.m.
Lecture: The Plumbicon Tube.
By: E. F. de Haan and S. L. Tan.

Joint I.E.R.E./I.E.E. Education Group

Date: 12 May. Time: 5.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London. W.C.2.
Discussion: The New Higher National Certificate.

Merseyside Section

Date: 27 May. Time: 7.30 p.m.
Held at: The Walker Art Gallery, Liverpool.
Lecture: Video Telephones.
By: H. A. Mumford.

SOCIETY OF ENVIRONMENTAL ENGINEERS

Date: 6 May. Time: 6 p.m.
Held at: The Royal Aeronautical Society, Hamilton Place London, W.1.
Lecture: RAE Farnborough on the Development of Test Facilities to Simulate Thermal Environments Encountered in Supersonic Aircraft.
By: J. Rudman.
Date: 13 May. Time: 6 p.m.
Held at: The Imperial College of Science, Mechanical Engineering Department, Exhibition Road, London, S.W.7.
Discussion: Vibration Damping in Engineering, to be followed by the Annual Dinner.
By: P. Grootenhuis.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Systems Engineering Section

Date: 6 May. Time: 7 p.m.
Held at: Manson House, 26 Portland Place, London, W.1.
Lecture: A Punched Card System for the Speed Control of an Extrusion Press.
By: P. W. Hicks and C. D. M. Johnston.

PUBLICATIONS RECEIVED

MINISTAC MANUAL — STC Semiconductor Division (Rectifiers) has recently produced a manual describing the various ways that their Ministac modules can be used to accommodate complete circuits of standard components. The Ministac system of modular circuit construction enables a large quantity of passive and active components to be assembled in the minimum space, with the inside conducting walls of the modules forming pre-punched circuit paths. Reference is made to the Ministac Kits now available which provide the design engineer with the materials and simple tools to enable him to make up prototype Ministacs. The Manual (Publication MF/501) is available on request from STC Semiconductor Division (rectifiers), Edinburgh Way, Harlow, Essex.

ERICSSON DSU EQUIPMENT FOR TELEPHONE EXCHANGES is the title of two booklets describing this newly developed system for the housing of electronic and/or electromechanical equipment. Copies of these booklets B63/06/1 and /1A describing in detail this new approach to equipment accommodation are available on request to Ericsson Telephones Ltd, D.S. Dept., Beeston, Nottingham.

STEROMOTORS is the title of an advance technical bulletin on the subject of the new Steromotor which is shortly to be manufactured in this country by Evershed and Vignoles Ltd. Copies are obtainable free on request to the company at Acton Lane Works, Chiswick, London, W.4.

THE WELLS SERIES OF NUCLEONIC INSTRUMENT MODULES—A new booklet which describes the Wells series of Nucleonic Instrument Modules is now available from EMI Electronics Ltd. These all-transistor modular units can be assembled in any order to provide a wide variety of facilities in the nucleonic instrument field. Individual instruments which can be made from the modules include a simple or an automatic counting system, a gamma spectrometer, and a simple or a comprehensive ratemeter for Geiger-Muller and scintillation counter probes. These can all be mains or battery operated. Plug-in printed circuit boards make for ease in servicing. A special panel can be supplied so that the modules will fit into a standard 19in Copies of this booklet can be obtained from EMI Electronics Ltd, Instrument Division, Hayes, Middlesex.

ELECTRICAL INSTRUMENTS is the title of a 36-page illustrated publication, List 163/2, giving abridged specifications and performance data for several hundred electrical instruments. It describes a number of entirely new instruments. including a 6-dial precision bridge with an accuracy of $\pm 0{\cdot}005$ per cent over the greater part of its range (1Ω—111MΩ), a multi-purpose portable potentiometer for laboratory and workshop use, an a.c./d.c. comparator with an accuracy of $\pm 0{\cdot}05$ per cent between 25c/s and 10kc/s and a delta resistance bridge that measures deviations of resistance and temperature coefficients of resistance. Copies of the booklet are available on request to the Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1.

MULLARD ELECTRONIC POWER CONTROL DEVICES is the title of a pamphlet giving brief information on the current range of devices—thyratrons, ignitrons and thyristors—for the design and maintenance engineer. Also included are lists of Mullard equivalents for other internationally accepted types. Requests for copies of this publication should be addressed the Mullard Ltd, Government and Industrial Valve Division, Mullard House, Torrington Place, London, W.C.1.

EE 69 **170** *for further details*

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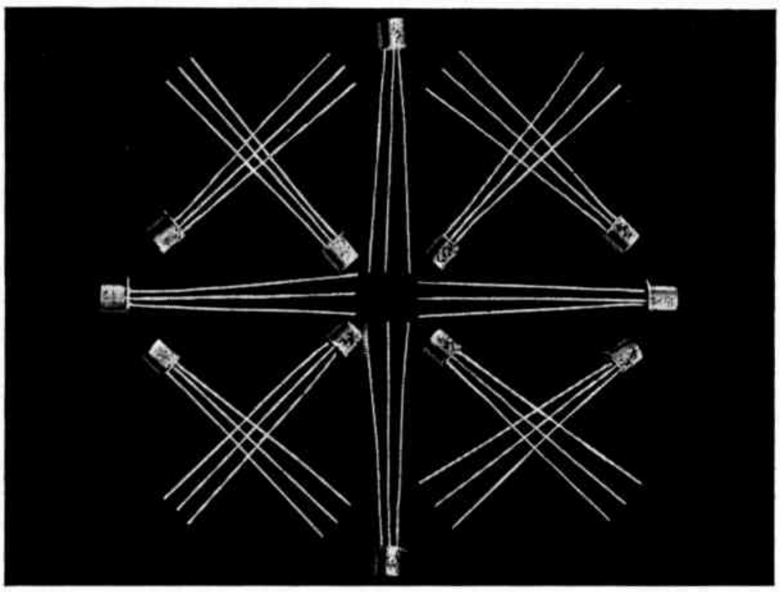

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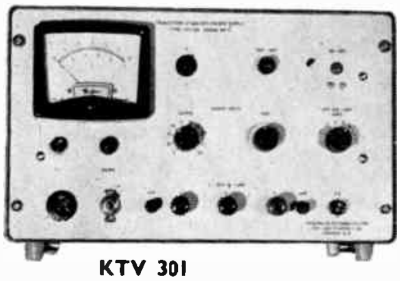
EE 69 171 *for further details*

KSM TRANSISTOR STABILISED POWER SUPPLIES



KTV 301

Continuously variable output voltage. All units can be supplied for constant voltage or constant current output. We are specialists in this field and we welcome your special requirements. Our other power supplies are silicon controlled rectifier and valve types giving voltages up to 10kV regulated.

MODEL	D.C. OUTPUT VOLTAGE VOLTS	D.C. OUTPUT CURRENT AMPS	OUTPUT RESISTANCE OHM	RIPPLE PEAK-TO-PEAK M.V.	CABINET DIMENSIONS IN INCHES H.	W.	D.	WEIGHT LBS.
KTV 35	0–30	½A	.01	1	9¼	13	9	16
KTV 301	,,	1	,,	,,	7¼	12¼	7½	18
KTV 302	,,	2	,,	,,	9¼	13	9	30
KTV 303	,,	3	,,	,,	,,	,,	,,	45
KTV 305	,,	5	,,	,,	12	21⅞	11½	60
KTV 310	,,	10	.008	,,	12	21⅞	16½	80
KTV 320	,,	20	.005	2	16½	21	16¼	110
KTV 330	,,	30	.003	,,	21	17¼	18	160
KTV 340	,,	40	.002	5	24	17¼	18	220
KTV 350	,,	50	.001	,,	27	20	20	300
KTV3100	,,	100	.001	,,	60	24	24	
KTV 501	0–50	1	.01	1	9¼	13	9	20
KTV 502	,,	2	,,	,,	,,	,,	,,	30
KTV 503	,,	3	,,	,,	12	21⅞	11½	45
KTV 505	,,	5	,,	,,	,,	,,	,,	65
KTV 510	,,	10	.008	,,	12	21⅞	16½	80
KTV 520	,,	20	.005	,,	16½	21	16½	110
KTV1001	0–100	1	.01	2	12	21⅞	11½	45
KTV1002	,,	2	,,	,,	,,	,,	,,	50
KTV1003	,,	3	,,	,,	12	21⅞	16½	60
KTV1005	,,	5	,,	,,	,,	,,	,,	85
KTV1010	,,	10	.008	5	21	20	20	125
2TV 35	2×0–30	2×½A	.01	1	20	9½	7¼	28
2TV 301	,,	2×1A	,,	,,	12	21⅞	11½	35
2TV 302	,,	2×2A	,,	,,	,,	,,	,,	45
2TV 303	,,	2×3A	,,	,,	,,	,,	,,	60
2TV 305	,,	2×5A	,,	,,	12	21⅞	16½	75
2TV 501	2×0–50	2×1A	,,	,,	12	21⅞	11½	40
2TV 502	,,	2×2A	,,	,,	,,	,,	,,	48
2TV 503	,,	2×3A	,,	,,	,,	,,	,,	65
2TV 505	,,	2×5A	,,	,,	16½	21⅞	16½	80

KTV 350

All models can be supplied in cabinets, on 19" standard P.O. Panels, or as open chassis units. All models supplied for 200/250V 50c/s except KTV3100 which uses 440V 3Ph 50c/s. Other voltages can be supplied.

PROTECTION: In the event of a short circuit or overload the constant voltage output is changed into a constant current output where the current is limited to a pre-determined value. As a further safeguard a cut-out is fitted. This cut-out can be set to trip at any desired current by means of a calibrated front panel control.

All models have four terminal outputs to eliminate volt drop in leads to the load. The output terminals are fully floating and a separate earthing terminal is provided.

All models can be supplied as constant current units as well as constant voltage.

VOLTAGE STABILISATION: .02%.

KSM ELECTRONICS

(PROPRIETORS OF KASAMA ELECTRONICS)

139-149 FONTHILL ROAD, FINSBURY PARK, LONDON, N4

TELEPHONE: ARCHWAY 6160

Le Salon International des Composants Électroniques

Une description, basée sur des renseignements fournis par les fabricants, de certains des composants français exposés au Salon International des Composants Electroniques à Paris du 7 au 12 février 1964

Traduction des pages 336 à 339

COGIE

3 et 5, boulevard Anatole-France, Aubervilliers (Seine)

BLOCS REFROIDISSEURS THERMOÉLECTRIQUES

(*Illustration à la page 336*)

Ces blocs thermoélectriques 'Frigogies' (à effet Peltier), type FC. 8-7 A se composent de huit couples au tellurure de bismuth assemblés par isolement thermique et reliés en série électriquement. L'évacuation de chaleur maxima est de 32 watts.

En changeant la polarité de la tension d'alimentation en continu on inverse les jonctions chaude et froide ce qui assure un excellent moyen de régulation de température. La différence maxima entre les jonctions chaude et froide est de 57° C, la jonction chaude étant à 25° C.

Il existe également des modèles de format plus réduit.

EE 69 751 *pour plus amples renseignements*

C.O.P.R.I.M.

(Compagnie des Produits Élémentaires pour Industries Modernes)

7, Passage Charles-Dallery, Paris 11e

POTENTIOMÈTRES MINIATURES

(*Illustration à la page 336*)

Ce nouveau modèle (E 086) appartient à la famille des potentiomètres à piste de carbone; il est conçu pour insertion sur câblage imprimé, au pas de 2,54 mm.

Il est plus petit que le modèle E097, mais de présentation et technologie analogue, et sa puissance nominale est de 0,1 W à 40° C.

Ces pièces répondront aux désirs des utilisateurs, partout où un gain de place est souhaitable et dans le cas de puissances faibles.

Les valeurs s'échelonnent de 100 Ω à 20 kΩ et peuvent atteindre 500 kΩ sur demande.

EE 69 752 *pour plus amples renseignements*

ELNO

(Etabl. Lailler-Pecquet et Cie)

18 et 20, rue du Val-Notre-Dame, Argenteuil (Seine-et-Oise)

ÉQUIPEMENT DE TÊTE LÉGER

Cet équipement de tête léger pèse moins de 150 grammes et comprend un serre-tête en fil d'acier, enrobé de mousse polyester gainée plastique qui supporte à l'une de ses extrémités un boîtier écouteur léger portant sur sa face antérieure une oreillette et sur sa face postérieure le dispositif d'amarrage du bras porte microphone. Le bras porte microphone est orientable verticalement et latéralement. Les cordons légers du boîtier écouteur et du microphone sont reliés à la boîte de raccordement et fixés au vêtement de l'utilisateur.

La gamme de fréquence de l'écouteur est de 20 Hz à 4 kHz ±3 dB. Six impédances standard de 15 Ω à 2 kΩ sont prévues.

La sensibilité du microphone est de 0,1 mV/μB. La courbe de réponse est ascendante à raison de 6 dB par octave jusqu'à 1,5 kHz, puis plate jusqu'à 4 kHz. Le microphone est à 6 impédances standard de 15 Ω à 2 kΩ.

EE 69 753 *pour plus amples renseignements*

HAUTE PARLEUR D'OREILLER

(*Illustration à la page 336*)

Cet appareil est spécialement conçu pour l'écoute d'un programme de radiodiffusion par des personnes alitées dans des hôpitaux. Il comporte un boîtier circulaire de section ellipsoïdale ayant sur chacune de ses faces un évent de diffusion du son et un écouteur électromagnétique. Ce dernier est à 10 impédances standard de 15 Ω à 18 kΩ.

EE 69 754 *pour plus amples renseignements*

FÉRISOL

18, avenue Paul-Vaillant-Couturier, Trappes (Seine et Oise)

LIGNES COAXIALES DE MESURES

Les lignes de mesures type UB 200 et UA200 sont les lignes coaxiales fendues d'impédance—50 Ω pour le type UB200 ou 100 Ω pour le type UA200.

Elles sont équipées d'un chariot mobile identique comportant un dispositif plongeur en profondeur et destiné à recueillir la tension HF. Ce dispositif est apériodique et ne comporte pas de système d'accord. La tension HP est disponible sur une fiche "N". Le chariot mobile contient également un détecteur coaxial à cristal destiné à détecter la tension HF, modulée ou non. On peut ainsi mesurer le T.O.S. de l'impédance étudiée en associant à la ligne de mesure l'indicateur de T.O.S. type AG.101.

Le déplacement mécanique du chariot sur la ligne de mesure est repéré avec précision sur une échelle graduée en centimètres et millimètres et un vernier gradué en dixième de millimètre. La précision de mesure ainsi obtenue est de l'ordre de 0,1 mm. Une précision de mesure plus élevée peut être obtenue (0,01 mm) pour des faibles déplacements (course maximum 50 mm) par l'utilisation d'un dispositif spécial de fixation pour comparateur mécanique au 1/100°.

EE 69 755 *pour plus amples renseignements*

MÉGOHMMÈTRE ÉLECTRONIQUE

Le mégohmmètre type RM200 est un appareil permettant la mesure des résistances dont la valeur est comprise entre 0,2 MΩ et 10^8 MΩ.

La valeur de la résistance inconnue est affichée en lecture directe sur le cadran d'un galvanomètre grâce à une nouvelle technique de la mesure de très faibles tensions continues. Contrairement aux dispositifs classiques à pont de Whetstone où l'opérateur est obligé de remettre le pont en équilibre à chaque variation de l'inconnue, le nouveau mégohmmètre type RM200 affiche cette valeur sans aucune intervention extérieure. Il permet de ce fait non seulement la mesure d'une valeur instantanée de l'inconnue, mais l'étude complète de son comportement dans le temps, sous l'influence d'autres paramètres tels que température, humidité, etc. Une sortie "enregistreur" délivre dans ce but une composante continue dont la valeur est liée par une fonction simple à la valeur de l'inconnue.

Les nombreuses tensions d'essai fournies par l'appareil lui-même en font un instrument universel, adapté aux techniques les plus modernes. Ainsi la tension d'essai 10 volts permet l'étude des composants relevant de la technique transistor.

D'autre part, la structure de l'appareil permet l'utilisation d'une tension d'essai provenant d'une source extérieure de valeur quelconque. Un calcul très simple établit la valeur réelle de l'inconnue, compte tenu de la valeur affichée par le cadran du galvanomètre de mesure et de la tension d'essai appliquée.

EE 69 756 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Lancashire

AMPLIFICATEURS PARAMÉTRIQUES

(Illustration à la page 337)

L'amplificateur paramétrique de 570 MHz (type VCA/L10) donne une amplification de 16 dB pour une largeur de bande de 6 MHz (par rapport aux points de 1 dB) et une valeur totale de bruit inférieure à 1 dB. Il est intéressant de noter que cette valeur de bruit est bien moindre que la valeur du bruit de fond provenant du ciel reçu à cette fréquence.

La fréquence de 570 MHz est la fréquence d'émission du nouveau programme de télévision à 625 lignes de la British Broadcasting Corporation, et le nouvel amplificateur a été principalement réalisé en vue de cette émission. Une commande de ces amplificateurs a été passée par Rediffusion Research Ltd qui s'en servira pour permettre à ses abonnés à la télévision de recevoir le nouveau programme sur un circuit de télédiffusion à Southampton.

Un autre amplificateur paramétrique exposé était le modèle VCA/S30. Cet amplificateur a été conçu pour le fonctionnement à une fréquence centrale de 3000 MHz et il permet une amplification de 20 dB pour une largeur de bande de 70 MHz (par rapport aux points de 3 dB) et un facteur de bruit de 2,7 dB. Les principales applications des amplificateurs paramétriques pour les fréquences de bande S, sont dans les récepteurs radar où ils augmentent la portée radar effective d'un facteur minimal de trois.

EE 69 757 *pour plus amples renseignements*

CELLULES PHOTOVOLTAÏQUES

(Illustration à la page 337)

Une nouvelle cellule photovoltaïque au silicium à grande surface, appelée type MS40, a été exposée. Elle est principalement prévue comme dispositif efficace et économique pour la transformation de l'énergie solaire en courant électrique pour systèmes d'emmagasinage par batterie tampons. La cellule MS40 se compose de quatre cellules individuelles reliées en parallèle à une source froide commune. Un encapsulage à lentille cylindrique moulé en une seule pièce augmente la surface effective de captage des rayons de soleil et cela, allié à d'autres caractéristiques de réalisation, assure une efficacité de conversion du spectre solaire d'environ 8 % au niveau du sol.

Avec une source de tungstène à 2854° K (incidence normale) et une densité d'éclairement de 3000 lumens par pied carré, la nouvelle cellule fournira (à une température ambiante de 25° C) un courant minimum de 5000 mA dans une charge de 0,6 Ω. Une attention particulière a été apportée lors de la réalisation à la forte corrosion et aux contraintes mécaniques ambiantes auxquelles les panneaux solaires construits avec des modules MS40 seront soumis. Les montants d'assemblage et connexion électrique sont en alliage Monel 400 dont le taux de corrosion est suffisant pour les milieux marins directs. Un encapsulage robuste en époxyde plastique est utilisé qui résiste à la gamme de températures ambiantes de fonctionnement et d'emmagasinage de −20° C à +85° C.

Un assemblage d'essai de modules MS40 vient d'être expédié au Golfe Persique où il sera essayé sur des structures marines fixes afin de pouvoir obtenir des données complémentaires pour compléter les essais de service. Cet essai en service final est effectué pour déterminer la réalisation optimale d'alimentations à grande échelle en énergie solaire-électrique pour l'utilisation dans des lieux éloignés.

EE 69 758 *pour plus amples renseignements*

HEWLETT-PACKARD S.A.

54 Route des Acacias, Genève, Suisse

GÉNÉRATEURS DE BALAYAGE MICRO-ONDES

(Illustration à la page 337)

Les générateurs de balayage, types 691A/B (1 à 2 GHz) et 692A/B (2 à 4 GHz) comportent trois modes de balayage. Dans le premier mode, appelé "normal," la fréquence de sortie est balayée dans les deux sens entre deux limites à réglage continu. Dans le mode " F," la fréquence est balayée autour d'une fréquence médiane réglable. De plus, ces dispositifs de balayage comprennent deux marqueurs de fréquence à réglage indépendant pouvant être réglés numériquement avec une résolution de ±0,05 % de la fréquence de sortie. Ces marqueurs modulent en amplitude la sortie dans les modes "F" et "normal", un avantage très important pour les enregistrements x-y ou l'observation du balayage d'un oscilloscope. Dans un troisième mode de balayage, on peut employer ces mêmes marqueurs de fréquence pour établir un mode de balayage semblable au mode "normal" et appelé "balayage marqueur" dans lequel les marqueurs déterminent la gamme de balayage. La sélection par marqueur permet (par exemple) de balayer d'abord toute la bande passante d'un filtre par le mode normal, puis de balayer la zone de rejet du même filtre en passant au "balayage marqueur" sans réajuster les limites de balayage.

Les types 691B et 692B emploient des sections à bande à diodes pour l'atténuation et la modulation. La modulation d'amplitude qu'effectuent ces diodes à broche est entièrement indépendante du tube de sorte que tout glissement de fréquence est virtuellement éliminé. Ces réseaux à diodes à broche sont également utilisés pour produire un nivellement du signal de sortie de 0,3 dB.

EE 69 759 *pour plus amples renseignements*

RÉCEPTEUR DE FRÉQUENCE STANDARD

(Illustration à la page 337)

Le récepteur type 5090A a été conçu pour la réception des transmissions de 200 kHz de Droitwich. Etant donné que la stabilité de la fréquence porteuse de Droitwich est de 5×10^{-10} par jour, il constitue un moyen simple et peu coûteux d'obtenir un étalon de fréquence d'une grande précision à l'usage des laboratoires.

Le récepteur 5090 A a des fréquences de sortie de 100 kHz et de 1 MHz.

EE 69 760 *pour plus amples renseignements*

I.E.R.

(Impression Enregistrement de Résultats)

6, rue Blondel, Courbevoie (Seine)

IMPRIMANTE SUR BANDE

(Illustration à la page 337)

L'imprimante numérique rapide sur bande à entrée parallèle, modèle 205, peut recevoir des résultats de mesures exprimés en code binaire-décimal et assure automatiquement le décodage et l'impression à la cadence maximale de 10 lignes par seconde. La capacité maximale étant de 13 colonnes, le débit maximal est de 130 caractères par seconde.

Elle a été développée et mise au point à la suite de l'expérience acquise avec le modèle 203 fonctionnant à 6 lignes par seconde.

EE 69 761 *pour plus amples renseignements*

LIE-BELIN

296, avenue Napoléon-Bonaparte, Rueil-Malmaison, (Seine-et-Oise)

ALIMENTATION T.H.T.

(Illustration à la page 338)

Cette alimentation T.H.T. entièrement transistorisée est présentée en tiroir standard Esone 2M ou en version capotée avec poignée de transfert.

Ses caractéristiques principales sont les suivantes:

Tension de sortie: 2 kV, 2 mA.
Stabilité: 5.10^{-5} pour ±10 % de variation de tension secteur.
Résistance interne: 50 kΩ.
Réglage de la tension: par bonds de 80 V. réglage fin; par potentiomètre 10 tours.
Contrôle: par voltmètre sur face avant.
Alimentation: par secteur 50 Hz 127 ou 220 V ou 24 V c.c.

EE 69 762 *pour plus amples renseignements*

ANALYSEUR AUTOMATIQUE DE FILTRES

(Illustration à la page 338)

Cet appareil est destiné à la mesure automatique ou manuelle de l'activité α ou β des filtres de prélèvement de poussières.

Chaque filtre à mesurer est amené devant un photomultiplicateur suivi d'une électronique de comptage, avec affichage et impression des résultats.

Dans le cas de la détection α, le scintillateur (Zns) est propre à chaque filtre et placé contre celui-ci dans le porte-filtre. Pour la détection B, le scintillateur est unique et fait corps avec le photomultiplicateur.

L'appareil se compose d'un coffret divisé en deux parties. La partie supérieure comprend le détecteur (54 AVP) avec préamplificateur, amplificateur et mise en forme et éventuellement un château en plomb (analyse β); le magasin pour les porte-filtres, les dispositifs d'avance du magasin, de prélèvement des filtres dans le magasin, d'avance et de retour des filtres, d'impression des résultats sur chaque porte filtre et sur bande de papier; l'électromécanique avec le programmeur, le relayage et les sécurités.

La partie inférieure comprend un châssis standard 4U3 contenant l'électronique et les alimentations avec 5 décades de comptage, avec affichage et impression par monodécades mécaniques.

La capacité des chargeurs est de 60 filtres de 160 × 160 mm.

EE 69 763 *pour plus amples renseignements*

MÉTRIX

(Compagnie Générale de Métrologie)

Chemin de la Croix-Rouge, Annecy, (Haute-Savoie)

TRANSISTORMÈTRE

(Illustration à la page 338)

Le transistormètre type 302A a été créé pour vérifier les diodes de détection et de redressement, les diodes Zener, les transistors de quelques milliwatts jusqu'aux transistors de puissance.

Le contrôle de ces éléments réside en une vérification du courant inverse, mesuré à l'aide d'un galvanomètre de grande sensibilité. Une résistance en série dans le circuit protège le galvanomètre en cas de branchement défectueux. La diode est soumise pendant ces essais à une tension ne dépassant pas 4 volts.

Lorsqu'on contrôle des diodes Zener, la diode est placée en série avec une source de 18 V et une résistance de limitation du courant.

L'appareil de mesure, dont la consommation est négligeable, est branché aux bornes de la diode et mesure la tension régulée par celle-ci.

La puissance dissipée dans la diode est très faible (30 mW environ).

Dans le cas des transistors, la mesure du courant ICBO est identique à la mesure du courant inverse des diodes.

Pour la mesure du gain statique h21E ou hFE, les circuits de l'appareil sont conçus pour permettre l'ajustage du courant collecteur à l'une des valeurs 1, 10, 100 mA ou 1 A.

Lorsque cette opération est effectuée, la tension collecteur-émetteur est de 2 volts. La puissance dissipée dans le transistor est donc respectivement de: 2 mW, 20 mW, 200 mW ou 2 W, suivant le courant collecteur choisi. Cette puissance est au maximum au moment de la mesure. Il n'y a donc aucun risque de surcharge pendant l'opération de tarage du courant collecteur.

Cette opération préliminaire étant effectuée, il suffit de mesurer le courant de base pour obtenir le gain statique exprimé par le rapport $\frac{IC}{IB}$. Le galvanomètre mesurant IB est gradué directement en valeur du gain.

EE 69 764 *pour plus amples renseignements*

ORÉGA

(Société Électronique et Mécanique)

106, rue de la Jarry, Vincennes (Seine)

BLOC D'ACCORD DE TÉLÉVISION

(Illustration à la page 338)

Appelé "Rotomatic," ce bloc d'accord/amplificateur très haute fréquence de format très réduit pour télérécepteurs est prévu pour l'accord mécanique ou électrique.

Le modèle à lampes serait vraisemblablement le plus petit sur le marché et sa spécification technique correspond au système à 819 lignes. Il a un gain élevé et un facteur de bruit réduit de 5 à 7 dB. Il utilise des lampes types ECC189 et ECF801.

Il existe également une version à transistors à grand gain et à facteur de bruit encore plus faible. La consommation électrique est de 17 mA à 12 V.

Les deux modèles sont pour les normes françaises ou étrangères.

EE 69 765 *pour plus amples renseignements*

QUENTIN ET CIE

2, Rue Hoche, Ermont (Seine-et-Oise)

ALIMENTATIONS STABILISÉES

(Illustration à la page 339)

Ces alimentations stabilisées sont conçues pour une alimentation secteur de 110 à 127 V ou de 220 à 240 V, de 48 à 65 Hz. Elles fournissent une tension de sortie de 0 à 150 V. Trois modèles sont prévus: AS-280, AS-281 et AS-282, dont les valeurs nominales de sortie sont de 0 à 1 A, 0 à 2 A et 0 à 5 A respectivement.

La régulation de la tension de sortie est supérieure à 0,05 % pour une variation de tension d'entrée de ±15 %. La régulation de charge est supérieure à 0,5 % pour une variation de tension de sortie de 0 au maximum.

La stabilité est telle que la tension de sortie varie de moins de 0.1 % en huit heures, après trente minutes de fonctionnement. La tension de sortie varie de moins de 0.5 % par °C pour des changements de température ambiante de −10° C à +50° C. L'ondulation est inférieure à 3 mV crête.

L'intensité de sortie est progressivement réglable de 20 % à 110 % de la sortie nominale.

EE 69 766 *pour plus amples renseignements*

RIBET-DESJARDINS

13-17, rue Périer, Montrouge (Seine)

OSCILLOSCOPE TRANSISTORISÉ

L'oscilloscope miniature 349A, entièrement transistorisé réunit sous un faible volume un véritable appareil de mesure de performances élevées. Il fonctionne sur secteur ou sur batterie rechargeable.

L'amplificateur vertical a une bande passante, en continu, de 0 à 1,5 MHz et un coefficient de déviation de 50 mV/div et, en alternatif, de 8 Hz à 1,5 MHz et un coefficient de déviation de 10 mV/div. Il comporte, en outre, un atténuateur calibré, compensé, à plots de 10 mV/div à 20 V/div.

La synchronisation peut être automatique intérieure ou extérieure. La tension minima nécessaire est de 10 V crête à crête et l'entrée est protégée contre les surtensions jusqu'à 300 V.

Un calibreur incorporé délivre des signaux rectangulaires de 1 V crête à crête.

EE 69 767 *pour plus amples renseignements*

SOCAPEX

9, rue Edouard-Nieuport, Suresnes (Seine)

CONNECTEURS MULTIBROCHES

(Illustration à la page 339)

Les connecteurs de la série 67—connecteurs hermaphrodites empilables à 6 contacts (3 mâles + 3 femelles) à tension de service élevée—se sont très sensiblement améliorés au cours de l'année.

Les connecteurs sous capots ont maintenant un de leurs pavés d'extrémité, ainsi que le crochet de verrouillage correspondant, d'une couleur différente de façon à faciliter le détrompage.

Le principe de marquage adopté permet de repérer instantanément un contact quelconque, qu'il soit sur fiche ou sur embase.

EE 69 768 *pour plus amples renseignements*

SOVIREL

27, rue de la Michaudière, Paris 2e

RÉSISTANCES ET CONDENSATEURS MINIATURES

(Illustration à la page 339)

Les résistances à couches d'oxyde

métallique des domaines haute stabilité et très haute stabilité (de 1/8 Ω à 6 kΩ) et les condensateurs à diélectrique verre, allant de 0,5 pF à 0,01 μF avec un angle de pertes inférieures à 10×10^{-14}, permettent de résoudre les problèmes de qualité et d'encombrement posés par les circuits modernes.

Les résistances à fiabilité prouvée dans les types CR, A et HR, dont le taux de défaut atteint 0,00069 %/1000 h (70 000 résistances essayées pendant 2000 heures, à 2,5 fois la dissipation nominale), constituent les nouveautés 1964.

Ces dernières comprennent également des condensateurs miniatures à diélectrique verre de 0,1 à 30 pF, spécialement conçus pour les blocs d'accord.

EE 69 769 *pour plus amples renseignements*

TACUSSEL SOLEA

2 et 4, rue Carry, Lyon, 3 (Rhone)

MULTIMÈTRE ÉLECTRONIQUE

(Illustration à la page 339)

Le multimètre électronique type VE-7 se caractérise par une impédance d'entrée très élevée lorsqu'il est utilisé comme voltmètre. L'instrument comprend un amplificateur muni d'un tube électromètre et de dix transistors (la plupart au silicium). Il est alimenté par batteries intérieures.

Ses fonctions principales sont la mesure de: tensions continues de 1 mV à 1 kV; précision de déviation totale de 1 %; impédance d'entrée infinie jusqu'à 3 V.

—courants continus de 1 pA (10^{-12} A) à 1 mA; précision de déviation totale de 1 %.

—résistances de 2 Ω à 1 000 000 MΩ; précision de 2 à 3 %.

Plusieurs accessoires sont prévus pour les mesures suivantes: intégration de quantité d'électricité, condensateurs, légères variations de tension, conductances, valeurs de pH.

EE 69 770 *pour plus amples renseignements*

DISTORSIOMÈTRE

(Illustration à la page 339)

Le distorsiomètre harmonique, type DH2, est un instrument à large bande pour la mesure de valeurs de distorsion harmonique de 0,05 à 100 %, en signaux dont les fréquences fondamentales peuvent s'étendre de 3 Hz à 300 kHz. Il comporte un filtre électronique à pont de Wien et un millivoltmètre électronique alternatif pour l'examen oscilloscopique du résidu (harmoniques).

Il peut ainsi être employé comme millivoltmètre alternatif pour des signaux de 0,5 mV à 50 V (3 Hz à 1 MHz).

EE 69 771 *pour plus amples renseignements*

TELEFUNKEN A.G.

Ernst-Reuter-Platz 7, 1 Berlin 10, Allemagne

BLOC D'ACCORD DE TÉLÉVISEUR UHF

Ce dispositif d'accord a été conçu pour la bande de télévision de 470 à 860 MHz. Son robuste coffret plat est divisé en quatre espaces d'une grandeur presque égale au moyen de cloisons qui constituent des circuits de tank et peuvent être accordés à la fréquence voulue à l'aide d'un triple assemblage de condensateurs rotatifs. Le dispositif d'accord type 132, pour l'accord manuel, contient des lampes du type EC88 et EC86. Le bloc d'accord type 140 est logé dans le même coffret que le type 132. Il est muni de deux transistors AF139, le premier fonctionnant dans l'étage d'entrée et le second dans l'étage mélangeur à auto-amorçage. Les deux dispositifs sont à demi-ondes. Cependant, à partir du milieu de l'année 1964, un dispositif d'accord à quart d'onde sera fourni pour le marché français et sera également muni de deux transistors AF139.

EE 69 772 *pour plus amples renseignements*

ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 340 à 343

AMPLIFICATEUR MAGNÉTIQUE

Distributeurs: Auto-Electronics Ltd, Peel Grove, London, E.2

(Illustration à la page 340)

Réalisé par la société Serelec de Paris, l'appareil type AM-TH-PP, est un amplificateur magnétique push-pull à haute performance et d'une grande souplesse d'emploi pour la mesure des signaux de tension continue à niveau réduit de thermocouples, d'extensomètres, de divers ponts, de dispositifs à effet Hall, etc. Il peut être également utilisé comme préamplificateur stable dans des systèmes de régulation à cadre.

Il se distingue par sa bonne linéarité, sa stabilité et sa solidité, avantages auxquels il faut ajouter son encombrement réduit, son faible poids et son prix réduit. Il ne comporte pas de pièces mobiles et s'il est utilisé dans les limites prévues, il n'est pas sujet à perte d'efficacité. Sa durée est illimitée. Le secteur mort (qui constitue une défectuosité fréquente de ce genre d'amplificateur) a été éliminé et des signaux de 0 à 100 μV sont amplifiés aussi fidèlement que ceux de 0 à 100 mV. Il est, en outre, insensible à tout champ magnétique extérieur.

Il se caractérise enfin par une entrée dynamique et une impédance de sortie élevées, les résistances de la source et de la charge ayant un effet négligeable sur son gain et sa précision. On élimine ainsi les effets dûs aux variations de résistance des conducteurs ou des contacts. Il peut aussi être utilisé avec des thermocouples coaxiaux.

EE 69 773 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera le Stand No L502 à l'Olympia de Londres durant le Salon des Instruments, de l'Électronique et de l'Automation du 25 à 30 mai, et les visiteurs y seront les bienvenus.

CONDENSATEURS AU MICA ARGENTÉ

Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(Illustration à la page 340)

Deux versions standard différentes du condensateur au mica d'argent sont offertes maintenant par Plessey UK Ltd. Le modèle trempé dans la cire utilise une cire spéciale ne se fissurant pas et un revêtement de laque; il est prévu pour une gamme de température de $-25°$ C à $+70°$ C. Les modèles à encapsulage en résine d'époxyde assurent la protection contre l'humidité selon les normes H.2 et étendent la gamme de température de $-40°$ C à $+85°$ C.

Les condensateurs peuvent être fournis dans n'importe quelle valeur jusqu'à 1,0 μF à 200 V en service normal; 0,75 μF à 350 V; 0,5 μF à 500 V; et 0,05 μF à 750 V. Quatre formats de plaques sont utilisés dans la gamme standard, de 11,1 mm × 6,4 mm, à 25 mm × 13 mm, selon la capacité. Les terminaisons sont en fil de cuivre étamé (normalement radial mais axial sur demands), fixées de manière à assurer une bonne connexion électrique et une grande puissance mécanique. Des condensateurs au mica spéciaux pouvant être utilisés à des températures allant jusqu'à 250° C sont livrables sur commande.

EE 69 774 *pour plus amples renseignements*

MACHINE D'ÉQUILIBRAGE DYNAMIQUE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 340)

La nouvelle machine d'équilibrage dynamique type 1250C constitue un moyen simple et précis d'équilibrage d'éléments rotatifs de tous genres sur chaîne de montage.

Elle consiste en un bloc compact (mesurant 38,1 × 31,75 cm à la base, d'une hauteur de 38,1 cm et pesant 24,94 kg) pour banc d'essai. Un rail robuste, monté sur une console contenant tous les circuits, permet l'ajustage latéral de deux berceaux pouvant recevoir des pièces de 5 à 33 cm de long, pesant de 28 g à 4,5 kg. La pièce soumise à l'essai est entraînée, au moyen d'une courroie légère, par un moteur électrique incorporé, à des vitesses de 600 à 3 000 tours minute.

Les berceaux comportent des blocs en V montés de manière à avoir libre jeu dans la direction transversale. Les vibrations dues au déséquilibre dans les pièces soumises à l'essai sont transmises par les blocs en V à des bobines de captage dans un champ magnétique puissant. Une tension alternative proportionnelle à l'amplitude de vibration des pièces est produite dans les bobines. Cette tension est injectée, après amplification, à un appareil indicateur. Elle est également employée pour la mise en action d'un stroboscope à lumière blanche monté au dessus de la pièce. Une bande numérotée, ou collier, appliquée à la pièce est ainsi "figée" et indique, selon une aiguille réglable, s'il y a lieu d'ajouter ou de soutraire des poids (suivant la position d'un sélecteur) pour rétablir l'équilibre.

Pour l'équilibrage dynamique, le réglage de poids doit s'effectuer sur deux plans transversaux. Un interrupteur supplémentaire permet de choisir le berceau de droite ou de gauche pour une indication indépendante.

La machine type 1250C comporte un circuit à filtre variable pour le filtrage des vibrations extérieures. Chacun des deux berceaux comprend des circuits connexes d'étalonnage et de réglage de la sensibilité. Ces derniers sont préréglés pour le travail sur chaîne de montage, de manière à ce que même un ouvrier non spécialisé puisse déterminer le déséquilibre dynamique dans une pièce en 20 secondes environ, après quelques brèves indications.

Le stroboscope à lumière blanche à haute intensité permet d'utiliser la machine à la lumière du jour. Les paliers en V ouverts simplifient la charge et la décharge. Les deux paliers et le sélecteur obvient à la nécessité de renverser la pièce, accélérant ainsi le fonctionnement.

EE 69 775 *pour plus amples renseignements*

POTENTIOMÈTRE UNIVERSEL

Cambridge Instrument Co. Ltd. 13 Grosvenor Place, London, S.W.1

(Illustration à la page 341)

Ce nouvel instrument, qui remplace le potentiomètre d'atelier Cambridge type 44226, est un appareil compact et entièrement autonome principalement conçu pour le contrôle "sur place" et l'étalonnage de thermocouples ainsi que des indicateurs et enregistreurs connexes. Il se prête aussi à de nombreux contrôles de courant et de tension en laboratoire.

Ce nouveau potentiomètre a une seule gamme de 0,01 à 101 mV, le choix se faisant par commutateur à prises ou fil à contact glissant. La précision du réglage est de $\pm 0,1$ % sur le commutateur ou d'une demi-division sur le fil à contact glissant, suivane celle qui est supérieure. L'échelle de potentiel principal qui commande le commutateur à prises comporte 100 degrés de 1 mV et l'échelle de potentiel précis qui commande le fil à contact glissant a une gamme de 1 mV et elle est divisée en intervalles de 0,01 mV permettant l'évaluation à 0,002 mV près.

Le commutateur de fonction à cinq directions choisit l'un ou l'autre des deux circuits d'essai, qui sont principalement utilisés pour le contrôle des thermocouples d'essai ou standard. Une troisième direction relie l'alimentation en millivolts à réglage continu aux bornes d'essai pour la déviation des instruments à cadre mobile. La quatrième direction permet d'employer l'instrument comme source de potentiel étalonné pour le contrôle d'instruments potentiométriques.

L'instrument est entièrement autonome et comprend un galvanomètre incorporé, une cellule standard et des batteries sèches. Les bornes extérieures permettent de relier un galvanomètre plus sensible sans enlever le galvanomètre incorporé, de sorte que la gamme de l'instrument peut être portée à 500 V c.c. et à 10 A c.c. à l'aide de quelques accessoires élémentaires.

Le potentiomètre peut être normalisé à n'importe quel moment au cours des essais sans modifier les réglages de l'interrupteur de fonction ou d'échelle, c'est à dire en pressant simplement la clef de normalisation et en réglant un seul rhéostat.

EE 69 776 *pour plus amples renseignements*

RÉSISTANCES DE PRÉCISION

Electrothermal Engineering Ltd, 270 Neville Road, London, E.7

(Illustration à la page 341)

La société Electrothermal Engineering Ltd vient d'ajouter à sa série G 101 de résistances de précision bobinées en matière moulée deux nouveaux modèles à cosses.

Il s'agit des types 3505T de $\frac{1}{4}$ W et 3507T de $\frac{1}{2}$ W qui comprennent tous deux les dispositifs d'ancrage enrobés qui caractérisent toutes les résistances de la série G 101. Ils sont donc de construction robuste et assurés d'un facteur élevé de fiabilité et scellés à 100% contre l'humidité.

Les valeurs prévues vont de 0,1 Ω à 1 MΩ pour le type à $\frac{1}{4}$ W et de 0,1 Ω à 2 MΩ pour le type à $\frac{1}{2}$ W. Les tolérances vont de 0,1% à 0,01% selon la valeur, la stabilité étant supérieure à $\pm 0,02$%. Les tensions de service maxima sont de 550 V et 750 V respectivement.

EE 69 777 *pour plus amples renseignements*

RELAIS À LAME VIBRANTE

Erg Industrial Corporation, Luton Road Works, Dunstable, Bedfordshire

(Illustration à la page 341)

Un nouveau relais compact à lame vibrante, type AB.1618, spécialement conçu pour répondre aux besoins du contrôle des procédés industriels, des calculatrices et du matériel analogue, vient d'être réalisé par la Erg Industrial Corporation.

Le nouveau relais se distingue par son temps de commutation rapide—jusqu'à une milliseconde—allié à une durée de fonctionnement de quelques millions d'opérations.

Les contacts dorés sont hermétiquement scellés dans une atmosphère inerte et ils sont livrables soit à une seule commutation (unipolaire à deux directions) soit normalement ouverts (unipolaire à une direction). Le montage du commutateur à lame vibrante permet de les échanger tandis que le blindage du relais minimise les effets des champs magnétiques extérieurs.

La bobine du relais est à moulage de résine d'époxyde renforcée de fibre de verre, les conducteurs d'enroulement à moulage intégral étant spécialement conçus pour répondre à la méthode la plus récente de montage de circuits imprimés.

Des pattes de fixation, incorporées au blindage métallique, type AB.1652, peuvent être fournies sans supplément de prix.

EE 69 778 *pour plus amples renseignements*

IMPÉDANCEMÈTRE

Distributeors: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Illustration à la page 341)

L'impédancemètre, type GB11, fabriqué par la société danoise Radiometer, est distribué par Livingston Laboratories Ltd.

Cet appareil peut mesurer l'impédance, sous le rapport de la grandeur et l'angle de phase, de 1 Ω à 1,1 Ω et de 0° à ±90°. La précision de base est de ±1 % pour la grandeur et de ±0,5° pour l'angle de phase. Douze fréquences de contrôle-surprise sont assurées par le générateur interne et les sources extérieures peuvent être utilisées jusqu'à 1 MHZ. Les courants de contrôle peuvent être variés de 3,2 μA à 1 A. On peut effectuer la mesure de l'impédance flottante, à la masse ou équilibrée à la masse, ainsi que la polarisation du courant continu. L'appareil comprend des circuits transistorisés, un dispositif de lecture en ligne de l'amplitude et un instrument permettant la lecture directe de l'angle de phase.

Ses applications comprennent la mesure de l'impédance des réseaux de filtrage, des bobines, des haut-parleurs et des transformateurs ainsi que l'impédance différentielle des selfs de choc à noyau métallique et des diodes, et l'impédance effective des dispositifs non-linéaires à des courants de contrôle divers.

EE 69 779 *pour plus amples renseignements*

GÉNÉRATEUR TRÈS BASSE FRÉQUENCE

Distributeurs: Claude Lyons Ltd, Hoddesdon, Hertfordshire

(Illustration à la page 341)

Le générateur type GB 860, fabriqué par la société de Constructions Radio-électriques et Electroniques du Centre, fournit des ondes sinusoïdales, triangulaires ou carrées dans la gamme des très basses fréquences de 0,001 HZ à 1 kHZ avec une puissance de sortie de l'ordre de 0,5 W. Une sortie à phase variable est prévue, qui rend l'instrument apte à la mesure du déphasage. Il est particulièrement prévu pour l'étude des servomécanismes et des déformations et contraintes mécaniques, ainsi que pour l'étude des lignes de transmissions, des transformateurs pour fréquences infra-acoustiques, la recherche médicale et géophysique. Il fournit en outre des signaux précis de programmation.

Le circuit oscillateur comporte une bascule de Schmitt en un intégrateur Miller et il produit des ondes triangulaires ou rectangulaires symétriques ou des ondes carrées de fréquence et d'amplitude à variation continue, la gamme de fréquences étant couverte en six gammes à étalonnage linéaire. Le circuit correcteur de formes d'ondes à écrêteur à diode en cascade produit la sortie d'ondes sinusoïdales à partir de la forme d'onde triangulaire.

L'amplificateur de puissance de sortie transistorisé permet de fournir chacune de ces trois formes d'ondes à une faible impédance, équilibrée ou déséquilibrée.

L'impulsion de déclenchement positive obtenue par différentiation de la sortie de l'oscillateur à ondes carrées assure la synchronisation de la base de temps oscilloscopique. Une deuxième sortie d'impulsions, variable en phase sur 360°, permet de déterminer avec précision le déphasage. Lorsque cette sortie est utilisée avec un oscilloscope à deux traces, tel que le C.R.C. type OC728, l'angle de phase entre la sortie du générateur et celle du dispositif soumis à l'essai peut être mesuré à plus d'un degré près.

EE 69 780 *pour plus amples renseignements*

TRANSDUCTEUR CONVERTISSEUR

S.E. Laboratories (Engineering) Ltd, North Feltham, Middlesex

(Illustration à la page 342)

La société S.E. Laboratories (Engineering) Ltd vient de réaliser un système transducteur/convertisseur, type S.E. 905, à une ou deux voies; c'est un appareil autonome, portatif et à bas prix. Il est prévu pour l'indication et (ou) l'enregistrement d'informations fournies par des transducteurs inductifs de type extensométrique.

Il permet donc d'effectuer sur place des mesures de déplacement, d'accélération, de contrainte et de pression. Il se caractérise en particulier par son bloc d'alimentation à batterie incorporée rechargeable, permettant de l'utiliser sans avoir recours au courant secteur. Il contient un oscillateur stable de 3 kHz.

Deux types d'amplificateurs à tiroir interchangeable peuvent être fournis: le modèle S.E.423, principalement conçu pour les mesures statiques et dynamiques par transducteurs du type extensométrique ou par extensomètres, et le modèle S.E.449, utilisé pour les mesures statiques et dynamiques par transducteurs de type inductif et par transformateurs différentiels.

Un instrument de mesure incorporé peut être étalonné directement sur demande du client. Il y a également des sorties pour la mise en oeuvre d'enregistreurs à bande magnétique et galvanométriques et d'oscilloscopes.

EE 69 781 *pour plus amples renseignements*

BLOCS D'ALIMENTATION À FAIBLE TENSION

Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire

(Illustration à la page 342)

Le bloc d'alimentation stabilisée L.P.401 est le premier d'une nouvelle gamme de blocs de grande stabilité et de précision à bas prix dont le niveau de performance est comparable à celui d'instruments d'un prix beaucoup plus élevé.

Des versions de 0,5 A à 1 A et de 10 à 30 V peuvent être fournies avec sorties simples et doubles, leur performances étant semblables dans chacun des cas.

La tension, le courant et le calage de courant constant réglable sont contrôlés par le grand instrument de mesure à échelle de 9,52 cm sans recours à la commutation de gamme.

Les commandes de réglage de tension précis et approximatif facilitent l'ajustage vernier de la sortie, la commande de réglage précis ayant une gamme totale de ±0,75 V à tous les niveaux de sortie.

Un soin tout particulier a été apporté à la construction des circuits de commande de courant constant qui assurent la commande extrêmement sensible, stable et précise de la limitation de courant dans une gamme très étendue, garantissant ainsi la protection totale de matériels sensibles de tous types.

Tous les éléments de la série 400 peuvent résister à des courts-circuits prolongés de la sortie, sans dommage aucun.

EE 69 782 *pour plus amples renseignements*

DOUILLE DE CIRCUIT IMPRIMÉ

Oxley Developments Co. Ltd, Priory Park, Ulverston, Lancashire

(Illustration à la page 342)

La douille de circuit imprimé Oxley type 50S/PCB a été conçue pour le montage direct sur plaquette de circuit imprimé modulaire de 0,1 ″ avec trous d'un diamètre minimum de 0,050 ″. Le montage est effectué de manière à ce que les fiches puissent être insérées sur un plan parallèle à la plaquette de circuit imprimé, permettant ainsi l'empilage serré de plaquettes dans un équipement. La tension de claquage est de 5 kV c.c. et la capacité ajoutée au circuit est inférieure à 0,7 pF par sortie. La hauteur totale au dessus de la plaquette de circuit imprimé est de 0,63 cm. Des douilles de 1 à 12 directions peuvent être fournies avec codage de couleur. Dans ce cas, le nombre de sorties doit être indiqué ainsi que la disposition des couleurs de gauche à droite. Les couleurs prévues sont: noir, brun, rouge, orange, jaune, vert, bleu, violet, rose, gris et blanc.

EE 69 783 *pour plus amples renseignements*

COMPTEUR DE LOTS

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(*Illustration à la page 342*)

La société Advance Controls Ltd a réalisé un nouveau compteur de lots à minuterie: le type TB 42.

Cet instrument peut être utilisé pour minuter un procédé ou mesurer un lot continûment ou indépendamment. Des vitesses de minutage de 0,01 sec peuvent être atteintes et le comptage des lots peut être effectué au rythme maximum de 1000 lots par seconde. L'appareil est à réenclenchement instantané, éliminant ainsi toute perte de comptage ou de temps sur le cycle à répétition. Cette dernière peut être à cyclage automatique, par commande à distance ou par bouton-poussoir manuel.

Le TB 42 comporte des tubes cathodiques qui assurent le contrôle visuel de la précision du comptage. Il comprend en outre un dispositif de signalisation à 99 coups ou 0,99 sec avant la fin d'un comptage préréglé.

Le relais de sortie peut être réglé de manière à effectuer un réenclenchement automatique après une période de temps déterminée. En variante, le réenclenchement peut se faire par commande extérieure.

L'appareil mesure 22,86 cm de large sur 16,51 cm de haut et 13,97 cm de profondeur. Il peut être fourni avec une gamme complète de capteurs (photoélectriques, magnétiques, à commutation à lame et de proximité).

EE 69 784 *pour plus amples renseignements*

MINUTERIE ÉLECTRONIQUE

Pioneer Designs Ltd, Crown House, Walton-on-Thames, Surrey

(*Illustration à la page 343*)

Cette minuterie comprend un circuit de déclenchement à impulsions à sensibilité variable par commande individuelle sur le panneau frontal. La sensibilité peut être contrôlée de manière à ce que les tensions continues et alternatives appliquées à partir d'une source extérieure au circuit de déclenchement d'entrée a impédance élevée puissent mettre en oeuvre le cycle minuté. En variante, on peut augmenter la sensibilité jusqu'au point où s'effectue le cyclage automatique.

Deux gammes de temps sont incorporées dans ce modèle: 0.2 à 1.0 et 1 à 5 secs. La période de minutage est à variation infinie et réglée sur un cadran de 12,7 cm.

On peut, en outre, déclencher le cycle de minutage au moyen d'un bouton-poussoir sur le panneau frontal ou par court-circuit, c'est à dire, par exemple, à l'aide d'un microcommutateur, entre les deux bornes.

La sortie se fait par un relais, faisant fonction de commutateur électrotechnique unipolaire. Des bornes d'ouverture et de fermeture sont prévues pour un circuit extérieur à fusion indépendante de 2 A. En changeant la liaison de connexion intérieure, on peut faire fournir à la minuterie des tensions de 230 V, 50 Hz, allant jusqu'à 500 W aux bornes de sortie. La consommation de la minuterie est inférieure à 8 W.

EE 69 785 *pour plus amples renseignements*

CONDENSATEURS SANS DÉCHARGE

F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(*Illustration à la page 343*)

La société F. C. Robinson and Partners Ltd vient de réaliser une gamme complète de condensateurs sans décharge d'ionisation.

Lorsqu'on effectue des essais de décharge d'ionisation avec n'importe quel détecteur, on n'obtient une sensibilité maxima que lorsque la valeur du condensateur de blocage est élevée par rapport à celle du spécimen soumis à l'essai. Tenant donc compte de ce fait, des valeurs peuvent être produites de manière standard, de:

1000 pF à 0,05 μF, à 50 kV efficaces
2000 pF à 0,1 μF, à 25 kV efficaces
5000 pF à 0,5 μF, à 10 kV efficaces
et 0,01 μf à 2 μF, à 5 kV efficaces.

Des tensions de régime plus élevées ou des valeurs de capacité plus importantes peuvent être obtenues en reliant certains éléments en série ou en parallèle. Notre photographie montre un banc parallèle de modèles de 0,05 μF, 50 kV.

Certains types sont de construction tubulaire ordinaire, tandis que d'autres sont incorporés à des réservoirs d'acier remplis d'huile ou à des conteneurs rigides en chlorure polyvinylique. Tous les modèles sont soigneusement contrôlés à tous les stades de fabrication de manière à s'assurer qu'ils sont sans décharge.

EE 69 786 *pour plus amples renseignements*

SÉLECTEUR DE LIMITE NUMÉRIQUE

Digital Measurements Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(*Illustration à la page 343*)

La société Digital Measurements Ltd vient d'annoncer la mise au point d'un sélecteur de limite numérique, type DM5080, et d'un élément logique, DM5006, qui forment un ensemble avertisseur pour voltmètres numériques, enregisteurs de données numériques, compteurs, etc.

L'élément logique peut recevoir jusqu'à trois décades décimales et actionner jusqu'à 16 sélecteurs de limite. Les sélecteurs "de limite élevée" produisent un signal lorsque l'entrée au système avertisseur est égale ou supérieure au niveau choisi. Les sélecteurs de "basse limite" produisent un signal lorsque l'entrée au système est inférieure au niveau choisi. Les sélecteurs peuvent être aisément convertis de la "limite élevée" à la "limite basse" et réciproquement par une modification des connexions de liaison.

La sortie peut être utilisée pour mettre en oeuvre une action automatique de rétablissement ou pour avertir l'opérateur d'une installation losque les conditions sont "hors limites" au moyen de panneaux éclairés, de signaux d'alarme audibles, etc. En variante, lorsqu'on enregistre les entrées provenant de sources différentes à l'aide d'un système de concentration, la touche de couleur de la machine à écrire ou de la machine à imprimer peut être mise en action automatiquement.

Les éléments sont de conception modulaire; un cadre de bâti de 48,2 cm, d'une hauteur de 17,78 cm, peut recevoir soit huit sélecteurs soit l'élément logique qui fonctionne sur courant alternatif de 100 à 125 V ou de 200 à 250 V.

EE 69 787 *pour plus amples renseignements*

ELÉMENTS NUMÉRIQUES ANALOGIQUES

Trainer and Simulator Division, Elliott Brothers (London) Ltd, Chobham Road, Frimley, Nr. Aldershot, Hampshire

Un système simulateur ou de calcul analogique/numérique de n'importe quelle configuration voulue peut être construit grâce à une nouvelle gamme d'éléments standard mis au point par la Trainer and Simulator Division de la société Elliott-Automation Ltd. Ces éléments qui sont le résultat de travaux effectués sur un simulateur d'un modèle perfectionné permettent au réalisateur la construction d'appareils d'une précision et d'une performance déterminés à un prix minimum. Le champ d'applications est extrêmement vaste et comprend, entre autres, le contrôle des systèmes analogiques et numériques, l'enseignement de la théorie de contrôle technique et la construction de simulateurs d'instruction ou de réalisation.

Les éléments de base, pouvant tous être fournis séparément, comprennent une plaquette d'asservissement électronique, comportant à son tour un modulateur-préamplificateur combiné et un amplificateur de puissance pour l'entraînement de moteurs d'asservissement standard, une boîte de vitesses de grande précision à sept arbres de sortie un amplificateur de puissance pour le fonctionnement de composants asservis dans la gamme de dimensions 08 à 15 et une plaquette de démodulateur pour la conversion des signaux de 400 Hz et des amplificateurs universels de calculatrices à courant alternatif ou continu. Tous ces éléments, ainsi que leurs blocs d'alimentation, peuvent être logés dans un système de châssis modulaire.

Ces nouveaux éléments analogiques Elliot peuvent être combinés de façon à former une gamme étendue de systèmes d'asservissement, tels qu'un système d'asservissement de position à courant continu ou un système de vélocité à entrées à courant continu ou à 400 Hz c.a., ainsi que des systèmes de résolution ou de synchronisation. Les sorties peuvent être sous forme de rotations d'arbre ou de signaux provenant de potentiomètres, de synchros, de résolveurs et d'encodeurs à arbre. L'angle d'arbre peut être affiché sur le panneau frontal du système à l'aide d'accouplements des arbres de la boîte de vitesses.

EE 69 788 *pour plus amples renseignements*

CONTRÔLEUR DE SOUDABILITÉ

Multicore Solders Ltd, Maylands Avenue, Hemel Hempstead, Hertfordshire

(*Illustration à la page 343*)

La demande de taux de production accélérés dans l'assemblage du matériel électronique exige une méthode simple et précise pour déterminer les propriétés de soudabilité des fils d'extrémité des composants.

La société Multicore Solders Ltd, en collaboration avec la Electronic Engineering Association, a réalisé une machine qui serait la seule du genre pouvant contrôler la soudabilité des extrémités des fils.

La machine à contrôler la soudabilité pose lentement un échantillon de fil métallique, dans lequel on a fait passer un flux, sur quelques gouttes de soudure fondue. Le temps en secondes mis par la soudure pour couler autour du fil et s'unir au dessus indique le degré de soudabilité du fil. Un bloc d'alimentation incorporé alimente le thermostat qui contrôle l'élément de chauffe, la température étant indiquée par un pyromètre à batterie d'une précision de 1 %.

La machine à contrôler la soudabilité est fournie complète avec pyromètre, chronomètre étalonné à 1/100^e de seconde et flux liquide, avec emballage d'exportation.

EE 69 789 *pour plus amples renseignements*

Résumés des Principaux Articles

Un instrument de mesure du taux d'impulsions à temps de réponse réduit et son application dans la cardiotachymétrie

par P. A. Tove et J. Czekajewski

Résumé de l'article aux pages 290 à 295

On a réalisé une méthode permettant de mesurer le taux de répétition des impulsions avec réponse rapide aux changements de fréquence. Le principe consiste à mesurer les intervalles de temps entre les impulsions en chargeant alternativement deux condensateurs durant les intervalles successifs d'impulsions. La tension sur le condensateur qui n'est pas chargé est indiquée sur un instrument de mesure ou un oscilloscope et représente la mesure de la dernière période. L'indication directe sur l'échelle de la longueur de l'intervalle d'impulsion ou de la fréquence est effectuée et la possibilité d'obtenir des échelles de fréquence linéaires et logarithmiques est examinée. Il est montré que la simple charge exponentielle des condensateurs correspond à peu près aux formes d'échelle désirées dans des régions de fréquence limitée. Des méthodes sont décrites qui permettent d'obtenir des échelles linéaires et logarithmiques dans des zones étendues. Des circuits pratiques sont indiqués pour les applications médicales dans la cardiotachymétrie et l'enregistrement de la fréquence respiratoire lorsqu'il est indispensable d'avoir une réponse rapide aux changements du taux d'impulsions. Une utilisation supplémentaire réside dans les systèmes de transmission d'informations basés sur la modulation de la longueur d'impulsion.

Phénomènes d'échantillonnage de télémesure

par S. Poole

Résumé de l'article aux pages 296 à 299

Cet article traite de divers aspects de l'échantillonnage de télémesure, dont la largeur de bande et le balayage de recherche. Il s'agit en général de cas où l'analyse visuelle des formes d'onde obtenues est exigée. Après une étude théorique des phénomènes, quelques résultats de travaux de laboratoire sont indiqués.

Un circuit de mesure de la tension efficace réelle par J. C. Cluley

Résumé de l'article aux pages 300 à 303

Le circuit redresseur classique à lecture de moyenne utilisé dans les régulateurs de tension automatiques pour alternateurs peut provoquer des erreurs considérables dans la tension de sortie si la forme d'onde mécanique comporte des composants harmoniques. L'étendue de telles erreurs et leur assujettissement à l'amplitude et à la fréquence de l'harmonique sont analysés dans cet article qui pécrit également un circuit pouvant produire une tension continue exactement proportionnelle à la tension d'alternateur efficace. Le circuit prévient l'écart de temps propre aux dispositifs mécaniques ou aux éléments sensibles à la température et il comprend une diode diélectrique au sulphure de cadmium, de conception récente et possédant une caractéristique quadratique.

Amplificateur transistorisé à courant continu de faible dérive par J. I. Pinto

Résumé de l'article aux pages 304 à 308

Cet article décrit les variations des paramètres de courant continu de transistors, qui sont importants dans la réalisation des amplificateurs c.c. à faible dérive. L'article analyse la "paire à longue queue" classique et les modifications nécessaires à l'amélioration de la performance et évalue les résultats obtenus. Il décrit ensuite un nouvel amplificateur équilibré à "image-miroir" et examine sa performance et ses possibilités.

Circuits amplificateurs hybrides à impédance d'entrée élevée et à basse tension par F. H. Laishley

Résumé de l'article aux pages 309 à 311

Un groupe de circuits amplificateurs hybrides accouples à la tension continue sont décrits. Ces circuits fonctionnent sur alimentation en tension basse et sont à impédance d'éntrée élevée.

La technique employée consiste à faire usage d'une seule lampe exigeant une faible tension anodique dans l'étage d'entrée de chacun des circuits et de faire suivre cet étage par des circuits semi-conducteurs classiques. On obtient ainsi des circuits simples à haute impédance, de dimensions réduites et à faible consommation életrique.

Ces circuits ont été conçus à l'origine pour la mesure de tensions emmagasinées dans des condensateurs à faible valeur.

Détecteurs accordés à circuit T jumelé par T. Spencer

Résumé de l'article aux pages 315 à 321

Cet article traite d'une utilisation particulière du circuit à T jumelé dans les amplificateurs à transistors. Le circuit T jumelé est utilisé comme élément de réaction shunt, amplifiant une bande étroite de fréquences centrées autour de la fréquence d'encoche du T jumelé, avec un étage Q^F d'environ 5 à 10 cycles; en montant en cascade deux de ces étages on parabolise le Q^F. Un résumé de la théorie complète du T jumelé par l'emploi d'une source finie et d'impédances de charge est donné en appendice. Un nomogramme est indiqué, en outre, pour une réalisation optimale de ce réseau. Le gain d'étage de l'amplificateur est démontré, ainsi que le fait que le Q_F d'un amplificateur accordé, avec circuit à T jumelé dans la course de réaction, est directement proportionnel à la transmission de courant continu du circuit, cette dernière étant son Q intrinsèque. Ainsi les étages montés en cascade qui ne sont pas accordés individuellement, à moins qu'on ne veuille un effet de bande passante, et en n'utilisant qu'un seul circuit à T jumelé dans la course de réaction globale, sont préférables. L'accord est très simple, la stabilité extrêmement bonne et presque n'importe quelle largeur de bande est réalisable. Des courbes universelles sont données afin de faciliter l'étude d'un amplificateur accordé. Il est intéressant de relever qu'on s'écarte très peu de l'étude des circuits à lampes.

Une méthode simple de mesure des courbes de rayonnement d'antennes basée sur l'emploi d'un ballon
par C. S. L. Keay et R. E. Gray

Résumé de l'article aux pages 322 à 325

Les courbes de rayonnement des antennes fonctionnant à des fréquences de 20Mc à 500Mc peuvent être mesurées facilement par l'emploi d'un petit émetteur autonome et d'un dipôle suspendus à un ballon météorologique. Le ballon est tenu et commandé par quarte lignes de nylon qui le forcent à se déplacer dans un arc à rayon constant centré sur le système d'antenne soumis au contrôle. Un diagramme complet, précis à 10% près, peut être obtenu très rapidement, parfois même en moins d'une heure, selon la fréquence de travail et les détails exigés.

Analyse d'un amplificateur de réaction par C. D. West

Résumé de l'article aux pages 326 à 329

L'auteur analyse le circuit d'un amplificateur d'impulsions dont le coefficient de gain est à température réduite et qui se compose d'un amplificateur de différence accouplé à la tension continue et d'un étage émetteur commun.

Il indique les valeurs de composants d'un amplificateur à gain de 10 et à coefficient de température de 50^{10-6}°C. Plusieurs étages peuvent être montés en cascade.

Un stabilisateur série stable et de grande fiabilité par D. J. Collins

Résumé de l'article aux pages 330 à 331

Les applications nécessitant des tensions continues stables offrent au technicien de circuits un choix étendu de montages possibles. Ce choix est généralement dicté par les paramètres fondamentaux de puissance, la qualité du rendement et le coût à prévoir. Si la plupart de ces conditions peuvent être aisément remplies, il se produit, néanmoins, des cas où se pose un problème particulier auquel il faut apporter une solution appropriée. Cet article décrit un circuit spécialement conçu pour pouvoir constituer une source de tension stable, compacte et peu coûteuse pour amorcer les transducteurs de pression extensométriques.

DIE FRANZÖSISCHE BAUELEMENTE-AUSSTELLUNG

Beschreibung einiger der auf der Internationalen Ausstellung für Elektronische Bauelemente in Paris von 7.—12. Februar 1964 ausgestellten französischen Bauelemente nach Angaben der Hersteller

Übersetzung der Seiten 336 bis 339

COGIE

3 et 5, boulevard Anatole-France, Aubervilliers (Seine)

THERMOELEKTRISCHE KÜHL-BAUGRUPPEN

(Abbildung Seite 336)

Die thermoelektrischen Baugruppen "Fricogie" FC8-7A beruhen auf dem Peltier-Effekt und bestehen aus acht Wismuttellurid-Elementen, die in Serie geschaltet und mit thermischer Isolation zusammengebaut sind. Die maximale Wärmeentziehung ist 32 W.

Durch Umpolen der Gleichstromversorgung werden die warmen und kalten Flächen umgekehrt, so dass sich ein ausgezeichnetes Mittel zur Temperaturregelung ergibt. Der Höchstunterschied zwischen warmen und kalten Flächen ist 75° C, wenn die warme Fläche eine Temperatur von 25° C hat.

Kleinere Ausführungen dieser Baugruppen sind auch lieferbar.

EE 69 751 *für weitere Einzelheiten*

C.O.P.R.I.M.

(Compagnie des Produits Élementaines pour Industries Moderues)

7, Passage Charles-Dallery, Paris 11e

MINIATURPOTENTIOMETER

(Abbildung Seite 336)

Das neue Modell EO86 gehört zum Fertigungsprogramm für Kohleschicht-Potentiometer und ist für gedruckte Schaltungen mit 2,54 mm Rastermass bestimmt.

Es ist kleiner als Modell 4EO97, trotzdem Aussehen und Konstruktion ähnlich sind; das Modell EO86 ist bei 40° C mit 0,1 W belastbar.

Diese Bauelemente sind besonders dort für Verwendung geeignet, wo der Raum knapp und eine niedrigere Belastbarkeit zulässig ist.

Sie sind in Werten von 100Ω bis zu 20 kΩ, im Sonderauftrag bis zu 500 kΩ lieferbar.

EE 69 752 *für weitere Einzelheiten*

ELNO

(Etabl. Lailler-Pequet et Cie)

18 et 20, rue du Val-Notre-Dame, Argenteuil (Seine-et-Oise)

LEICHTE KOPFHÖRERAUSRÜSTUNG

Diese Kopfhörer-Mikrofonkombination wiegt weniger als 150 g und besteht aus einem mit Polyesterschaum umhüllten Stahlbügel, an dessen einem Ende ein leichter Hörer befestigt ist. Die Vorderseite des Hörers ist mit einem Geräuschschutz versehen, auf der Rückseite ist der Mikrofonbügel befestigt, der sowohl senkrechte wie seitliche Verstellung erlaubt. Die Leitungen vom Hörer und Mikrofon gehen zu einem Verteilerkästchen, das an der Kleidung des Benutzers befestigt wird.

Der Frequenzgang des Hörers ist ±3 dB von 20 Hz ... 4 kHz; er ist in sechs Standard-Impedanzen von 15Ω bis zu 2 kΩ lieferbar.

Das Mikrofon hat eine Empfindlichkeit von 0,1 mV/μB. Der Frequenzgang ist +6 dB/Oktave bis zu 1,5 kHz und dann linear bis zu 4 kHz. Es ist in sechs Standard-Impedanzen von 15Ω bis zu 2 kΩ lieferbar.

EE 69 753 *für weitere Einzelheiten*

KOPFKISSENLAUTSPRECHER

(Abbildung Seite 336)

Dieser Lautsprecher ist für Krankenhauspatienten bestimmt, die ans Bett gebunden sind. Er besteht aus einem kreisrunden Gehäuse mit elliptischem Querschnitt und Öffnungen für die Klangzerstreuung. Im Gehäuse ist ein elektromagnetischer Hörer eingebaut, der in zehn Standard-Impedanzen von 15Ω bis zu 18 1Ω lieferbar ist.

EE 69 754 *für weitere Einzelheiten*

FERISOL

18, avenue Paul-Valliant-Couturier, Trappes (Seine et Oise)

KOAXIALE MESSLEITUNGEN

Die Messleitungen UB200 und UA200 sind geschlitzte Koaxialleitungen mit Impedanzen von 50Ω für Type UB200 und 100Ω für Type UA200.

Sie sind mit identischen Schlitten ausgerüstet, von denen ein Koppelstift mit veränderlicher Tiefe in die Leitung hineinragt und die HF-Spannung abtastet. Der Abtaster ist aperiodisch und hat keine Abstimmeinrichtung. Die HF-Spannung kann über einen N-Stecker abgenommen werden. Auf dem Schlitten ist ausserdem ein koaxialer Kristalldetektor für modulierte oder unmodulierte HF-Spannungen vorgesehen. Das Stehwellenverhältnis für die zu untersuchende Impedanz lässt sich daher direkt durch Anschluss des Messgerätes AG101 für Stehwellenverhältnisse messen.

Die mechanische Verschiebung des Schlittens auf der Messleitung kann man auf einer in Zentimeter und Millimeter unterteilten Skala mit einem für Zehntel Millimeter unterteilten Nonius genau ablesen. Die erreichbare Genauigkeit ist in der Grössenordnung von 0,1 mm. Über einen kleineren Bereich (maximal 50 mm Weg) kann mit Hilfe eines mechanischen Vergleichers, für den eine Befestigungsvorrichtung vorhanden ist, eine Ablesegenauigkeit von 0,01 mm erzielt werden.

EE 69 755 *für weitere Einzelheiten*

ELEKTRONISCHER MEGOHMMESSER

Der Megohmmesser RM200 ist für Widerstandsmessungen zwischen 0,2 MΩ und 10^8 MΩ ausgelegt.

Durch eine neue Messtechnik für sehr niedrige Gleichspannungen kann man den unbekannten Widerstand unmittelbar auf der Skala eines Galvanometers anzeigen. Im Gegensatz zu herkömmlichen Geräten wie z.B. der Wheatstoneschen Brücke, bei der der Benutzer bei jeder Änderung der Unbekannten die Brücke wieder abgleichen muss, zeigt das neue Messgerät RM200 diesen Wert ohne irgendeinen externen Eingriff an. Dadurch wird es nicht nur möglich, den momentanen Wert der Unbekannten zu untersuchen, sondern auch ihr Verhalten, wenn sie anderen Parametern wie Temperatur, Feuchtigkeit usw. ausgesetzt wird. Für Verwendung dieser Art gibt ein Ausgang für Registriergeräte eine Gleichstromkomponente ab, deren Wert durch eine einfache Funktion mit dem Wert der Unbekannten verbunden ist.

Die zahlreichen im Gerät zur Verfügung stehenden Prüfspannungen erlauben Einsatz als Universalmessgerät für die neuste Technik. So ermöglicht z.B. die Prüfspannung von 10 V Untersuchung von Transistoren.

Ausserdem ist das Gerät so konstruiert, dass externe Prüfspannungen jeder Höhe benutzt werden können. Eine einfache Berechnung gibt den wahren Wert der Unbekannten mit Hilfe des an der Galvanometerskala angezeigten Wertes und der angelegten Prüfspannung.

EE 69 756 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Lancashire, England

Parametrische Verstärker

(Abbildung Seite 337)

Der parametrische Verstärker VCA/L10 für 570 MHz hat bei 6 MHz Bandbreite eine Verstärkung von 16 dB (auf die 1-dB-Punkte bezogen) und einen Gesamtrauschfaktor von unter 1 dB. Es ist interessant, dass dieser Rauschfaktor wesentlich unter dem bei dieser Frequenz empfangenen Weltraumrauschen liegt.

570 MHz ist die Frequenz des neuen 625-Zeilen-Fernsehprogramms der British Broadcasting Corporation, und der neue Verstärker wurde speziell im Zusammenhang mit diesen Sendungen entwickelt. Rediffusion Research Ltd hat für diesen Verstärker einen Auftrag erteilt und wird sie einsetzen, um den Teilnehmern ihres Drahtfernsehdienstes in Southampton den Empfang des neuen Programms zu ermöglichen.

Weiterhin wurde der parametrische Verstärker VCA/S30 gezeigt, der für Betrieb bei einer Mittenfrequenz von 3 GHz ausgelegt ist. Bei einer Bandbreite von 70 MHz und auf die 3-dB-Punkte bezogen hat er eine Verstärkung von 20 dB und einen Rauschfaktor von 2,7 dB. Das Hauptverwendungsgebiet für parametrische Verstärker, die für Betrieb im S-Band konstruiert sind, sind Radarempfänger, deren effektive Reichweite mindestens um einen Faktor von 3 erhöht wird.

EE 69 757 *für weitere Einzelheiten*

Sperrschicht-Photozelle

(Abbildung Seite 337)

Eine neue grossflächige Silizium-Sperrschicht-Photozelle MS40 wurde ausgestellt, die hauptsächlich als sehr leistungsfähiges und wirtschaftliches Mittel zur Umwandlung von Sonnenenergie in elektrischen Strom gedacht ist, der sich besonders zum Laden von Batterien für Pufferspeichersysteme eignet. Die MS40 besteht aus vier parallelgeschalteten Einzelzellen auf einer gemeinsamen Kühlplatte. Ein integraler Teil der gepressten Umhüllung ist eine zylindrische Linse, die die effektive Sammelfläche für Sonnenstrahlungen erhöht. Dadurch wird zusammen mit anderen Konstruktionsmerkmalen in Bodenhöhe ein Umwandlungswirkungsgrad von 8% für das Sonnenspektrum erreicht.

Bei Beleuchtung durch eine Wolframquelle bei 2854°K, senkrechtem Einfall und einer Beleuchtungsstärke von 3000 lm/Fuss^2 (rund 280 Lux) gibt die neue Zelle bei 25°C Umgebungstemperatur einen Mindeststrom von 500 mA an 0,6 Ω Abschluss ab. Bei der Konstruktion wurde der starken Korrosion und mechanischen Umgebungsbeanspruchung, denen aus MS40-Moduln aufgebaute Sonnenbatteriefelder ausgesetzt werden, besondere Aufmerksamkeit gewidmet. Fassungen und Verbindungsstützpunkte sind aus der Monellegierung 400 gefertigt, deren Einstufung für Korrosionswiderstand für Verwendung in direkter Seeumgebung ausreicht; die robuste Epoxydharzumhüllung kann Betrieb und Lagerung in Umgebungstemperaturen von −20° ... +85°C aushalten.

Ein versuchsmässiger Zusammenbau von MS40-Moduln wurde vor kurzem nach dem Persischen Golf verschifft, wo er an ortsfesten Marinekonstruktionen Tests untergehen wird, um weitere Daten für die Vervollständigung der Erprobung zu gewinnen. Diese endgültigen Erprobungsversuche werden zur Bestimmung der optimalen Konstruktion grosser Sonnenkraftwerke zur Versorgung abgelegener Gegenden vorgenommen.

EE 69 758 *für weitere Einzelheiten*

HEWLETT-PACKARD S.A.

54 Rte des Acasias, Genf, Schweiz

Mikrowellen-Frequenzwobbler

(Abbildung Seite 337)

Die Frequenzwobbler 691A/B (1 ... 2 GHz) und (2 ... 5 GHz) haben drei Durchlauf-Betriebsarten. In der ersten, sogenannten "Normal"-Betriebsart wird die Ausgangsfrequenz in beiden Richtungen zwischen zwei kontinuierlich verstellbaren Grenzfrequenzen gewobbelt. In der Betriebsart "F" wird die Frequenz um eine verstellbare Mittenfrequenz gewobbelt, und ausserdem haben diese Durchläufe zwei unabhängig voneinander verstellbare Frequenzmarken, die digital mit einer Unsicherheit von ±0,05% der Ausgangsfrequenz einstellbar sind. Diese Marken amplitudenmodulieren den Ausgang sowohl in der Normal-Betriebsart wie auch bei F-Betrieb—eine für X-Y-Aufzeichnungen oder Beobachtung der Ablenkung auf einem Oszillografen sehr wichtige Einrichtung. In der dritten Betriebsart werden dieselben Frequenzmarken dazu benutzt, eine der Normal-Betriebsart ähnliche sogenannte "Marken-Wobbelung" herbeizuführen, in der die Marken den Wobbelbereich bestimmen. Wahl der Betriebsart ermöglicht es z.B., den ganzen Durchlassbereich eines Filters in "Normal" zu durchlaufen und das Sperrgebiet desselben Filters durch Umschalten auf "Marken-Wobbelung" ohne Änderung der Grenzfrequenzen zu untersuchen.

Im 691B und 692B werden Diodenstreifen-Leiterabschnitte für Modulation und Dämpfung benutzt. Die mit diesen Stiftdioden vorgenommene Amplitudenmodulation ist vollkommen unabhängig von der Rückwärtswellen-Oszillatorröhre, so dass Mitziehen eliminiert wird. Diese Stiftdioden-Netzwerke verwendet man auch, um das erzeugte Ausgangssignal innerhalb 0,3 dB zu halten.

EE 69 759 *für weitere Einzelheiten*

Normalfrequenzempfänger

(Abbildung Seite 337)

Der Empfänger 5090A ist für Empfang des Senders Droitwich auf 200 kHz bestimmt. Da die Trägerfrequenzkonstanz von Droitwich innerhalb 5×10^{-10} pro Tag ist, erhält man auf diese Weise ohne Schwierigkeiten und mit geringem Aufwand ein sehr genaues Frequenznormal für Labors.

Der 5090A hat Ausgangsfrequenzen von 100 kHz und 1 MHz.

EE 69 760 *für weitere Einzelheiten*

I.E.R.

(Impression Enregistrement des Resultats)

6, rue Blondel, Courbevoie (Seine)

Streifendrucker

• *(Abbildung Seite 337)*

Der schnelle Digitalstreifendrucker 205 mit Paralleleingabe kann binärverschlüsselte Dezimaleingaben aufnehmen, entschlüsseln und mit einer Höchstgeschwindigkeit von zehn Zeilen je Sekunde ausdrucken. Die Höchstbreite ist 13 Stellenwerte und die Höchstgeschwindigkeit daher 130 Zeichen/Sekunde.

Die Ausrüstung ist eine Weiterentwicklung der Type 203, die mit sechs Zeilen/Sekunde arbeitet.

EE 69 761 *für weitere Einzelheiten*

LIE-BELIN

296, avenue Napoléon-Bonaparte, Rueil-Malmaison, (Seine-et-Oise)

Hochspannungsstromversorgung

(Abbildung Seite 338)

Diese volltransistorisierte Hochspannungsstromversorgung ist entweder als Standard-Esone-Baugruppe 2M oder in Gehäuseausführung mit Traggriff lieferbar.

Die technischen Hauptdaten sind wie folgt:

Leistung: 2 kV, 2 mA.
Konstanz: 5×10^{-5} für ±10% Netzschwankungen.
Innenwiderstand: 50 kΩ.
Ausgangsspannung: in 80-V-Stufen einstellbar und mittels 10gängigem Wendelpotentiometer feingeregelt.
Kontrolle: Messgerät auf Frontplatte.
Netzanschluss: 127 ... 220 V~ 50 Hz oder 24 V=.

EE 69 762 *für weitere Einzelheiten*

LUFTÜBERWACHUNGSANLAGE

(*Abbildung Seite 338*)

Die Ausrüstung ist für das automatische oder manuelle Messen der α- und β-Aktivität von in Filter gesaugten Staubproben ausgelegt.

Das zu untersuchende Filter wird vor einen Photovervielfacher gebracht, dem ein elektronischer Zähler komplett mit Anzeige und Druckwerk nachgeschaltet ist.

Für den α-Nachweis wird ein getrennter Szintillationskopf (ZnS) benutzt, der gegen jedes Filter in seinem Halter gehalten wird. Für den β-Nachweis wird ein mit dem Photovervielfacher zusammengebauter Einzelszintillator verwendet.

Die Ausrüstung besteht aus einem in zwei Fächer unterteilten Gehäuse. Im oberen Fach befinden sich der Detektor (54 AVP) mit Vorverstärker, Verstärker, Former und—im Bedarfsfall—eine Bleikammer (β-Nachweis); Filterhalterkassette, Kassettenvorschub, Einrichtung zur Selektion der Filter aus der Kassette sowie für ihren Vorschub und Rücklauf, Einrichtungen zum Aufdrucken der Ergebnisse auf jeden Filterhalter und eine Papierstreifenrolle; ausserdem elektromechanische Teile mit Programmierer, Übertragungs- und Schutzeinrichtungen.

Im unteren Fach ist ein Standard-Chassis 4U3 mit den elektronischen und Stromversorgungs-Systemen mit fünf Zähldekaden und deren Anzeige, sowie die Druckeinrichtung mit mechanischen Einzeldekaden untergebracht.

Die Kassette hält 60 Filter von 160 mm × 160 mm Grösse.

EE 69 763 *für weitere Einzelheiten*

METRIX

(Compagnie Generale de Metrologie)

Chemin de la Croix-Rouge, Annecy, (Haute-Savoie)

TRANSISTOR-TESTER

(*Abbildung Seite 338*)

Der Transistor-Tester 302A wurde für die Prüfung von Dioden und Gleichrichtern, Zenerdioden und Transistoren von Kleinsignal- bis zu Leistungstypen entwickelt.

Dioden werden durch Messen des Sperrstroms mittels eines hochempfindlichen Galvanometers geprüft. Ein Serienwiderstand in der Schaltung schützt das Galvanometer gegen Beschädigung durch eine fehlerhafte Verbindung. Während der Prüfung wird die Diode mit einer 4 V nicht überschreitenden Spannung beaufschlagt.

Bei der Zenerdiodenprüfung liegt die Diode in Serie mit einer 18-V-Stromquelle und einem strombegrenzenden Widerstand.

Das Messgerät, dessen Stromverbrauch vernachlässigbar klein ist, wird mit den Diodenanschlüssen verbunden und die durch die Diode geregelte Spannung gemessen.

Die Verlustleistung der Diode ist äusserst niedrig (ungefähr 30 mW).

Für Transistor-Tests wird I_{CO} auf dieselbe Weise gemessen wie der Dioden-Sperrstrom.

Die Schaltungen zum Messen der statischen Verstärkung h_{21E} sind in diesem Gerät so ausgelegt, dass man den Kollektorstrom auf 1 mA, 10 mA, 100 mA oder 1 A einstellen kann.

Beim Ausführen dieser Prüfung ist die Kollektor-Emitterspannung 2 V und die Verlustleistung je nach eingestelltem Kollektorstrom 2 mW, 20 mW, 200 mW oder 2 W. Die Leistung hat ihren Maximumwert während die Messung vorgenommen wird, und es besteht daher keine Überlastungsgefahr beim Eichen des Kollektorstroms.

Sobald diese vorbereitenden Arbeiten erledigt sind, kann der Bassisstrom gemessen werden, um die durch das Verhältnis I_C/I_B ausgedrückte statische Verstärkung zu erhalten. Das zum Messen von I_B benutzte Galvanometer ist direkt in Verstärkungswerten geeicht.

EE 69 764 *für weitere Einzelheiten*

ORÉGA

(Société Électronique et Méchanique)

106, rue de la Jarry, Vincennes (Seine)

FERNSEH-TUNER

(*Abbildung Seite 338*)

Dieser mit "Rotomatic" bezeichnete VHF-Tuner-Verstärker hat sehr kleine Abmessungen und ist für tragbare Fernseher gedacht; er kann mechanisch oder elektrisch abgestimmt werden.

Die röhrenbestückte Ausführung soll die kleinste auf dem Markt sein, die den Anforderungen der 819-Zeilen-Norm genügt. Sie hat eine hohe Verstärkung, der Rauschwert ist nur 5 ... 7 dB, und sie ist mit den Röhren ECC 189 und ECF 801 bestückt.

Eine Transistor-Ausführung mit hoher Verstärkung und sogar noch niedrigeren Rauschzahlen ist auch lieferbar und hat bei 12 V eine Stromaufnahme von 17 mA.

Beide Typen sind sowohl für die französische wie auch ausländische Normen lieferbar.

EE 69 765 *für weitere Einzelheiten*

QUENTIN ET CIE

2, rue Hoche, Ermont (Seine-et-Oise)

TRANSISTORISIERTE STROMVERSORGUNGEN

(*Abbildung Seite 339*)

Diese Konstant-Stromversorgungen sind für Anschluss an Netze mit 110 ... 127 V oder 220 ... 240 V, 48 ... 65 Hz bestimmt und geben Spannungen von 0 ... 150 V= ab. Die drei lieferbaren Modelle AS280, AS281 und AS282 sind für Ausgangsströme von 0 ... 1 A, 0 ... 2 A bzw. 0 ... 5 A bemessen.

Die Regelung der Ausgangsspannung ist bei Netzschwankungen von ±15% besser als 0,05%. Die Regelung der Ausgangsspannung bei Belastungsschwankungen von Leerlauf bis Vollast ist besser als 0,05%.

Nach 30 Minuten Einlaufzeit ist die Ausgangsspannung über eine Zeitspanne von acht Stunden innerhalb 0,1% konstant. Sie schwankt bei Änderungen der Umgebungstemperatur zwischen −10° C und +50° C weniger als 0,05%/° C. Die Restwelligkeit des Ausgangs ist niedriger als 3 mV_{SS}.

Eine Strombegrenzungsschaltung kann zwischen 20% und 110% des Ausgangs-Nennstroms eingestellt werden.

EE 69 766 *für weitere Einzelheiten*

RIBET-DESJARDINS

13-17, rue Périer, Montrouge (Seine)

TRANSISTORISIERTER OSZILLOGRAF

Der Oszillograf 349A ist volltransistorisiert und verbindet die Vorteile kleiner Abmessungen mit den Eigenschaften eines zuverlässigen Instruments für Hochleistungsmessungen. Er kann an das Netz angeschlossen oder mit einer aufladbaren Batterie betrieben werden.

Der Vertikalablenkungsverstärker hat 0...1,5 MHz Bandbreite bei einer Empfindlichkeit von 50 mV/Teilung und von 8 Hz ... 1,5 MHz bei 10 mV/Teilung. Er hat einen geeichten Abschwächer mit Stufen von 10 mV/Teilung bis zu 20 V/Teilung.

Synchronisierung kann automatisch, intern oder extern erfolgen. Die erforderliche Mindestfrequenz ist 10 V_{ss}, und der Eingang ist gegen Überspannungen bis zu 300 V geschützt.

Ein eingebauter Eicher gibt Rechteckwellen von 1 V_{SS} ab.

EE 69 767 *für weitere Einzelheiten*

SOCAPEX

9, rue Edouard-Nieuport, Suresnes (Seine)

MEHRPOLIGE STECKVERBINDUNGEN

(*Abbildung Seite 339*)

Steckverbindungen der Serie 67 sind sechspolige, stapelbare Steckverbindungen für hohe Betriebsspannungen mit drei Steckern + drei Buchsen und wurden im Laufe des letzten Jahren wesentlich verbessert.

Die Endbeschläge und entsprechenden Arretierungen der gekapselten Modelle werden jetzt in verschiedenen Farben ausgeführt, um Irrtümer zu vermeiden.

Durch das angewandte Markierungssystem kann man jeden Kontakt sofort erkennen, ganz gleich ob es sich um Stecker oder Buchse handelt.

EE 69 768 *für weitere Einzelheiten*

SOVIREL

27, rue de la Michadière, Paris 2e

MINIATURWIDERSTÄNDE UND -KONDENSATOREN

(Abbildung Seite 339)

Die Lösung der Qualitäts- und Raumprobleme der neuzeitlichen Schaltungstecknik kann in den Fertigungsprogrammen der hochkonstanten und sehr hochkonstanten Metalloxyd-Schichtwiderstände ($\frac{1}{2}$ Ω ... 6 kΩ) und Kondensatoren mit Glasdielektrikum mit Grenzwerten von 0,5 pF... 0,01 μF und Verlustwinkeln von weniger als 10×10^{-4} gefunden werden.

Widerstände der Typen CR, A und HR haben eine nachgewiesene Ausfallrate von nur 0,00069%/1000 h (es wurden 7000 Widerstände 2000 Stunden lang bei 2,5 × Nennspannung geprüft).

Speziell für Tuner entwickelte Miniaturkondensatoren sind in Werten von 0,1 bis zu 30 pF lieferbar.

EE 69 769 *für weitere Einzelheiten*

TACUSSEL SOLEA

2 et 4, rue Carry, Lyon, 3e (Rhone)

ELEKTRONISCHES VIELFACHINSTRUMENT

(Abbildung Seite 339)

Das elektronische Vielfachinstrument VE7 ist bei Einsatz als Voltmeter durch eine sehr hohe Eingangsimpedanz gekennzeichnet. Der Verstärker des Gerätes ist mit einer Elektrometerröhre und zehn Transistoren (vorwiegend Silizium) bestückt. Es wird aus internen Batterien gespeist und ist für folgende Hauptmessungen ausgelegt:

Spannung: 1 mV ... 1 kV=; Messunsicherheit 1% des Skalen-Endwertes.

Strom: 1×10^{-12} A ... 1 mA=; Messunsicherheit 1% des Skalenendwertes.

Widerstand: 2Ω ... 1000 GΩ; Messunsicherheit 2 ... 3%.

Zusätze sind für folgende Messungen lieferbar: Integration von Elektrizitätsmengen; Kapazitäten; kleine Spannungsschwankungen; Leitfähigkeit; pH-Werte.

EE 69 770 *für weitere Einzelheiten*

KLIRRFAKTORMESSER

(Abbildung Seite 339)

Der Klirrfaktormesser DH2 ist ein Breitbandinstrument zum Messen von Klirrfaktorwerten von 0,05% ... 100% in Signalen, deren Grundfrequenz in den Bereich 3 Hz ... 300 kHz fällt. Es enthält aktive elektronische Filter in Wien-Brückenschaltung und ein Breitband-Wechselstrom-Millivoltmeter. Ein niederohmiger Ausgang ist für Anschluss eines Oszillografen zur Untersuchung der Restwellenformen vorhanden.

Das Gerät kann auch als Wechselspannungsmesser für Signale von 0,5 mV ... 50 V (3 Hz ... 1 MHz) eingesetzt werden.

EE 69 771 *für weitere Einzelheiten*

TELEFUNKEN A.G.

Ernst-Reuter-Platz 7, 1 Berlin 10, Deutschland

UHF-FERNSEH-TUNER

Dieser Tuner ist für das Fernsehband 470 ... 860 MHz ausgelegt. Das robuste, flache Gehäuse wird durch Zwischenwände in vier fast gleichgrosse Räume unterteilt, die in jedem Fall als Schwingkreise dienen und mittels eines Dreifach-Drehkondensators auf die gewünschte Frequenz abgestimmt werden können. Der Tuner 132 ist für Handabstimmung ausgelegt und mit Röhren der Typen EC86 bestückt. Der UHF-Tuner 140 ist in demselben Gehäuse wie die Type 132 untergebracht. Er ist mit zwei Transistoren AF139 bestückt, der erste in der Eingangsstufe, der zweite in der selbsterregten Summierstufe. Beide Tuner sind in Halbwellentechnik konstruiert. Ab Mitte 1964 wird jedoch auch ein Viertelwellentuner nach Frankreich geliefert werden, der ebenfalls mit zwei Transistoren AF139 bestückt ist.

EE 69 772 *für weitere Einzelheiten*

ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 340 bis 343

Magnetischer Verstärker

Vertrieb: Aoto-Electronics Ltd, Peel Grove, London, E.2

(Abbildung Seite 340)

Der von Serelec in Paris hergestellte hochwertige, magnetische Gegentaktverstärker AM-TH-PP ist ein vielseitiges Gerät für das Messen von Gleichstrom-Kleinsignalen von Thermopaaren, Dehnstreifen, verschiedenen Brücken, Hall-Elementen usw. Er ist ebenfalls für Einsatz konstanter Vorverstärker in geschlossenen Regelsystemen geeignet.

Gute Linearität, Konstanz und robuste Konstruktion bei kleinen Abmessungen, geringes Gewicht und niedriger Preis zeichnen ihn aus. Diese Verstärker haben keine beweglichen Teile und unbeschränkte Lebensdauer, vorausgesetzt dass sie innerhalb der festgelegten Grenzen eingesetzt werden. Die tote Zone, ein ziemlich weitverbreiteter Nachteil dieser Verstärkerlasse, ist ausgemerzt, und Signale von 0 ... 100 μV werden genau so naturgetreu verstärkt wie die von 0 ... 100 mV. Das Gerät ist gegen externe Magnetfelder unempfindlich.

Der Verstärker hat hohe dynamische Eingangs- und Ausgangsimpedanzen, und Widerstandsschwankungen der Quelle oder Last haben einen vernachlässigbaren Einfluss auf Verstärkung und Genauigkeit. Dadurch werden die durch Widerstandsänderungen in Verbindungen oder Kontakten auftretenden Schwierigkeiten vermieden. Man kann den Verstärker auch mit koaxialen Thermopaaren benutzen.

EE 69 773 *für weitere Einzelheiten*

Glimmer-Kondensatoren

Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(*Abbildung Seite 340*)

Zwei alternative Standardausführungen der Kondensatoren mit versilberten Glimmerplättchen sind jetzt bei Plessey-UK Ltd lieferbar. Die wachsgetauchte Ausführung, die einen rissefreien Spezialwachs- und Lacküberzug hat, ist für Temperaturen von −25° C ... +70° C geeignet. Die Bauformen mit Epoxydharzumhüllung geben Feuchtigkeitsschutz nach Klasse H2 und haben den grösseren Temperaturbereich −40° C ... +85° C.

Die Kondensatoren können in jeder Kapazität bis zu 1,0 µF bei 200 V= Betriebsspannung, 0,75 µF bei 350 V, 0,5 µF bei 500 V und 0,05 µF bei 750 V geliefert werden. In der Standardausführung werden je nach Kapazität vier Plättchengrössen benutzt, und zwar von 11,1 × 6,4 mm bis zu 25 mm × 13 mm. Die Anschlussdrähte (üblicherweise radial, aber auf Wunsch axial) bestehen aus verzinntem Kupferdraht und sind in einer Weise befestigt, die gute elektrische Verbindungen und grosse Festigkeit gewährleistet. Spezial-Glimmerkondensatoren für Temperaturen bis zu 250° C sind auf Wunsch ebenfalls lieferbar.

EE 69 774 *für weitere Einzelheiten*

Auswuchtmaschine

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(*Abbildung Seite 340*)

Die neue Auswuchtmaschine 1250C wurde als einfaches und doch genaues Mittel zum Auswuchten rotierender Werkstücke aller Art in der Fertigungsstrasse durchkonstruiert.

Es ist eine kompakte Maschine für Aufbau auf einen Arbeitstisch (Grundplatte 381 × 318 mm, Höhe 381 mm, Gewicht 25 kg). Auf einem Untersatz, der alle Schaltungen enthält, ist eine robuste Schiene aufgebaut, auf der zwei offene Lagerstühle seitlich so verschoben werden können, dass sie Werkstücke von 50 ... 330 mm Länge und 28 g ... 4,5 kg Gewicht aufnehmen können. Das zu prüfende Werkstück wird durch einen leichten Riemen von einem eingebauten Elektromotor mit 600 ... 3000 UPM getrieben.

Die Lagerstühle sind mit Prismenauflagen ausgerüstet, und zwar so, dass sie in Querrichtung frei beweglich sind. Auf Unwucht des Werkstücks zurückzuführende Schwingungen werden durch die Prismenauflagen auf Sondenspulen in einem starken Magnetfeld übertragen. In den Sondenspulen wird eine der Schwingungsamplitude des Werkstücks proportionale Wechselspannung erzeugt, die nach Verstärkung einem Messgerät zugeleitet wird und gleichzeitig ein Stroboskop mit weisser Lichtquelle über dem Werkstück triggert. Ein numeriertes Band oder ein Ring am Werkstück wird dadurch "festgehalten" und zeigt gegenüber einem einstellbaren Zeiger, wo zum Auswuchten entweder Gewicht erforderlich ist oder entfernt werden muss (je nach Position eines Schalters).

Für dynamisches Gleichgewicht muss das Gewicht in zwei Querebenen justiert werden. Mittels eines weiteren Schalters kann entweder der rechte oder linke Lagerstuhl für unabhängige Anzeige geschaltet werden.

Die Maschine 1250C hat eine eingebaute, regelbare Filterschaltung, die Aussieben von Störschwingungen erlaubt. Jedem der beiden Lagerstühle sind Schaltungen für Eichung und Empfindlichkeitsregelung zugeordnet. Diese werden für Fertigungskontrolle voreingestellt, so dass selbst ungelernte Kräfte nach kurzer Unterweisung die Unwucht eines Werkstücks in 20 Sekunden bestimmen können.

Das hochintensive weisse Licht des Stroboskops erlaubt Einsatz der Maschine bei Tageslicht. Die offenen Prismenauflagen erleichtern Einlegen und Entfernen des Prüflings. Die beiden Lager und der Schalter machen Umkehr der Drehrichtung unnötig und beschleunigen den Arbeitsablauf.

EE 69 775 *für weitere Einzelheiten*

Mehrzweck-Potentiometer

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(*Abbildung Seite 341*)

Dieses neue Instrument ersetzt das Cambridge-Werkstoffpotentiometer 44226; es ist kompakt, in sich geschlossen und hauptsächlich für das Prüfen und Eichen von Thermopaaren und zugehörigen Anzeigegeräten an Ort und Stelle bestimmt.

Das Potentiometer hat einen Bereich von 0,01 ... 101 mV, der mit einem mehrpoligen Schalter und einem geeichten Schleifdraht eingestellt wird; die Einstellunsicherheit ist ±0,1% oder ½ Teilung des Schleifdrahtes, wobei jeweils die grössere gilt. Die den mehrpoligen Schalter kontrollierende Hauptspannungsskala hat 100 Stufen von je 1 mV; die den Schleifdraht kontrollierende Feineinstellskala hat einen Bereich von 1 mV mit Teilungen von 0,01 mV, die Schätzung von 0,002 mV erlauben.

Ein fünfstufiger Funktionsschalter stellt die eine oder andere von zwei Testschaltungen ein, die hauptsächlich zusammen für den Vergleich von Normal- und Test-Thermopaaren benutzt werden. Eine dritte Position legt eine kontinuierlich regelbare mV-Spannung zur Ablenkung von Drehspulinstrumenten an die Testklemmen, während eine weitere Stellung Verwendung als geeichte Spannungsquelle für potentiometrische Instrumente erlaubt.

Das Instrument ist mit eingebautem Galvanometer, Normalelement und Trockenbatterien völlig in sich geschlossen; der Bereich lässt sich mit ein paar einfachen Zusätzen auf 500 V= und 10 A= erweitern.

Das Potentiometer kann jederzeit während des Tests und ohne Verstellen von Skala oder Funktionsschalter durch Drücken einer Taste und Betätigung eines Drehwiderstands nachgeeicht werden.

EE 69 776 *für weitere Einzelheiten*

Präzisions-Widerstände

Electrothermal Engineering Ltd, 270 Neville Road, London, E.7

(*Abbildung Seite 341*)

Electrothermal Engineering Ltd hat ihr Programm für alle umpressten Präzisions-Drahtwiderstände der Serie G101 durch zwei Ausführungen mit Lötfahnen erweitert.

Es handelt sich um die Type 3505T ¼ W und 3507T ½ W, die dieselbe eingepresste Verankerung haben wie alle Precistors G101. Dadurch wird eine robuste Bauart mit hoher Zuverlässigkeit sowie 100 %ige positive Abdichtung gewährleistet.

Die ¼-W-Type ist in Werten von 0,1 Ω ... 1 MΩ, die ½-W-Type von 0,1 Ω ... 2 MΩ lieferbar. Die Toleranzen liegen je nach Wert zwischen 0,1 % und 0,01 %; die Stabilität ist besser als ±0,02 %. Die höchsten Betriebsspannungen sind 550 V bzw. 750 V.

EE 69 777 *für weitere Einzelheiten*

Schutzgasrelais

Erg Industrial Corporation, Luton Road Works, Dunstable, Bedfordshire

(*Abbildung Seite 341*)

Ein neues kompaktes Schutzgasrelais für die industrielle Verfahrenstechnik, Elektronenrechner und verwandte Ausrüstungen wurde von der Erg Industrial Corporation mit der Typenbezeichnung AB.1618 angekündigt.

Das neue Relais vereint kurze Schaltzeiten, und zwar bis zu 1 ms herunter, mit einer langen Lebensdauer von vielen Millionen Betätigungen.

Die hermetisch in einer inerten Atmosphäre verschlossenen vergoldeten Kontakte sind entweder als einpolige Umschalter oder Einzel-Arbeitskontakte lieferbar. Der Schutzgasschalter ist so montiert, dass er leicht ausgeschaltet werden kann, und die magnetische Abschirmung des Relais verringert den Einfluss magnetischer Fremdfelder.

Die Spulenkörper des Relais werden aus glasfaserverstärktem Epoxydharz hergestellt, und die eingepressten

Spulenanschlüsse sind nach der modernsten Montagetechnik für gedruckte Schaltungen ausgelegt.

Die Metallabschirmung kann ohne Extrakosten mit Befestigungslappen geliefert werden (Type AB. 1652).

EE 69 778 *für weitere Einzelheiten*

Impedanzmesser

Vertrieb: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Abbildung Seite 341)

Der von Radiometer in Dänemark hergestellte Impedanzmesser GB11 kann jetzt von Livingston Laboratories Ltd bezogen werden.

Mit dem Gerät lassen sich Impedanzen in Form von Grösse und Phasenwinkel von 1 Ω...1,1 MΩ und von 0°...±90° messen. Die Grund-Messunsicherheit ist ±1 % für die Grösse und ±0,5° für den Phasenwinkel. Der interne Generator gibt zwölf Testfestfrequenzen, und externe Quellen können bis zu 1 MHz benutzt werden. Der Teststrom ist von 3,2 μA bis zu 1 A regelbar. Es können erdfreie, geerdete oder erdsymmetrische Impedanzen gemessen werden, und Gleichstrom-Polarisation ist vorgesehen. Andere Kennzeichen sind transistorisierte Schaltungen, einzeilige Anzeige der Grösse und ein Messgerät für Direktanzeige des Phasenwinkels.

Als Einsatzmöglichkeiten werden Impedanzmessungen an Filter-Netzwerken, Spulen, Lautsprechern und Übertragern, sowie Messungen der Zusatzimpedanz von Drosseln mit Eisenkern und die effektive Impedanz nichtlinearer Elemente bei verschiedenen Testströmen genannt.

EE 69 779 *für weitere Einzelheiten*

Generator für Längstwellenfrequenz

Vertrieb: Claude Lyons Ltd, Hoddesdon, Hertfordshire

(Abbildung Seite 341)

Das von Constructions Radioélectriques et Électroniques du Centre hergestellte Gerät GB860 gibt Sinus-, Dreieck- oder Rechteckwellen im Längstwellenfrequenzbereich von 0,001 Hz...1 kHz mit 0,5 W Ausgangsleistung ab. Ein vorhandener Ausgang mit regelbarem Phasenwinkel erlaubt Einsatz des Gerätes zum Messen von Phasenverschiebungen. Es ist besonders für Untersuchungen an Servovorrichtungen, mechanischen Verformungen und Dehnungen gedacht, kann aber auch für Übertragungsleitungen, Übertrager für Unterschallfrequenzen, medizinische Forschung, Geophysik und zur Abgabe genauer Signale für Programmierungszwecke Verwendung finden.

In der Oszillatorschaltung werden die Schmitt-Trigger- und Miller-Integratorschaltung zur Erzeugung symmetrischer Dreieck- und Rechteckwellen stetig regelbarer Frequenz und Amplitude benutzt; der Frequenzumfang wird in sechs lineargeeichten Teilbereichen überstrichen. In einer wellenformenden Schaltung erzeugen Dioden-Abkapper in Kaskade Sinuswellen aus Dreieckwellen.

Der Endverstärker ist transistorisiert, wodurch jede der drei Wellenformen bei niedriger Impedanz symmetrisch oder unsymmetrisch zur Verfügung steht.

Für den Gleichlauf der Zeitablenkung eines Oszillografen wird durch Differenzierung des Rechteckwellen-Oszillatorausgangs ein positiver Triggerimpuls erzeugt. Ein zweiter Ausgangsimpuls mit über einen Bereich von 360° veränderlicher Phase ermöglicht die genaue Bestimmung von Phasenverschiebungen. Bei Benutzung dieses Ausgangs in Verbindung mit einem Zweistrahl-Oszillografen, z.B. dem C.R.C.-Modell OC728, kann man den Phasenwinkel zwischen Generator-Ausgang und zu prüfendem System mit weniger als ein Grad Unsicherheit messen.

EE 69 780 *für weitere Einzelheiten*

Messwertwandler

S.E. Laboratories (Engineering) Ltd. North Feltham, Middlesex

(Abbildung Seite 342)

S.E. Laboratories (Engineering) Ltd hat ein preisgünstiges, tragbares und in sich geschlossenes Ein- oder Zweikanal-Messwertwandlersystem S.E.905 herausgebracht. Die Ausrüstung ist für Anzeige und /oder Aufzeichnung der von induktiven oder Dehnstreifen-Messwertgebern abgegebenen Information geeignet.

Das Messen von Druck, mechanischer Beanspruchung, Weg usw. ist daher an Ort und Stelle durchführbar.

Ein Sondermerkmal des S.E.905 ist der Netzanschluss mit aufladbarer Batterie, die Verwendung des Gerätes im Aussendienst erlaubt. Die Ausrüstung hat einen konstanten 3-kHz-Oszillator.

Verstärkereinschübe sind in zwei Typen lieferbar: Type S.E.423 ist in erster Linie für statische oder dynamische Messungen mit Dehnungsstreifen oder Dehnungsstreifen-Messwertgebern bestimmt, Type S.E.449 für statische oder dynamische Messungen mit induktiven Gebern oder Differentialübertragern.

Ein eingebautes Messgerät mit Kreisskala kann Kundenwünschen entsprechend geeicht werden. Ausgangsklemmen für den Anschluss von Galvanometern, Magnetbandgeräten und Oszillografen sind vorhanden.

EE 69 781 *für weitere Einzelheiten*

Niederspannungs-Stromversorgung

Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire

(Abbildung Seite 342)

Die Konstant-Stromversorgung L.P.401 mit Messgerät ist die erste einer neuen Serie hochkonstanter, preiswerter Präzisionsgeräte, deren Leistung mit Geräten einer viel höheren Preislage vergleichbar ist.

Ausführungen für 0,5 und 1 A, 10...30 V sind mit Einzel- und Doppelausgang lieferbar, wobei das Pflichtenblatt in jedem Fall ähnlich ist.

Spannung, Strom und die über den vollen Bereich regelbare Einstellung des konstanten Stroms werden ohne Bereichumschaltung auf einem 89-mm-Messgerät mit klarer Skala überwacht.

Grob- und Feinregelung der Spannung ermöglicht äusserst genaues Einstellen des Ausgangs; die Feinregelung hat für alle Ausgangspegel einen Regelbereich von ±0,75 V.

Besondere Aufmerksamkeit wurde dem Entwurf der Regelschaltungen für den konstanten Strom gewidmet, die über einen sehr breiten Bereich äusserst empfindliche, stabile und genaue Strombegrenzungsregelung ergeben und damit empfindliche Geräte aller Art schützen.

Alle Geräte der Serie 400 können Kurzschluss des Ausgangs für längere Zeit ohne Beschädigung überstehen.

EE 69 782 *für weitere Einzelheiten*

Buchsen für gedruckte Schaltungen

Oxley Development Co. Ltd, Priory Park, Ulverston, Lancashire

(Abbildung Seite 342)

Die Oxley-Buchsenleiste für gedruckte Schaltungen Type 50S/PCB ist für Direktmontage auf gedruckte Schaltungen mit 2,54 mm Rastermass und 1,27 mm Mindestlochdurchmesser ausgelegt. Die Montage erfolgt in einer Weise, die Einführung des Steckers parallel zur Leiterplatte und damit enges Stapeln derselben ermöglicht. Die Überschlagspannung ohne Leiterplatte ist 5 kV=, die zusätzliche Schaltungskapazität je Kontakt liegt unter 0,7 pF, und die Höhe der Leiste über der Leiterplatte ist nur 6,4 mm. Die Buchsenleiste kann ein- bis zwölfpolig geliefert werden; sie ist für Aufnahme des Steckers 50P/156 ausgelegt, wodurch Testpunkte mit geringerem Abstand voneinander angeordnet werden können. Bei mehrpoligen Leisten sind Buchsen mit Farbkennzeichnung erhältlich, und in diesem Fall muss die Anzahl der Buchsen mit ihrer Farbkennzeichnung von links nach rechts gehend aufgegeben werden. Lieferbar mit folgenden Farben: schwarz, braun, rot, orange, gelb, grün, blau, violett, rosa, grau und weiss.

EE 69 783 *für weitere Einzelheiten*

Vorwahlzähler

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(Abbildung Seite 342)

Advance Controls Ltd hat einen neuen Zeitgeber-Vorwahlschalter TB42 angekündigt.

Man kann dieses Instrument entweder zur Zeitsteuerung eines Verfahrens oder für laufendes oder einzelnes Zählen einer Partie einsetzen. Steuergeschwindigkeiten von 0,01 s sind vorhanden, und Vorwahlzählen lässt sich mit Geschwindigkeiten von bis zu 1000/s durchführen. Der TB42 hat momentane Rückstellung, so dass bei wiederholtem Arbeitstakt Zähl- und Zeitverluste vermieden werden. Die Wiederholung kann automatisch, ferngesteuert oder mittels Drucktaste eingeleitet werden.

Bestückung mit Kaltkathodenröhren erlaubt optische Überwachung des korrekten Zählens, und das Gerät kann bis zu 99 Zählimpulse oder 0,99 Sekunden vor Erreichen der vorgewählten Zahl ein Signal auslösen.

Das Ausgangsrelais kann so angeordnet werden, dass es nach Ablauf einer bestimmten Zeit automatisch rückgestellt wird. Andererseits kann die Rückstellung auch extern gesteuert werden.

Der TB42 ist 229 mm breit, 165 mm hoch und 140 mm tief. Für dieses Gerät ist ein komplettes Zubehör-Sortiment von photoelektrischen, magnetischen, Schutzgas- und Nahwirkungs-Abtastern erhältlich.

EE 69 784 *für weitere Einzelheiten*

Elektronische Schaltuhr

Pioneer Designs Ltd, Crown House, Walton-on-Thames, Surrey

(Abbildung Seite 343)

Die Schaltuhr hat eine impulsbetätigte Triggerschaltung mit von der Frontplatte einstellbarer, regelbarer Empfindlichkeit. Die Empfindlichkeit kann so geregelt werden, dass von externen Quellen an den hochohmigen Eingang der Triggerschaltung gelegte niedrige Gleich- oder Wechselspannungen den gesteuerten Zyklus einleiten können. Andererseits lässt sich die Empfindlichkeit bis zu einem Punkt erhöhen, bei dem sich der Ablauf automatisch wiederholt.

Das Gerät hat zwei Zeitbereiche: 0,2 ... 1 s und 1 ... 5 s. Die Zeitspanne ist stetig regelbar und wird mit Hilfe einer 127-mm-Skala eingestellt.

Ausserdem kann die Zeitsteuerung auch durch Betätigung einer Drucktaste auf der Frontplatte oder durch Kurzschliessen von zwei Ausgangsklemmen, z.B. mittels Mikroschalter, eingeleitet werden.

Im Ausgang liegt ein als einpoliger Umschalter wirkendes Relais, und Ausgangsklemmen zum Schliessen oder Unterbrechen getrennt für 2 A gesicherter externer Kreise sind vorhanden. Bei Umlegen einer internen Verbindungslasche können bis zu 500 W bei 230 V 50 Hz an die Ausgangsklemmen gelegt werden. Der Verbrauch der Schaltuhr liegt unter 8 W.

EE 69 785 *für weitere Einzelheiten*

Entladungsfeste Kondensatoren

F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(Abbildung Seite 343)

F. C. Robinson & Partners Ltd geben bekannt, dass sie nunmehr ein komplettes Sortiment ionisationsentladungsfester Kondensatoren liefern können.

Bei der Ausführung von Ionisationsentladungstests wird nur dann mit beliebigen Messeinrichtungen Höchstempfindlichkeit erreicht, wenn der Wert des Kopplungskondensators im Vergleich mit dem Prüfling hoch ist. Die folgenden Standardwerte wurden unter Berücksichtigung dieses Umstandes gewählt:

1000 pF ... 0,05 μF bei 50 kV_{eff}
2000 pF ... 0,1 μF bei 25 kV_{eff}
5000 pF ... 0,5 μF bei 10 kV_{eff}
0,01 μF ... 2 μF bei 5 kV_{eff}

Höhere Betriebsspannungen und grössere Kapazitätsspannungen kann man durch Serien- und Parallelschalten gewisser Einheiten erreichen. Die Abbildung zeigt einen Parallelsatz von Kondensatoren von 0,05 μF, 50 kV.

Je nach Nennspannung und Kapazitätswert werden einige Typen in der bekannten Rohrkonstruktion, andere in ölgefüllten Stahlgehäusen oder in Hart-PVC-Behältern geliefert. Alle werden in jeder Herstellungsstufe sorgfältig geprüft, um gewährleisten zu können, dass sie entladungsfest sind und bleiben.

EE 69 786 *für weitere Einzelheiten*

Digital-Grenzwächter

Digital Measurements Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Abbildung Seite 343)

Digital Measurements Ltd hat einen Digital-Grenzwächter DM5080 und ein Logik-Gerät DM5006 herausgebracht, die zusammen ein Alarmsystem für Einsatz mit Digital-Voltmetern, Anlagen für Digital-Messerfassung, Zähler usw. bilden.

Das Logik-Gerät kann bis zu drei Dezimaldekaden aufnehmen und bis zu 16 Grenzwächter treiben. Wächter für die "obere Grenze" geben einen Ausgang ab, wenn das in das Alarmsystem gespeiste Signal dem vorgewählten Pegel gleich ist oder ihn übersteigt. Wächter für die "untere Grenze" geben einen Ausgang ab, wenn das in das System gespeiste Signal unter den vorgewählten Pegel fällt. Die Wächter können durch Umlegen von Verbindungen ohne Schwierigkeiten von einem Typ in den anderen umgewandelt werden.

Das Ausgangssignal kann entweder automatisch Abhilfsmassnahmen einleiten oder den Anlagenwächter durch beleuchtete Felder, hörbare Warnung usw. davon in Kenntnis setzen, dass ein Vorgang ausserhalb der Toleranz läuft. Andererseits können auch die Eingangssignale von verschiedenen Quellen mittels eines Erfassungssystems, in dem die Farbbandumschaltung der Schreibmaschine oder des Druckwerks automatisch erfolgt, aufgezeichnet werden.

Die Einheiten sind in Modulbauweise ausgeführt; ein lieferbarer 178 mm hoher Einschub für ein 19"-Gestell kann acht Grenzwächter oder ein Logik-Gerät und vier Wächter aufnehmen. Die Wächter werden aus dem Netzanschluss des Logik-Gerätes, der für 100 ... 125 V~ oder 200 ... 250 V~ ausgelegt ist, gespeist.

EE 69 787 *für weitere Einzelheiten*

Analog-Digital-Bausteine

Trainer and Simulator Division, Elliott Brothers (London) Ltd, Cobham Road, Frimley, Nr. Aldershot, Hampshire

(Abbildung Seite 343)

Eine von der Trainer- und Simulator-Abteilung der Elliott-Automation Ltd entwickelte neue Serie von Standard-Bausteinen ermöglicht den Aufbau von Analog-Digital-Rechner- oder Simulatorsystemen in jeder gewünschten Verknüpfung. Die Bausteine wurden ursprünglich für Sonderzweck-Simulatoren modernster Konstruktion entworfen und ermöglichen System-Ingenieuren, mit geringstem Kostenaufwand zu bekannter Genauigkeit und Leistung zu bauen. Das äusserst breite Anwendungsgebiet umfasst unter anderem "Brettschaltungs"-Erprobung von Analog- und Digital-Systemen, Unterricht in der Theorie der Regelungstechnik und Bau von Konstruktions- und Ausbildungssimulatoren.

Die Grundbausteine, die alle einzeln lieferbar sind, bestehen aus einer Servoelektronik-Platte, die einen kombinierten Modulator-Vorverstärker und Leistungsverstärker zum Treiben von Standard-Stellmotoren enthält, einem sehr genauen Getriebe mit sieben Ausgangswellen, das Übersetzungen von 1 : 18750 für den Antrieb von Stell-Bauelementen der Grössen 08 bis 15 gibt, einer Demodulator-Platte für die Umsetzung von 400-Hz-Signalen und Wechselstrom- und Gleichstrom-Mehrzweckverstärker. Die Bausteine können zusammen mit ihren Stromversorgungen in einem modularen Gestellsystem untergebracht werden.

Aus diesen neuen Elliott-Analog-Bausteinen kann man Servosysteme in grosser Auswahl zusammenbauen, z.B. eine Gleichstrom-Positionsservoanlage oder ein Geschwindigkeits-Servosystem mit Gleichstrom- oder 400-Hz-Eingangssignalen und Drehfeldnachläufer oder Funktionsdrehmelder. Ausgangssignale können die Form von Wellenumläufen nehmen oder von Potentiometern, Drehfeldgebern, Funktionsdrehmeldern oder Wellendrehgebern abgegebenen Signalen hergeleitet werden. Wellenwinkel lassen sich mit Hilfe von Kupplungen auf den Getriebewellen auf der Frontplatte des Systems anzeigen.

EE 69 788 *für weitere Einzelheiten*

Lötbarkeitstests

Multicore Solders Ltd, Maylands Avenue, Hemel Hempstead, Hertfordshire

(Abbildung Seite 343)

Die geforderten höheren Produktionsleistungen im Zusammenbau elektronischer Ausrüstungen haben einen Bedarf für ein einfaches, genaues Verfahren zur Bestimmung der Lötbarkeit der Anschlussdrähte von Bauelementen geschaffen.

In Zusammenarbeit mit der Electronic Engineering Association hat Multicore Solders Ltd eine Maschine entwickelt, die die einzige lieferbare Ausrüstung zum Testen von Anschlussdrähten auf Lötbarkeit sein soll.

In Betrieb bringt die Maschine einen vorher in Flussmittel getauchten Prüfdraht sanft auf einen geschmolzenen Löttropfen herab. Wieviele Sekunden das Lot braucht, um um den Draht zu fliessen und sich über ihm zu treffen, ist ein Massstab der Lötbarkeit des Drahts. Eine eingebaute Stromversorgung speist das Thermoelement, das die Heizkörpertemperatur regelt, und die Temperatur wird mittels eines batteriegespeisten Temperaturfühlers mit 1% Messunsicherheit angezeigt.

Die Testmaschine für Lötbarkeit wird komplett mit Temperaturfühler, in 0,1 s geeichter Stoppuhr, Bedienungsanweisung, Mustern von Lotpastillen und flüssigem Flusmittel für Export verpackt geliefert.

EE 69 789 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Für Kardiotachometrie eingesetzter Impulsperiodenmesser mit kurzer Ansprechzeit von P. A. Tove und J. Czekajewski

Zusammenfassung des Beitrages auf Seite 290-295

Eine Messmethode für Impulsfolgefrequenzen, die schnell auf Frequenzänderungen anspricht, wurde entwickelt. Das zugrundeliegende Prinzip ist die Bestimmung der Intervalle zwischen Impulsen durch umschichtiges Aufladen von zwei Kondensatoren während aufeinanderfolgender Impulsintervalle. Die Spannung an dem Kondensator, der nicht aufgeladen wird, wird auf einem Messgerät oder Oszillografen angezeigt und ist ein Massstab der letzten Intervalldauer. Dieses Prinzip ist anderen Zeit-Spannungswandlern mit Kondensatorenaufladung überlegen. Die Intervalldauer oder Frequenz der Impulse kann direkt auf der Skala abgelesen werden, und die Möglichkeit, lineare oder logarithmische Frequenzskalen zu erreichen, wird besprochen. Es wird gezeigt, dass einfaches Exponentialladen der Kondensatoren für begrenzte Frequenzbereiche ungefähr die gewünschte Skalenform ergibt. Methoden zum Erlangen linearer und logarithmischer Skalen für Grossbereiche werden besprochen. Praktische Schaltungen für medizinische Anwendungsmöglichkeiten in der Kardiotachometrie und Registrierung der Atemfrequenz, bei denen es auf schnelles Ansprechen auf Änderungen der Impulsrate ankommt, werden beschrieben. Andere Anwendungsmöglichkeiten bestehen in Informationsübertragungssystemen mit Pulsbreitenmodulation.

Telemetriedurchmusterungserscheinungen von S. Poole

Zusammenfassung des Beitrages auf Seite 296-299

In diesem Beitrag werden verschiedene Fragen der Telemetriedurchmusterung besprochen, und zwar (1) Bandbreite, (2) Informationsverkennung und (3) eine vorgeschlagene "Such"-Abtastung. Im allgemeinen werden sie für den Fall, dass visuelle Analyse der Sampling-Wellenformen erforderlich ist, behandelt. Ausser der theoretischen Besprechung werden Einzelheiten der im Zusammenhang mit diesen drei Themen ausgeführten Laborarbeiten gegeben.

Eine Schaltung zum Messen der wahren Effektiv-Spannung von J. C. Cluley

Zusammenfassung des Beitrages auf Seite 300-303

Die in der automatischen Spannungsregelung für Wechselstromgeneratoren benutzten herkömmlichen Gleichrichterschaltungen mit Mittelwertanzeige können für wesentliche Ausgangsspannungsfehler verantwortlich sein, wenn die Wellenform der Maschine harmonische Komponenten enthält. Die Grösse dieses Fehlers und seine Abhängigkeit von der Amplitude und Frequenz der Harmonischen wird besprochen und eine Schaltung beschrieben, die eine Gleichspannung erzeugt, deren Wert der Effektiv-Spannung des Generators genau proportional ist. Die Schaltung vermeidet die mit mechanischen Einrichtungen oder temperaturempfindlichen Elementen verbundene Zeitverzögerung und benutzt eine vor kurzem entwickelte Cadmium-Sulfid-Diode, die eine quadratische Charakteristik hat.

Transistor-Gleichstromverstärker mit kleiner Drift von J. I. Pinto

Zusammenfassung des Beitrages auf Seite 304-308

Der Beitrag behandelt die Abweichungen der Gleichstromparameter von Transistoren, die für den Entwurf von Gleichstromverstärkern mit kleiner Drift wichtig sind. Der herkömmliche Differenz-Verstärker und die für Verbesserung der Leistungsdaten erforderlichen Abwandlungen werden besprochen und die Ergebnisse beurteilt. Abschliessend wird ein neuer, symmetrischer "Spiegelbild"-Verstärker, seine Leistung und Fähigkeiten beschrieben.

Niederspannungs-Hybridverstärkerschaltungen mit hochohmigem Eingang von F. H. Laishley

Zusammenfassung des Beitrages auf Seite 309-311

Der Beitrag beschreibt eine Schar gleichspannungsgekoppelter Hybridverstärkerschaltungen, die mit Niederspannungen betrieben werden und sehr hochohmige Eingänge haben.

In der Eingangsstufe jeder Schaltung wird eine Einzelröhre, die mit niedriger Anodenspannung arbeitet, benutzt, während weitere Schaltungen in herkömmlicher Halbleitertechnik ausgeführt sind. Dadurch ergeben sich einfache hochohmige Schaltungen kleiner mechanischer Abmessung und niedriger Stromaufnahme.

Die Schaltungen wurden ursprünglich für das Messen von in klein bemessenen Kondensatoren gespeicherten Spannungen entwickelt.

Abgestimmte Detektoren mit Doppel-T-Schaltung von T. Spencer

Zusammenfassung des Beitrages auf Seite 315-321

Dieser Beitrag befasst sich mit der spezifischen Anwendung der Doppel-T-Schaltung in Transistor-Verstärkern. Das Doppel-T wird als Parallel-Gegenkopplungselement einer Stufe zur Verstärkung eines engen Frequenzbandes eingesetzt, dessen Mitte ungefähr mit der Sperrfrequenz des Doppel-T zusammenfällt und bei 10 Hz ein Stufen-Q_F von ungefähr 5 hat; durch Kaskadenschalten von zwei dieser Stufen wird das Q_F quadriert. Ein Abstrakt der vollständigen Theorie des Doppel-T mit endlichen Quellen und Abschlussimpedanzen wird im Anhang gegeben. Ein Nomogramm für den optimalen Entwurf dieses Netzwerks wird gegeben. Dass das Q_F eines abgestimmten Verstärkers mit einem Doppel-T-Netzwerk im Gegenkopplungsweg direkt der Gleichstromübertragung des Netzwerks angepasst ist, die Güte seiner unbelasteten Schaltung und die Stufenverstärkung des Verstärkers, ist nachgewiesen, und daher braucht man die kaskadengeschalteten Stufen nicht einzeln abzustimmen, es sei denn, dass ein Bandpass-Effekt gewünscht und nur ein Doppel-T-Netzwerk im Gesamtgegenkopplungsweg vorgezogen wird. Abstimmen bereitet keine Schwierigkeiten, die Stabilität ist sehr gut und fast jede Bandbreite möglich. Universalkurven, die den Entwurf eines abgestimmten Verstärkers erleichtern, werden gegeben, und es ist beachtlich, dass nur wenige Abweichungen vom Entwurf der Röhrenschaltungen bestehen.

Eine einfache Ballontechnik zum Messen der Strahlungscharakteristik von Radioantennen von C. S. L. Keay und R. E. Gray

Zusammenfassung des Beitrages auf Seite 322-325

Die Strahlungscharakteristik eines Antennensystems für Betrieb bei Frequenzen von 20 . . . 500 MHz kann ohne Schwierigkeiten mit einem kleinen, in sich geschlossenen Sender und Dipol, die unter einem Pilotballon hängen, bestimmt werden. Der Ballon ist angepflockt und wird mittels vier Nylonseilen kontrolliert, die ihn zwingen, sich mit konstantem Radius um das zu untersuchende Antennensystem zu bewegen. Auf diese Weise kann schnell eine komplette Charakteristik mit 10% Genauigkeit bestimmt werden, und zwar manchmal in weniger als einer Stunde, je nach Betriebsfrequenz und den erforderlichen feineren Einzelheiten.

Analyse eines gegengekoppelten Verstärkers von C. D. West

Zusammenfassung des Beitrages auf Seite 326-329

In dem Beitrag wird die Analyse einer Impulsverstärkerschaltung behandelt. Die Schaltung, deren Verstärkung einen niedrigen Temperaturbeiwert hat, besteht aus einem gleichspannungsgekoppelten Differentialverstärker und einer Stufe in Emitterschaltung.

Für einen Verstärker mit einem Verstärkungsfaktor von 10 und einem Temperaturbeiwert von $50 \times 10^{-6}/°C$ werden Bauelementwerte gegeben. Es können mehrere Stufen hintereinandergeschaltet werden.

Ein einfacher, leistungsfähiger Serienkonstanthalter von D. J. Collins

Zusammenfassung des Beitrages auf Seite 330-331

Wenn konstante Gleichspannungen benötigt werden, so steht dem Konstrukteur eine breite Auswahl von Schaltungsverknüpfungen zur Verfügung. Seine Wahl wird durch die Netzgrundparameter, die Leistungsklasse und den zulässigen Kostenaufwand bestimmt. Der grösste Teil der Forderungen lässt sich leicht befriedigen, jedoch gibt es gelegentlich ein besonderes Problem, das eine Sonderlösung erfordert. Der Beitrag beschreibt eine Schaltung, die speziell als preisgünstige, kompakte Konstanthaltungsquelle für die Speisung unaufgeklebter Dehnungsmessstreifen in Druckgebern entworfen ist.