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Commentary

THE exhibition of electronic components and valves held in Paris during February of this year is the twenty-third exhibition of its kind to be organized by the French electronic industry since 1934.

For the third year in succession it has become international in character and of the 410 exhibitors present, no less than 60 were from countries abroad, represented principally by 21 exhibitors from America, 18 from West Germany and five from this country.

The *Salon International de la Piece Detachée et des Tubes Electroniques* is more than an exhibition of components and corresponds more to a combination of our own Radio and Electronic Component Manufacturers' (R.E.C.M.F.) Exhibition and the electronic instrument section of the Instruments, Electronics and Automation (I.E.A.) Exhibition with the Audio Fair thrown in for good measure.

As such, it is probably the largest exhibition of its kind in Western Europe and is at the same time an excellent showplace for the French electronic industry.

Comparisons both of size, efficiency and technical level with our own industry are inescapable but such comparisons are difficult to make and in any case are apt to be misleading since the two industries are not organized on the same lines.

One generalization which may be permissible is that with one or two notable exceptions the French electronic component industry is made up of a large number of small firms all specializing in a limited range of products. For example, there are, according to the *Fédération National des Industries Electroniques* (F.N.I.E.), some fifteen separate manufacturers of variable capacitors, twenty-seven of fixed capacitors and no less than thirty-eight of transformers.

In parenthesis the F.N.I.E. corresponds to our Radio Industry Council and is similarly subdivided into a number of organizations or associations. It has a membership of over 500 and represents about ninety per cent of the French electronic industry.

It should also be borne in mind that the French electronic industry is not burdened to the same extent as our own with matters of defence and is therefore concentrating its efforts largely on the more conventional outlets such as communication systems, industrial applications and the entertainment market for radio and television receivers. The manufacture of radio and television receivers represents nearly half the total output of the industry and while figures are not yet available for 1959, it has been estimated that the total output will have been of the order of half a million television receivers and 1½ million radio sets amounting to a turnover of some £60M. To this figure may be added a further £13M. for audio frequency equipment, giving a total turnover for the past year of about £180M.

The French electronic industry is thus considerably smaller than that in this country which on a census of production had a gross output last year of some £470M; neither does its export figures for the same period compare with that of £28M. for the capital goods side alone of the British electronic industry. But it should be emphasized that the French electronic industry manufactures primarily for the home market and is not traditionally one of the French exporting industries.

But if it is smaller than the British industry it is by no means lagging behind from the technical standpoint and there is every evidence of a live and virile industry.

* * *

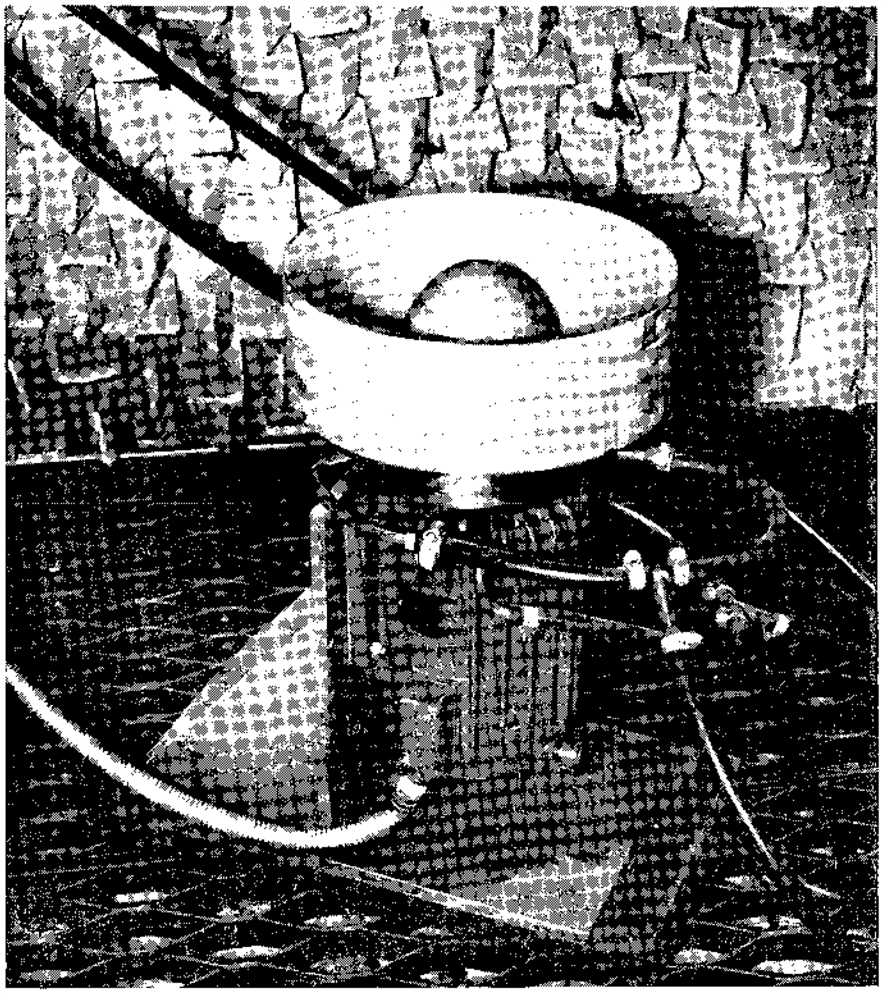
Before the war, the electronic industry of this country was mainly occupied with the production of radio and television equipment for the entertainment market. In the years since the war the capital goods and industrial electronics side of the industry has assumed an ever-growing importance and we therefore welcome the announcement of the formation of the Electronics Industry Council.

This new Council, which has been formed by the Electronic Engineering Association who with the Radio and Electronic Component Manufacturers' Federation and a number of other bodies have been discussing the matter for some time, will be concerned with electronic instruments, sound and television transmitters, radio communication equipment, radar and radio navigational aids, computers, industrial electronic control equipment and industrial television and the electronic components used therein.

Its constitution will provide for the adherence of associations or federations of manufacturers concerned wholly or partly in the manufacture in the United Kingdom of electronic components, apparatus and equipment except for those used in the broadcast radio and television receiving industry and for public telephone services.

The new Council, being in the capital goods field, is complementary to the Radio Industry Council which for more than a year now has been dealing only with the associations whose members make broadcasting receivers, or the components and valves for them and this sharing of the responsibilities previously carried by the R.I.C. should make for a more sound and efficient organization of the industry as a whole.

High Intensity Swept Frequency Acoustic Test Equipment for the 1 to 60kc/s range



By D. J. Birchall*

This article describes the control gear for a high-intensity, swept-frequency siren and the instrumentation designed for measuring sound pressure levels, and comprising low noise pre-amplifiers, band-pass selective amplifiers, and logarithmic amplifiers.

This siren is housed in an anechoic chamber, and the whole equipment is used to carry out acoustic tests of various descriptions.

The main feature of interest is the method of employing thermistors as the frequency controlling elements whereby the siren and selective amplifiers track with a frequency reference. The selective amplifiers are thereby tuned to the fundamental frequency of the sound field, and track together when the frequency reference is swept, thus providing a system for automatic swept-frequency monitoring on narrow-band channels.

Frequency range of the equipment is 1 to 60kc/s, covered in one sweep without range switching, at a maximum rate of two octaves/minute.

Threshold sensitivity for sound measurements is about 70dB, and dynamic range of the logarithmic amplifiers is 60dB.

(Voir page 260 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 265)

A HIGH-INTENSITY acoustic environment may affect the functioning of certain types of equipment. This susceptibility may be investigated by conducting tests under a simulated environment, in this case, comprising a swept-frequency sound field. Results will be obtained in the form of a frequency response of cause and effect, and remedial measures may be analysed similarly.

The equipment described was designed to test an electronic system susceptible to acoustic noise over a 1 to 60kc/s bandwidth.

The high sound levels required over the frequency range limited the choice of sound source to that of a siren. This is driven by a high-speed three-phase induction motor, and the sound field frequency is a multiple of motor speed. Power for the motor is obtained from three Bryan Savage v.l.f. 1kVA amplifiers, whose inputs, in turn, are obtained from a three-phase oscillator which acts as the means for speed control.

An inherently bad feature of the siren is the high harmonic content in its sound field; if data in the form of a frequency response is to be obtained filtering of the harmonics must be resorted to.

The necessity for inclusion of frequency selective amplifiers in the monitoring equipment led to the development of a three-phase oscillator and selective amplifiers which track together, enabling sound levels to be recorded as the siren sweeps. Six sets of monitoring amplifiers are provided. These each comprise the three units mentioned in the summary.

The microphones used have as the pressure-sensitive element a barium titanate ceramic disk. Frequency response is flat to 20kc/s, above which various resonances and anti-resonances give rise to variations about the normal of the order of ± 10dB.

A feature of the pre-amplifiers that is worthy of note is the method of feedback employed, analogous to the feedback principle of summing amplifiers, whereby the microphone signal is amplified by a factor which is the ratio of the microphone capacitance to a feedback capacitance

The above photograph shows the siren and its induction motor drive

* *Hunting Engineering, Luton (formerly with English Electric Aviation Ltd.)*

(C_m/C_1). The capacitance of the connecting lead between microphone and pre-amplifier has little loading effect on sensitivity, although it may be several times that of the microphone itself.

Intended for calibration purposes, but also useful as a signal source for a loudspeaker (in place of the siren), a 1 to 60kc/s constant amplitude oscillator is provided. This is also designed on the thermistor principle and tracks to the frequency reference.

The frequency reference is a direct voltage derived from either a manually variable or a motor-driven potentiometer. Frequency is proportional to voltage, and the potentiometers have an exponential law such that log f is proportional to angular rotation.

An electronic servo loop is used on each oscillator and selective amplifier to reduce the individual alignment errors. External supplies for the servos comprise a 25c/s constant amplitude signal and a 10c/s push-pull signal, generated by a 25c/s and 10c/s oscillator.

The equipment outlined above, together with power supplies, is housed in a console adjacent to the anechoic chamber. Used in conjunction with a suitable recorder, the equipment enables frequency plots of sound pressure levels to be obtained automatically.

Frequency Alignment System for the Oscillators and Selective Amplifiers

DEVELOPMENT OF THE SYSTEM

Mention has been made of the high distortion level of the sound field. Distortion factors in excess of 100 per cent have been noted, which, together with high background noise levels, necessitated the incorporation of narrow-band amplifiers in the monitoring channels. The choice of Q factor for the amplifiers is dependent upon the probable maximum value for distortion factor, which in turn, depends upon the nature of the acoustic investigations that are envisaged—distortion factor in an acoustically resonant system would be magnified.

The decision to adopt an automatic tracking system for the selective amplifiers, with the necessity to make allowance for tracking error, made the choice of Q a compromise. A Q value of 6 was decided on, which reduces the distortion factor by about 20dB, and to date this choice appears quite adequate.

An alternative to this type of filter would be one based on the superheterodyne principle, the filtered output being at a suitable carrier frequency. The present system was pursued with one or two further applications in mind, where the requirement is simply for a reduction in harmonic content and noise level of a signal. An example of the latter type would be a filter required to reduce distortion in the output signal of a modulator valve used to control acceleration level of a vibration generator.

Control of frequency is effected by using indirectly heated thermistors as the resistance elements of resistance-capacitance selective amplifiers and oscillators, frequency variation being achieved by the temperature variation of the thermistors.

Thermistors are resistance elements having a high negative temperature coefficient. The indirectly heated types are wound with a small heater, the whole assembly being sealed in an evacuated glass envelope. Wide variation of resistance is possible by variation of heater current, the 60:1 resistance change implicit in a six octave frequency coverage being easily achieved.

The heater current/resistance characteristics of all thermistors in a unit are matched, and an extra matched thermistor is included from which is derived the frequency of the unit in terms of a voltage (ω). Heaters are connected in series and supplied from a heater power supply whose output is controlled by the error between ω and the frequency reference, the unit aligning itself to the frequency reference.

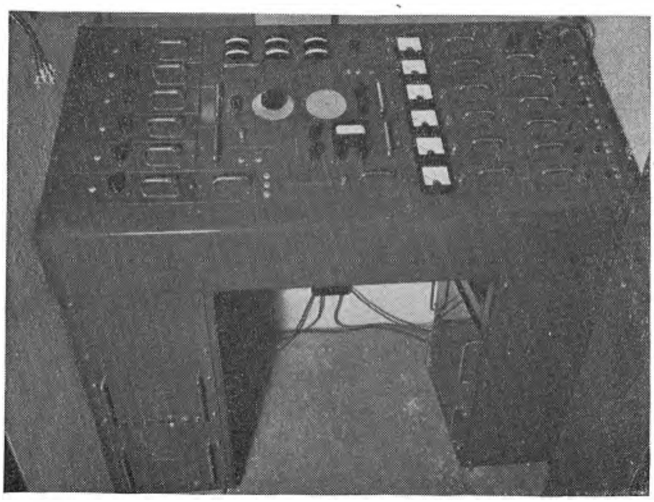
The frequency of the oscillators and selective amplifiers is a function $(1/R \cdot n/C)$ of the resistance capacitance elements composing the oscillator or filter sections.

That the thermistors are matched, and their heaters are in series, implies that the thermistor resistances are nominally equal to R, i.e., that within the limits of matching accuracy a unit will be at its nominal frequency and that this frequency may be derived from the spare thermistor of the set.

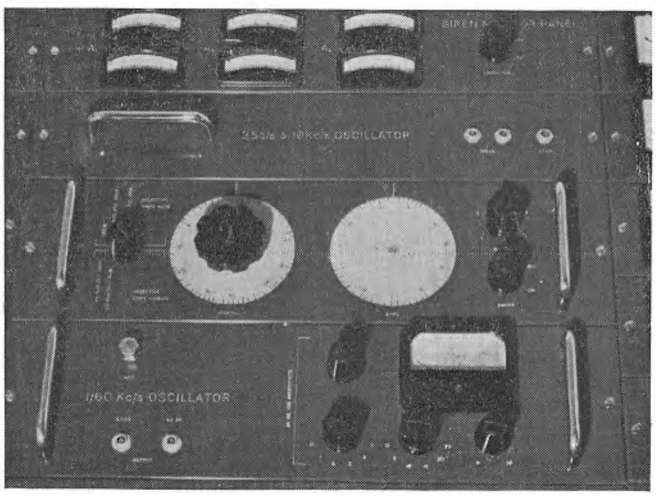
The unit frequency is derived by the following connexion of the reference thermistor (Fig. 1):

Control of frequency is obtained by deriving the error between ω and the frequency reference (ω_r), and after amplification using it to control the heater current to the thermistors. In the absence of an error, this current assumes

The completed acoustic testing equipment. Low noise amplifiers are on the extreme left. The logarithmic and selective amplifiers occupy the two right-hand bays



Close-up of the centre-left bay. The frequency reference voltage is generated by one of two potentiometers mounted behind the panel of the lower centre unit



a maximum value I_m, hence the unit will be at its maximum frequency.

The system diagram is shown on Fig. 2.

The loop equation is:

$$\omega = \beta\,[i_M - A_2\,(\omega - \omega_r)]$$

$$\omega\,(1/\beta + A_2) = I_m + A_2\omega_r$$

$$\omega = \frac{\beta\,I_m}{1 + \beta\,A_2} + \frac{\omega_r}{1 + 1/(\beta A_2)}$$

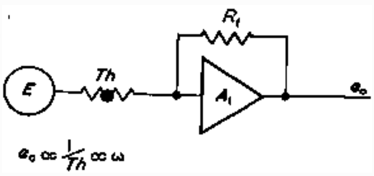


Fig. 1. Frequency voltage convertor

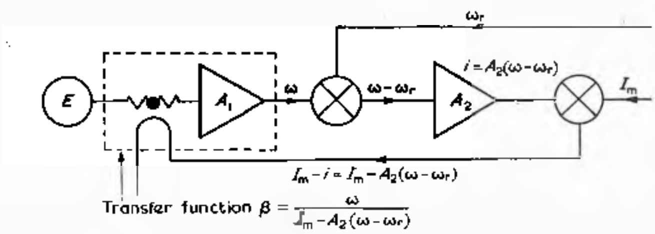


Fig. 2. Frequency alignment system

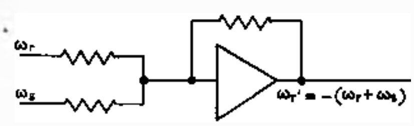


Fig. 3. Summing amplifier for slip correction

The frequency maximum corresponding to I_m, ω_m is equal to βI_m. βA_2 is the loop gain of the servo, G. Hence:

$$\omega = \frac{\omega_m}{1 + G} + \frac{\omega_r}{1 + 1/G}$$

$$\omega \simeq \omega_r + (\omega_m - \omega_r)/G$$

ω is higher than the reference frequency by an amount $\omega_m - \omega_r/G$. This error assumes a maximum value for $\omega_r \to 0$.

The error at the low frequency end is about 0·2 octave high, and correction is obtained by means of a pre-set potentiometer which shunts the reference thermistor.

The servo is an a.c. type for reasons of simplicity. E is a 25c/s voltage of controlled peak amplitude. ω_r is a direct voltage. In the absence of an error signal, ω would tend to a value in excess of ω_r. Comparison between ω and ω_r is made using a diode, peak values of ω exceeding ω_r composing the error.

The error, in the form of a train of pulses, is amplified, rectified, and applied as the control-grid bias to a pair of push-pull variable-mu valves. In the absence of bias, I_m is supplied by these valves to the heaters. Bias reduces I_m, and the bias may be considered as a current i opposing I_m. Error rate damping and integral control are incorporated after the bias rectifier to stabilize the loop.

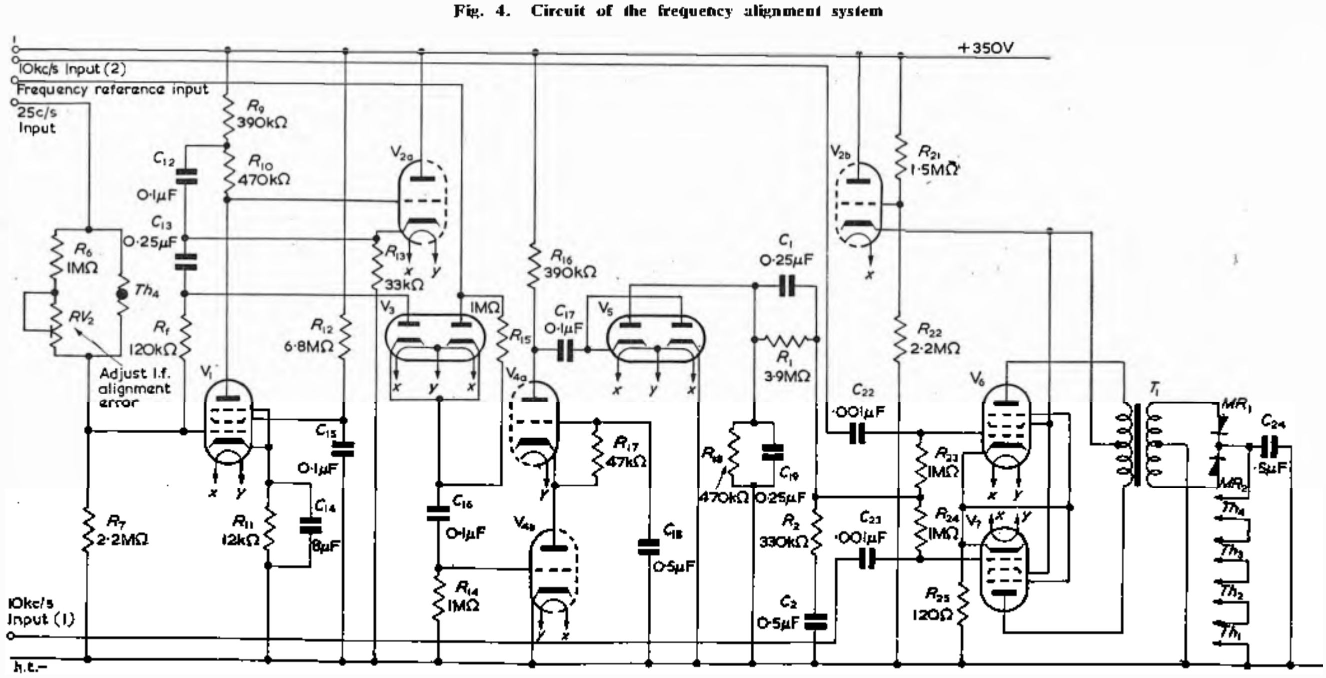
The sound field is about 500c/s below ω_r due to slip of the induction motor drive to the siren. Compensation is provided by adding to ω_r a voltage (ω_s) such that the reference (ω_r) for the three-phase oscillator is about 500c/s higher than that for the rest of the frequency-variable units (Fig. 3).

The servo control of frequency outlined above is applied to the three-phase oscillator, a 1 to 60kc/s oscillator, and the frequency selective amplifiers. A circuit description follows.

Circuit Description of the System

Fig. 4 is a circuit diagram of the servo system described in the first half of this section. Amplifier A_1, used with the reference thermistor Th_4 to derive frequency in terms of an analogue voltage, comprises valves V_1 and V_{2a}. This amplifier is of the 'bootstrap' variety and has an open-loop gain of 60dB with phase reversal. Output is linear up to levels of about 30V peak. The reference thermistor varies in value from 240kΩ (corresponding to 1kc/s) to 4kΩ (60kc/s), which with a feedback resistor of 120kΩ and an input of 1V peak, produces an output within the range 0·5V to 30V

Fig. 4. Circuit of the frequency alignment system



peak. The voltage error between A_1 output and the direct voltage analogous to the reference frequency is extracted by comparator diode V_3.

V_3 is a double diode, the halves being connected in series opposition. This cancels to some degree the effect of contact potential. Otherwise a large and random error would be introduced into the loop at the low frequency end of the range, where the levels being compared are of the half volt order.

The contact potential at the anode determines the conduction point of a diode, and is a function of heater voltage. In the case of the CV140 diodes used in the circuit, a 10 per cent heater voltage variation can produce a contact potential variation of about 300mV. The error can be reduced by about 5 : 1 simply by the connexion of a similar diode in series opposition.

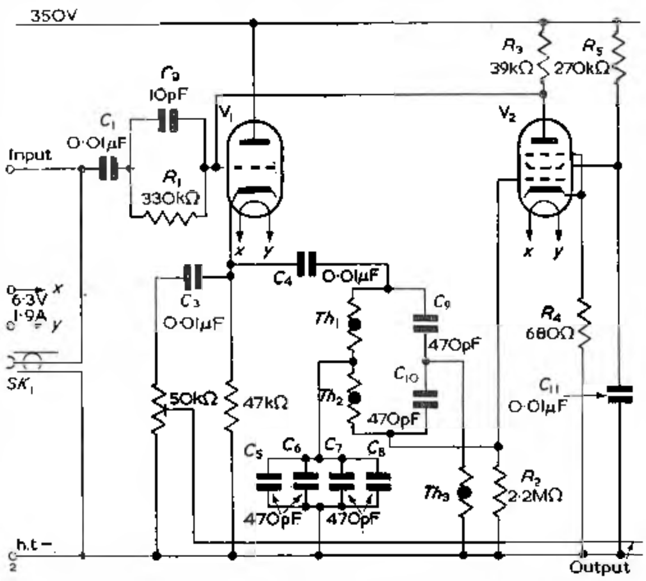


Fig. 5. Frequency selective amplifier

Error voltage from the comparator consists of a train of pulses. These are routed to V_4, a cascode connected double triode. V_4 has no auto-bias resistors. The contact potential effect mentioned previously provides a bias of about 500mV. Characteristic to amplifiers of this type are a low input impedance (in this case about 200kΩ), and of course the small bias value limits their use to signals having small positive peak amplitude. In this case the input level is about 100mV, maximum, from the comparator. The gain of V_4 is about 50dB.

V_4 is followed by a rectifier (V_5) and smoothing network, the error pulses finishing as a negative voltage which is applied as control grid bias to the push-pull variable-mu pentodes V_6 and V_7. The signal input to these valves is derived from the 10kc/s oscillator, and the output is rectified and provides heater current for the four thermistors. Bias controls the heater current level. In the absence of the error bias, output is a maximum I_m. The application of bias reduces I_m, i.e. the bias is converted to a current i opposing I_m. A_2 $(i/\omega - \omega_r)$ is therefore the gain from the comparator to the output of V_6 and V_7.

A network is inserted after V_5 to stabilize the loop. It comprises the capacitors and resistors C_1 R_1 (error rate damping, bias $\propto d\,(\omega - \omega_r)/dt$), and C_2 R_2 (integral control, bias $\int^T (\omega - \omega_r)\,dt$).

In the case of the three-phase oscillator the frequency reference is derived from a summing amplifier and is inverted in polarity. Modifications to the control system are therefore entailed and involve the reversing of diode V_3 and the reconnexion of V_4 in the form of a two stage cascade amplifier, a phase reversal not being required.

Further Topics Relating to the Frequency Alignment System

Thermistors

The thermistor sets used in the various oscillators and selective amplifiers are of S.T.C. manufacture, type B·65. Sets were selected to a tolerance of ±3 per cent of the mean of the set. Nominal resistance at an ambient temperature of 20°C is 500kΩ dropping to 4kΩ at a heater current of 15mA. The decrease is logarithmic. Provision must be made to ensure that there are no large temperature gradients between the individual thermistors of a set, and those used in the equipment are mounted in a small aluminium block.

Deviation from the required tolerance at extremes of the resistance range can be corrected by the addition of resistors either in parallel (at the high resistance end) or in series with the offending thermistors.

A thermistor having a constant percentage error can be employed as the reference thermistor Th_4 simply by an adjustment to the value of R_f.

Thermistor stability is of a high order[1], certainly the equipment has maintained its accuracy throughout its operational life since late 1958.

25c/s Oscillator

Peak amplitude of the 25c/s signal is a factor controlling frequency accuracy of the servos, any deviation from a peak value of 1V being reflected, *pro rata*, as a frequency deviation.

In the case of the selective amplifier and the 1 to 60kc/s oscillator, negative peak amplitude of the 25c/s oscillator signal is used, after amplification, as the internal frequency reference. Positive peak amplitude is used likewise in the three-phase oscillator.

The load supplied by the 25c/s oscillator comprises seven thermistors in parallel. The load is a function of frequency and drops to a minimum of about 550Ω.

Both a low output impedance and an absence of asymmetric distortion are therefore implicit in the maintenance of a close degree of alignment between the various servos and the frequency reference, and between the sound field and the selective amplifiers.

The Frequency Reference Voltage

A manually variable and a swept frequency reference were incorporated into the equipment. These each comprise a continuous rotation exponential low potentiometer. The swept frequency reference has ganged to it a frequency marker switch and is driven by a fractional horsepower d.c. motor through a 10 000 : 1 reduction gear box. The armature voltage is varied to provide a range of sweep speeds between ¼ and 2 octaves/minute. Various combinations of manual and swept reference voltages to the various units enable alignment error and frequency response checks to be made on the selective amplifiers.

Selective Amplifiers, Three-Phase Oscillators, and 1 to 60kc/s Oscillator

These units each comprise an appropriate thermistor-capacitance network with the thermistor heaters linked in series with the reference thermistor and connected to the thermistor heater supply.

Frequency Selective Amplifiers

Low Q selective amplifier design presents no difficult problems[2]. That developed for this equipment is shown in Fig. 5. A twin-T filter and the feedback amplifier (V_2) are connected between cathode and grid of cathode-follower V_1. Input is also to the grid of V_1, output from the cathode.

Maximum transmission, at the rejection frequency of the twin-T filter, is R_3/R_1+R_3. Outside the pass-band the transmission falls to an asymptotic value $R_3/(A+1)\,R_1+R_3$ of the input, A being the gain of V_2.

The pass-band gain is flat to within 1 to 1½dB over the whole 6 octave range of the equipment. The input level varies from 50mV to 50V peak-to-peak. Rejection an octave away from the pass-band centre frequency is about 20dB.

Oscillators

The 1 to 60kc/s oscillator comprises a Wien bridge and maintaining amplifier. The selective network in the case of the three-phase oscillator consists of three single thermistor-capacitance phase-shift networks separated by cathode-followers.

The maintaining amplifier design is a problem relating to the frequency coverage of the particular oscillator under consideration, and usually a variable gain characteristic controlled by the amplitude of oscillation is required. From the standpoint of simplicity and freedom from distortion, the type of maintaining amplifier using a thermistor as a feedback resistor to control gain is superior to types where gain is controlled by means of a bias voltage variation. This latter type was developed here, and performs satisfactorily, but is more complex than a thermistor gain control system and suffers to a certain degree from non-linearity, the distortion level being in the region of 10 per cent at certain parts of the frequency range.

It may be mentioned that a variable frequency induction motor requires a voltage proportional to frequency. In consequence differentiators are incorporated into the three-phase oscillator.

Low-Noise Amplifiers (Fig. 6)

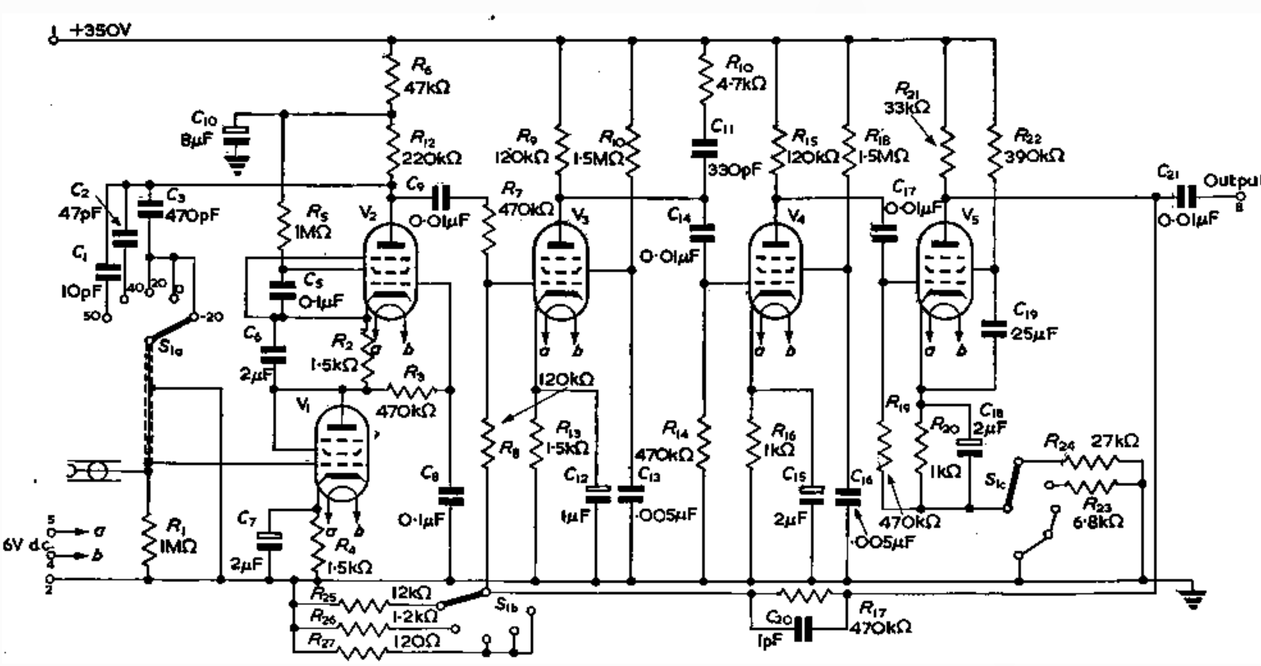
The dynamic range of the monitoring amplifier chain is 60dB and threshold sensitivity level is determined by the microphone sensitivity and gain of the low noise pre-amplifiers. The pre-amplifier gain is variable in fixed steps (one 10dB and three 20dB steps) such that the threshold sensitivity is adjustable over a range of sound intensities of from 65 to 135dB with respect to ·002dyn/cm².

A factor affecting the sensitivity of systems employing high-impedance transducers is the attenuation of transducer output due to its loading by the measuring circuit. In this particular case, the transducers are capacitive, but the loading effect of connecting lead capacitance, which is about three times that of the microphone itself, has been eliminated. The microphone self-capacitance determines pre-amplifier gain ($G=K.\ C_m/C_f$, where C_m is microphone capacitance and C_f is a feedback capacitor internal in the pre-amplifier), cable capacitance loading being a second order effect which may be rendered negligible.

The microphones used in the equipment each comprise a barium titanate pressure-sensitive ceramic disk suitably mounted in a cylindrical container which provides screening and a heavy mass backing. The assembly is about 1in in diameter by 1½in long. The first axial resonance of the ceramic disks themselves occur at about 80c/s, but constructional factors involved in the complete microphone assembly reduces the limit of linear operation to 20kc/s. Above this the sensitivity varies by as much as ±10dB about the mean. Microphone sensitivity up to 20kc/s is approximately 26dB with respect to 1μV/dyn/cm², and the capacitance of the ceramic disk is about 80pF.

In the main, the microphones are used to measure pressure levels at surfaces rather than for free field measurements, where their large size would lead to pressure doubling effects at quite low frequencies. Calibration is made by a sensitivity comparison with a Brüel and Kjær capacitor microphone in free field at 1kc/s, followed by a measurement of frequency response using a Brüel and Kjær electrostatic actuator.

Fig. 6. Low noise amplifier



Attention has been drawn to a feature of the pre-amplifier whereby microphone cable capacitance is reduced to a second-order effect with regard to sensitivity. The means by which this is achieved will be shown in the next few paragraphs. It is an adaption of the feedback technique used for summing amplifiers, although the application in this context is thought to be new.

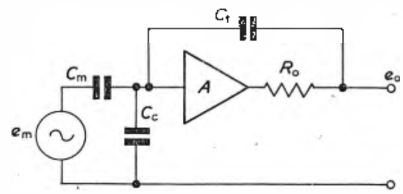


Fig. 7. Feedback system of the low noise amplifier

Microphone and pre-amplifier are connected as shown in Fig. 7.

e_m and C_m are microphone open circuit sensitivity and self-capacitance.

C_c is capacitance of the cable connecting microphone to pre-amplifier.

$-A$ and R_o refer to amplifier gain and output impedance. C_f is a feedback capacitor.

The effect of input resistance may be considered by considering C_c complex.

Gain of the system (G) equals $1/a$ of the system (a) matrix

$$\begin{bmatrix} a & b \\ c & d \end{bmatrix}$$

the system $[a]$ matrix is the product of the $[a]$ matrices:

$$[a_{em}, C_c] \times [a_{\text{pre-amp}}]$$

$$[a_{em}, C_c] = \begin{bmatrix} 1 + Z_m Y_c & Z_m \\ Y_c & 1 \end{bmatrix}$$

$$[a_{\text{pre-amp}}] = \begin{bmatrix} Y_f + Y_o & 1 \\ Y_f Y_o (A+1) & Y_f \end{bmatrix} \times \frac{1}{Y_f - A Y_o}$$

$$(1/a) = G = \frac{Y_f - A Y_o}{(Y_f + Y_o)(1 + Z_m Y_c) + Z_m Y_f Y_o (1 + A)}$$

Y refers to admittance, and Z to impedance.

For $A \to \infty$, $G = -(1/Z_m Y_f)$, or $-(C_m/C_f)$

Let this value of gain be G'.

Then:

$$G = \frac{-G' \left(1 - (Y_f/AY_o)\right)}{1 + \left((Y_f/AY_o) + (1/A)\right)(1 + Z_m Y_c)\, G'}$$

For $G \to G'$

The error $(G' - G)/G'$ between the actual and hypothetical gain values equals (approximately)

$$\Delta = \frac{j\omega C_f R_o}{A} + (G'/A)(1 + j\omega C_f R_o)\left(1 + (C_c/C_m)\right)$$

Gain Error at Low Frequencies

For $\omega = 0$,

$$\Delta \simeq \frac{C_m + C_c}{AC_f}$$

The amplifier input resistance may be lumped with C_c, Then:

$$\Delta \simeq \frac{C_m + C_c}{AC_f} + \frac{1}{j\omega R_{in} C_f \, . \, A}$$

The former part of the expression may be considered to be the mid-band error, and the latter an additional error at low frequencies due to input resistance R_{in}.

High-Frequency Error

For $\omega C_f R_o \gg 1$

$$\Delta \simeq \frac{j\omega R_o (C_f + C_m + C_c)}{A}$$

The actual pre-amplifier comprises two feedback amplifiers connected in cascade, that amplifying the microphone signal being based on the system analysed above, and having a gain of 200, output impedance of 200kΩ, and input resistance of 1MΩ. Feedback capacitor C_f is in fact one of three values selected by S_{1a}, providing a 30dB gain variation.

Changing C_f from 470pF to 47pF and then to 10pF produces gain variations of 20dB and then 10dB. The latter value for C_f is about 15 per cent lower than that computed from the hypothetical value for G, and takes into account the 15 per cent error that would ensue at the mid-band assuming values of 80 and 250pF for C_m and C_c respectively.

The low-frequency error is in quadrature to G', the difference between G and G' for a low-frequency error of 0·3 being approximately 10 per cent. The l.f. 1dB point, then equals $1/(0{\cdot}3A\, R_{in} C_f)$, and for $C_f = 10$pF. This occurs at 1·5c/s.

The high-frequency error is also an error in quadrature to G'. It is a maximum for $C_f = 470$pF, and on this range is about 10 per cent at 60kc/s.

Logarithmic Amplifiers

A logarithmic amplifier based on the circuit shown in Fig. 8 has been developed for inclusion in the monitoring channels. The large increase in dynamic range that accrues allows most frequency plots of sound levels to be recorded without range switching.

Referring to Fig. 8, diode parameters govern the choice of voltage increment at which the loading resistors are switched. In turn, the selection of diode is controlled by the characteristics of the voltage source E_{in}.

Assuming a value of switching increment e volts, and

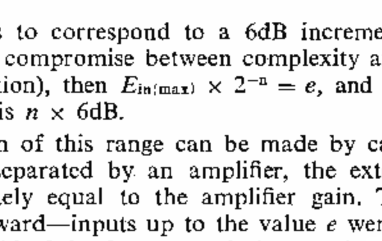


Fig. 8. Basic linear/logarithmic network

that this is to correspond to a 6dB increment of E_{in} (a reasonable compromise between complexity and degree of approximation), then $E_{in(max)} \times 2^{-n} = e$, and the logarithmic range is $n \times$ 6dB.

Extension of this range can be made by cascading two networks separated by an amplifier, the extension being approximately equal to the amplifier gain. This is quite straightforward—inputs up to the value e were not within the threshold of the first network, but are increased by the gain of the amplifier and come within the threshold of the second network by this gain factor. Peak input to the amplifier is of the order $(n + 1)\, e$ rather than $e \times 2^n$, hence amplifier overload problems inherent to a wide dynamic range are reduced.

Design of the logarithmic amplifier starts with design of the second logarithmic network, followed by the calculation of its input/output characteristic over the complete range. The input logarithmic network is now designed to match the input characteristic of the second network.

A few simple rules are necessary in the selection of the network diodes, and in the network design itself. Choice of diode is governed by the output resistance of the voltage source. The signal level may be below the operating threshold of the logarithmic network, but the diode leakage current will be drawn from the input and may cause unwanted attenuation. Germanium diodes were unsuited in this respect in the input network, but were incorporated in the second logarithmic network driven from the d.c. amplifier.

In the conduction direction, diodes are characterized by the following approximate formula. $e_d = e_b + Ri$, where e_d is the diode voltage drop for a forward current i, R is the forward resistance, and e_b is the junction voltage, which is about 200mV for germanium junction diodes and 500mV for silicon junction diodes, but is subject to wide variation between samples and to a variation with temperature. A switching increment e of about five times the normal barrier voltage will ensure a reasonable stability against the variations mentioned.

The logarithmic amplifier circuit is shown in Fig. 9. The input logarithmic network compresses the signal over the upper 30dB of the range. The output from this network is amplified by $6\frac{2}{3}$ by V_1 and V_2 and applied to the second network. The lower 12dB of the range is left linear.

The second network is followed by an output amplifier V_3 and V_4 having unity gain, whose zero is offset so that the 0dB point coincides with zero output. A monitoring meter is connected to the output of this amplifier for calibration purposes.

A table of voltage levels at various points through the amplifier is appended (Table 1). A gap in the output of the second logarithmic network larger than the nominal 1V/6dB increment will be noticed between the 6 and 12dB levels. This offsets the non-linear characteristic of the input demodulator at the low levels.

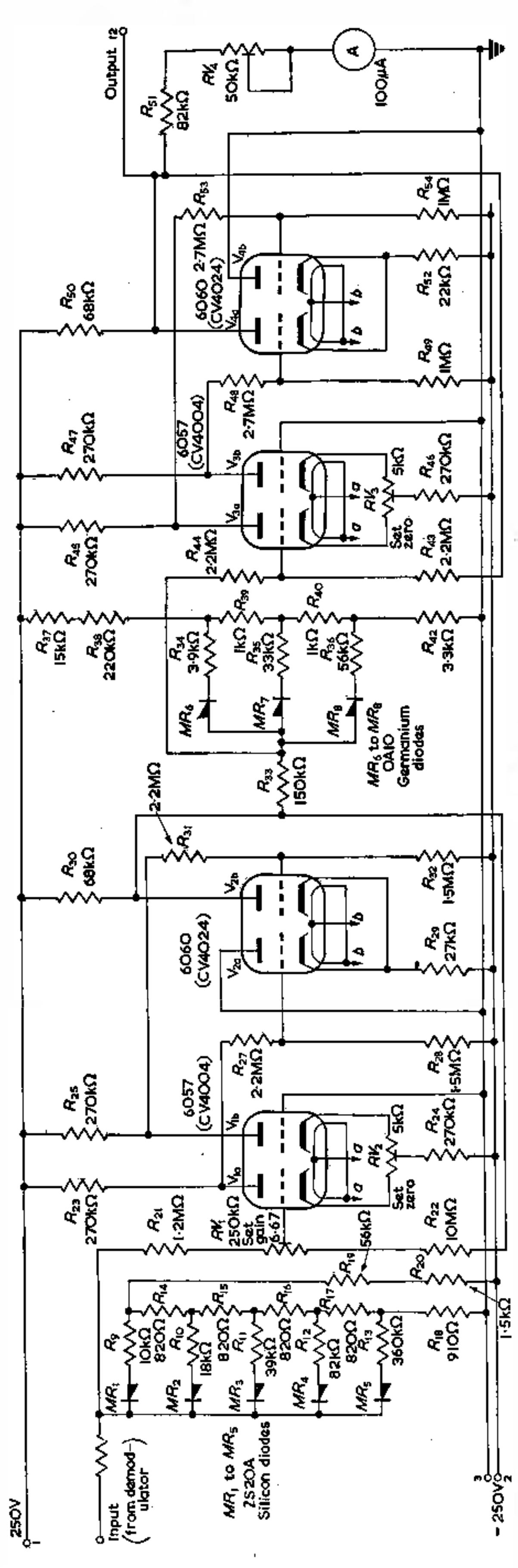


Fig. 9. Logarithmic amplifier

TABLE I

Voltage levels at reference points within the logarithmic amplifier

INPUT TO 1ST DIODE NETWORK (V)	OUTPUT TO 1ST DIODE NETWORK e_{01} (V)	$e_{01}.GV_1—V_2$ (V)	OUTPUT FROM 2ND DIODE NETWORK (V)	RELATIVE LEVEL
0·125	0·125	0.835	0·835	0
0·25	0·25	1·67	1·67	6
0·5	0·5	3·34	3·34	12
1	1	6·67	4·34	18
2	2	13·34	5·34	24
4	4	26.68	6·34	30
8	7·2	48	7·34	36
16	10·4	69·34	8·34	42
32	13·6	90·68	9·34	48
64	16·8	102	10·34	54
128	20	113·34	11·34	60

Drift of the amplifier is negligible after a 20 minute warming up period, and calibration stability also is of a high order.

Acknowledgments

The author would like to thank the Directors of the English Electric Aviation Co. Ltd for permission to publish this article. He is also indebted to Mr. G. D. Oliver, now with De Havilland Propellers Ltd, for his encouragement during the course of the work.

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High Speed Track Selection for a Magnetic Drum Store

Andrew D. Booth*, D.Sc., Ph.D., F.Inst.P., M.Brit.I.R.E.

A short account is given of the methods which have been adopted for the selection of tracks on a magnetic drum store. These are followed by a discussion of two conventional transistor head switches and finally by a description of a new head switch which uses a single symmetrical transistor and which has the advantage of not injecting a large transient into head and amplifier at the instant of switching.

(Voir page 260 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 265)

THE earliest magnetic drum storage system[1] was intended for use in conjunction with a relay computing system and hence it was natural that relays should be used to perform the operation of switching between the various tracks and the electronic amplifier. Early magnetic drum recording and reading heads had extremely low impedances and thus required a relatively expensive matching transformer either to lift the output signals to a satisfactory voltage level for input to an electronic amplifier or to transform the small currents available from normal thermionic valves to the ampere or so required by the head for writing. Thus the use of relays between heads and transformer was almost mandatory from the viewpoint of cost and complexity.

Quite apart from their unique property of providing a nearly zero resistance path for currents of either sign, relays have two additional virtues: first that their noise level is extremely low and second that they can be connected in series and thus enable the operations of decoding a selection indication and setting up the required path to be combined in a single unit.

A typical head selector of this type is shown schematically in Fig. 1. Because a number of extant machines still employ switching of this type, it is worth making a few observations on points of note in the design. The first is the choice of the relays themselves. If speed is not important, as when the magnetic drum is a mere backing store, the P.O. 3 000 type relay offers the advantages of low cost, mechanical robustness and reliability, and duplication of contacts. On the other hand, in this speed range (20msec), the T.M.C. cross bar selector provides up to 40 make contacts, each duplicated, on a single unit and has the additional advantage of a selector mechanism which removes the necessity of having many contacts in series.

When speed is required neither of the above devices is satisfactory and use must be made either of a polarized relay or of a Siemens high speed relay. Both of these relays easily achieve a transfer time of 1msec and, taking account of the fact that no current is broken by the relay and that it has to hold off only a few volts, it has proved possible to obtain bounce-free transit times of 0·25msec with the Siemens relay. This is only possible if the gap is reduced to ·0005in which is really too small for stability with the normal construction of the contact system. Daily maintenance checks are necessary and adjustment has to be made with a cathode-ray oscilloscope.

The second point which is worthy of mention is that the tree itself should be of balanced construction so that no change in power supply current results from a change in relay configuration. This presents no difficulty except at the apex (Relay 1, Fig. 1) and our practice has been to replace this relay by a pair.

Suitable contact materials are platinum for the high speed relays and silver or rhodium for P.O. 3 000 or cross bar selectors.

While the author has experienced little trouble with several relay head switches, the same good fortune does not seem to have attended the efforts of some other designers. This has lead to numerous attempts to replace the relays by a simple and inexpensive all-electronic switch. It is not

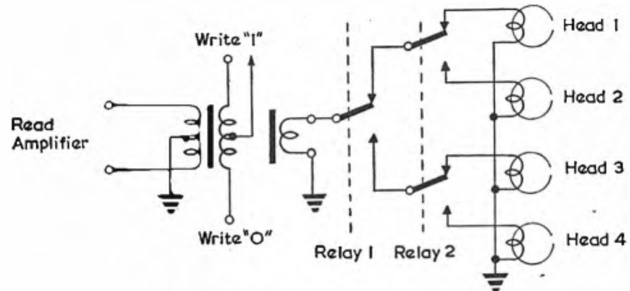

Fig. 1. Relay selector and decoder

worth detailing all of the early efforts in this direction, some of which are described in a thesis by G. J. Herring[2]. All electronic switches require either that each head has a separate matching transformer, or that special high impedance heads are wound. A high impedance head is difficult to screen and is apt to be bulky. For this reason the author prefers the individual matching transformer which, by suitable production techniques, can be manufactured quite cheaply.

Whichever alternative is adopted, however, it is important to minimize the transient which the act of switching introduces into the amplifier-transformer-switch-transformer-head system. In particular paralysis of the playback amplifier may be difficult to overcome in the 20 to 100μsec which are usually available between a track switch and the next reading operation.

Restriction of the low frequency response of the amplifier will usually reduce amplifier paralysis to acceptable limits, but experiments have shown that the transformers themselves are insensitive to small inputs for some time after heavy saturation of their cores. It is worth pointing out that this effect is quite as objectionable with ferrites as with metallic cores and that it obtrudes after most return-to-zero recording operations.

One of the first economical electronic head switch systems which, at the same time, avoids transformer saturation is that designed by R. Bird[3], and shown in Fig. 2. Here the switch elements are pairs of diodes, D_1 D_2, associated

* *Birkbeck College, University of London.*

with the matching transformers $T_1 \ldots T_n$. The selection mechanism is as follows: the triode section S_r associated with the head to be selected has its grid at earth potential, all of the other triodes have grids held well below cut-off voltage. Suppose that S_1 is selected, conduction will reduce the anode voltage, set by the power supply V_2 and the high value resistor R, to a value slightly less than that of the main transformer, M, centre-tap, V_1. It follows that the diodes D_1, D_2, associated with matching transformer T_1 will conduct and that signals appearing at T_1 from the associated head will be transmitted to M and thence to the reading amplifier. Notice that, because of the symmetry of the circuit, the act of selection sends equal currents to both halves of T_1 and M so that magnetization of the cores does not occur and transients are minimized. The act of switching does not cause spurious writing on the drum.

All diodes D_1, D_2 other than those associated with the selected transformer, T_1, are back biased by V_2 and R so that signals are not received from non-selected heads.

The operation of writing involves an impulse on one or other of the inputs to M, this signal appears as a positive

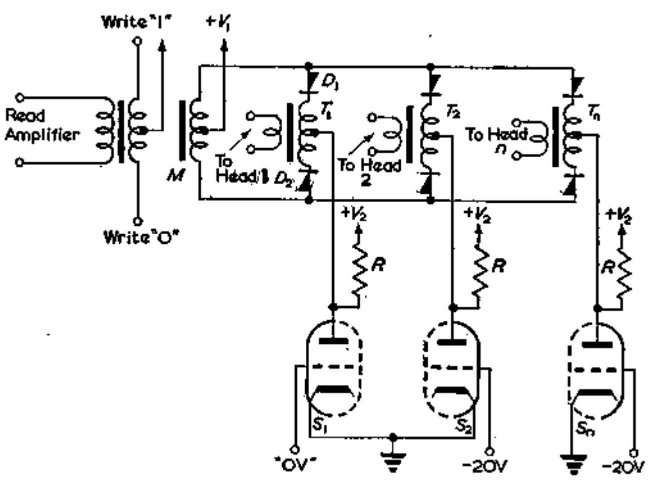

Fig. 2. Bird's electronic head switch

pulse on D_1 and an equal negative pulse on D_2 (or vice versa). D_1 conducts and the appropriate winding of T_1 transmits the current to the anode of S_1 where it flows to earth.

Naturally, since writing must not occur on non-selected heads, it is necessary that the bias on $S_2 \ldots S_n$ is adequate to hold the valves cut off when the write signal appears on their anodes.

A number of switches of this type have now been constructed and their vices are well known. The chief is that, because the diodes D_1 D_2 associated with non-selected heads are subjected to a high back bias, they tend to be noisy and some selection is necessary. Semiconductors are generally most suitable although the hot cathode type 6AL5 has proved satisfactory.

The second vice derives from the use of hot-cathode triodes for $S_1 \ldots S_n$. Since the current carrying capacity of these valves is only 10 to 20mA, a larger number of turns ($\sim$ 200) is needed on the transformers M and $T_1 \ldots T_n$. This high impedance makes the connecting leads liable to pick-up and careful screening is necessary.

Many of these defects are removed in the transistorized version of this switch described by Seader[4] and shown in Fig. 3. The mode of operation is self-evident and it is necessary only to remark that the replacement of the triodes $S_1 \ldots S_n$ by the transistors enables the matching transformers to have far lower impedance levels or, even for the transformers to be eliminated and suitable windings (10 to 20 turns) placed directly below the heads.

The advent of the symmetrical junction transistor leads inevitably to its applications as a head switch and a typical circuit is that of Ridler and Reynolds[5] shown in Fig. 4. the obvious disadvantage of this circuit is that, when a particular head is selected by applying a suitable negative bias to the base of the appropriate transistor, the base current divides itself between the head and the line matching transformer, M. While it is true that the base current may be only 1/10 to 1/50 as great as the writing current, disturbance of the written matter may still be produced by

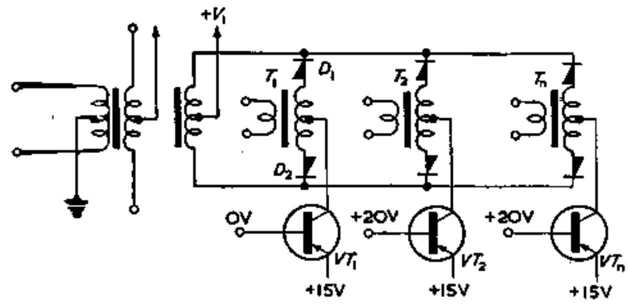

Fig. 3. Seader's head switch

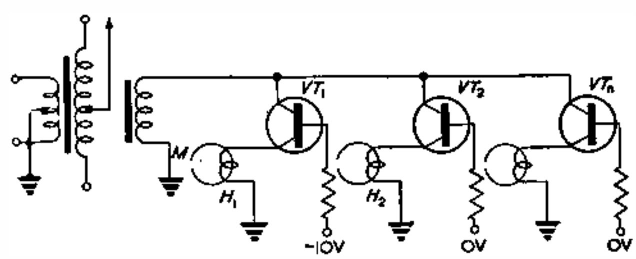

Fig. 4. Ridler and Reynold's switch

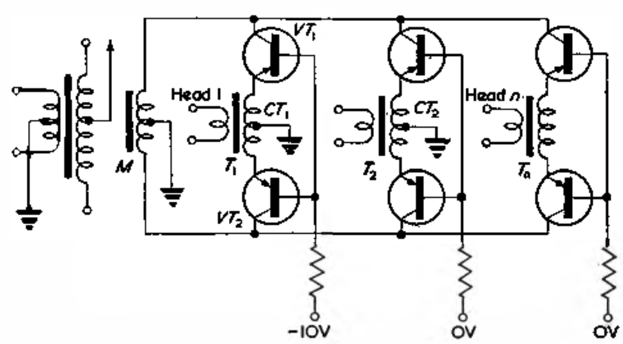

Fig. 5. Symmetrical circuit

the act of selection, and also considerable alteration can occur in the magnetic state of the transformer M. Some experiments conducted with a Ferroxcube cored transformer showed a transient whose duration could not be reduced below 200μsec. It is only fair to say that Ridler and Reynolds have proposed a slightly different version of this circuit in which the effect of the transient is minimized.

An obvious development from the circuits shown in Fig. 3 and Fig. 4 is given in Fig. 5. Here a pair of transistors, either symmetrical or normal, are used to replace the diodes of Fig. 3. The appropriate head is selected by suitably biasing the bases of the selected transistor pair while earthing the remainder. It is clear that no unbalanced d.c. flows in the circuit, so that, within the limits of component sym-

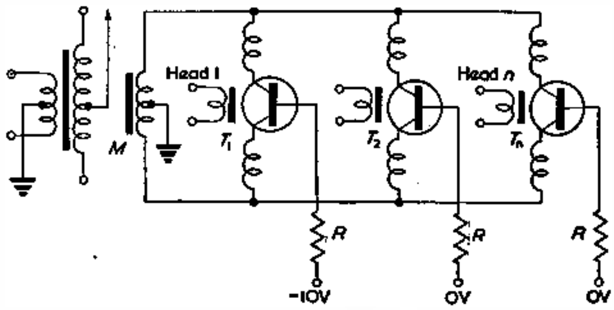

Fig. 6. Symmetrical, one transistor, circuit

metry, polarization of transformer cores does not occur. The circuit has been drawn with normal (asymmetrical) transistors for which the transformer $T_1 \ldots T_n$ (or the heads themselves if the high impedance version is used) require a centre tap $CT_1 \ldots CT_n$. If symmetrical transistors are used, however, this centre tap is unnecessary. This circuit operates in a satisfactory manner and generates a transient whose duration is only 12 to 20μsec; its disadvantage lies in the fact that two transistors are needed for each track selected. This is not a severe disadvantage since for switching times of the order mentioned quite inexpensive transistors are satisfactory. This is because the high speed impulses in the circuit derive from the transformer M, the transistors merely acting as open- or short-circuits.

The ultimate simplicity seems to be the circuit shown in Fig. 6. Here a single symmetrical transistor is used to make or break the head (or matching transformer) circuit at its centre point. In the system used on the Birkbeck College M.2 computer a Wharf Engineering Laboratories Mark 4 magnetic drum is used. This has low impedance heads so that matching transformers $T_1 \ldots T_n$ are needed. It proved possible, however, to use the small Mullard FX1011 pot cores and a compact switch unit was produced incorporating an Ediswan XS101 transistor and the transformer mounted on a B7G base.

The switching transient with these units is due almost entirely to the reading amplifier, and has a duration of less than 12μsec, but perhaps the most important advantage of the switch is its low noise level. Noise due to high back-biasing in the diode circuits has already been mentioned; in the transistor circuit, however, the unselected transistors have effectively zero voltage applied to them during reading and this makes them practically free from random noise. In the circuit used the signal breakthrough from unselected heads is such that 100 unselected heads produce about 10 per cent of the signal from a selected head. This effect can be reduced by using a matrix selection system of the type proposed by Seader[4], where groups of heads feed auxiliary transistor switches before reaching the matching transformer or amplifier.

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Valve Cooling for Reliability

By M. Virgon*

One of the most important features revealed by investigations into the reliability of electronic equipments is the nature of valve failures. The high intrinsic reliability of the thermionic valve is severely degraded when run at high bulb temperatures and this is responsible for seventy-five per cent of equipment failures. Two of the detrimental effects which arise from overheating are reversal of the gettering process and electrolysis of the glass envelope.

Improved versions of valves on the Preferred List are now available. Their extremely high potential reliability will be fully exploited and reflected in equipment reliability only when the engineering construction permits the valves to be operated at greatly reduced bulb temperatures. A 6BW6 (CV2136) dissipating 16W in conventional engineering attains a bulb temperature of 170°C above ambient. If this rise is reduced at least by half reliability is considerably improved. The heat withdrawn from the valves must not unduly increase the temperature of other components or the net gain in equipment reliability will be reduced.

For guided weapons test equipment where reliability is of paramount importance, Bristol Aircraft Limited has developed a chassis engineering system which eliminates the thermal deficiencies associated with conventional design. The system includes a specially designed valve mounting clip which carries the heat from the valve bulb to the chassis. The amount of heat removed by conduction can easily be made greater than that removed by convection and radiation.

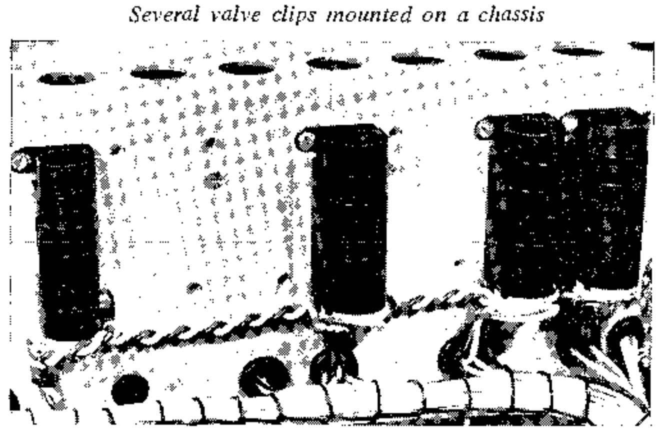
The Bristol clip is made from one piece of 26 s.w.g. brass blanked out and formed into six fingers which embrace the valve over the maximum area. There are effectively six separate clips overcoming variations of envelope diameter along its length. Each finger has its own thermal circuit to the base of the clip and thence to the chassis so that heat is not conducted via the clip from the hotter to cooler glass. Generally the hottest part of the bulb is halfway along its length, opposite the centre of the anode, and it would be detrimental to the valve to conduct this heat to the extremities since this would encourage reversal of the gettering process at the pip end and electrolysis and embrittlement of the glass at the other end.

The base of the clip is a trough section which lends rigidity thus enabling good thermal contact to be maintained between clip and chassis with only two point fixing. The large contact area between clip and chassis gives a good thermal path to cooler metal. The clip is made in two sizes similar in all respects except the diameters which are varied to suit B9A and B7G valves. Both sizes are suitable for pin and flying lead valves.

Since the metal clip totally envelopes the valve it serves as a screen and an earthing point is provided. Positive radial and longitudinal valve location is maintained even under severe vibration. A test at constant acceleration of 10g over the frequency range of 20 to 1 500c/s had negative results. The length of the clip affords complete mechanical protection for the longest miniature valve. Flying leads are protected against vibration fatigue and mechanical damage.

Tests show that two-thirds of the heat is removed by conduction through the Bristol clip which has a dissipation factor of less than 3°C rise above ambient per watt dissipated.

Several valve clips mounted on a chassis



*Bristol Aircraft Ltd.

AN EIGHT DIGIT WORD GENERATOR

using Surface Barrier Transistors and Direct Coupled Technique

(Part 2)

By P. L. Owen*, B.A., and T. R. H. Sizer*, A.M.I.E.E.

(Voir page 191 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 199)

The Invertor

Fig. 20 shows two circuits P and Q connected by means of a transistor in the grounded emitter configuration. The most convenient way of explaining the logical operation of this device is to regard the transistor as a variable impedance between the collector and emitter terminals, the magnitude of this impedance being determined by the amount of base current drawn.

Representing the input impedance of Q by Z the equivalent circuits for the two states are shown in Figs. 21(a) and (b). When the base-emitter voltage is large negative ($-0{\cdot}5$V) representing a 1 input the emitter base current is high and the collector-emitter impedance is low compared with Z and R. The collector voltage is then just slightly negative

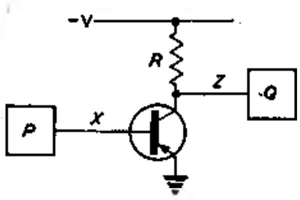


Fig. 20. The invertor

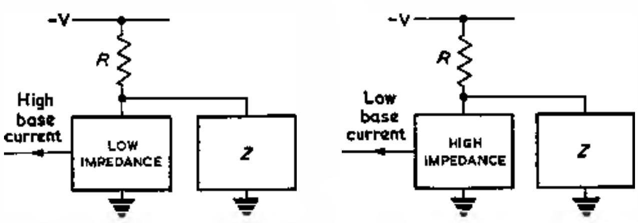


Fig. 21. Impedance states of the invertor

of earth ($-0{\cdot}03$V), its actual value being determined by the loading resistor R in series with the low collector-emitter impedance of the transistor. This voltage, by convention, represents a 0. When the emitter-base voltage is slightly negative of earth ($-0{\cdot}03$V), representing a 0 input, the emitter-base current drawn is small and the collector-emitter impedance is high compared with Z and R. The collector voltage is determined by the loading resistor R in series with Z. By suitable choice of R this voltage will be $-0{\cdot}5$V and the output signal thus represents a 1.

The relation between the input and output is summarized in Table 5 and is seen to be identical with Table 1 for the logical operation of inversion.

And and Or Gates

Fig. 22(a) shows two transistors VT_1 and VT_2 and a resistor in series connected between earth and the negative power supply. The collector of the top transistor VT_1 is connected to the base of a third transistor VT_3 which is also connected in series with a resistor R_2 between earth and the negative supply. Let the states of the inputs to the bases of VT_1 and VT_2 be represented by the logical symbols X and Y respectively, the state of the connexion between the collector of VT_1 and the base of VT_3 be M, and the state of the collector of VT_3 be Z.

The inputs to the device are X and Y and four combinations are possible. The equivalent circuits for these combinations are shown in Fig. 23.

In the case of the 0,0 input, the base currents of VT_1 and VT_2 are low and the transistors form two high impedances in series. The collector voltage of VT_1 is defined by the load resistor R in series with the base-emitter impedance of VT_3. R is chosen so that sufficient base current is drawn to make VT_3 low impedance and by using the same argument as in the invertor, the output of VT_3 is a 0.

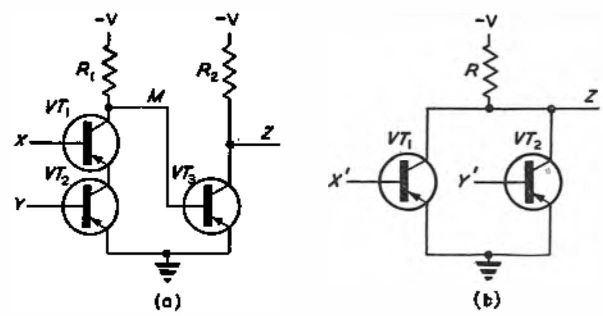


Fig. 22. 'And' gate configurations

In the case of a 1 input to VT_2 and a 0 input to VT_1 (Fig. 22(a)) these transistors form a high impedance in series with a low impedance and the base current to VT_3 is defined as in the 0,0 case giving a 0 output.

The case of the 0 input to VT_2 and a 1 input to VT_1 (Fig. 22(a)) while appearing to be similar to the previous case, in fact, requires a little closer examination. Since the input to the lower transistor is 0, only a low base current

TABLE 5

INPUT X	BASE CURRENT	COLLECTOR-EMITTER IMPEDANCE	OUTPUT Y
0	Low	High	1
1	High	Low	0

TABLE 6

X	TRANSISTOR VT_1	Y	TRANSISTOR VT_2	M	TRANSISTOR VT_3	Z
0	High impedance	0	High impedance	1	Low impedance	0
1	High or low impedance	0	High impedance	1	low impedance	0
0	High impedance	1	Low impedance	1	Low impedance	0
1	Low impedance	1	Low impedance	0	High impedance	1

* Royal Aircraft Establishment, Farnborough.

is drawn and therefore it is in the high impedance state. The voltage input to the upper transistor corresponds to a 1, but very little base current can flow from the emitter since the resistance to earth contains the high impedance of VT_2. Some current may flow, however, via the base-collector diode of VT_1 and the base of VT_3, the magnitude of which depends on whatever input device is connected to the base of VT_1. Thus VT_1 may be in the high or low impedance state. This does not affect the state of VT_3, however, since the path from the collector of VT_1 to the emitter of VT_2 is always high impedance. Thus VT_3 is in the low impedance state and the output is a 0.

In the final case (Fig. 22(a)) both VT_1 and VT_2 are in the low impedance state and the voltage of the collector of VT_1 is defined by the load R in series with the low impedance path to earth via VT_1 and VT_2. The voltage is thus slightly below earth and a low current is drawn from the base of VT_3. VT_3 is then in the high impedance state and the output is a 1.

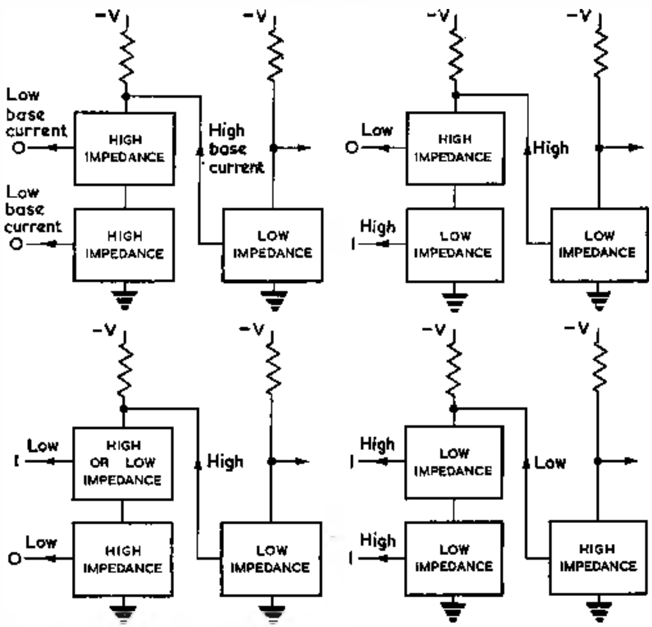


Fig. 23. 'And' gate impedance states

The behaviour of the circuit is summarized most conveniently in Table 6.

The relation between the output Z and the variables X and Y shows that the configuration functions as an AND gate.

If the inverse signals X' and Y' are available, it is possible to synthesize the AND function with the simpler configuration of Fig. 22(b). In this network only two transistors in parallel and a loading resistor are necessary. The state of the transistors and the outputs for the four input combinations are given by Table 7.

For the first three combinations at least one of the transistors is in the low impedance state and, since the transistors are in parallel, the common collector point must be at a potential of about −0·03V corresponding to a value 0 for the output variable Z. For the fourth combination neither transistor is in the low impedance state and the common collector point takes a potential defined by the loading resistor in series with the input impedance of the following circuit. R is chosen so that this potential is about −0·5V and the value of the output variable Z is 1. Thus, provided the inverse signals X' and Y' are available the network shown in Fig. 22(b) performs the logical AND function. Since only two transistors are used compared to three in Fig. 22(a), the introduction of this type of gate where possible results in a more economical use of transistors.

TABLE 7

X	X'	TRANSISTOR VT_1	Y	Y'	TRANSISTOR VT_2	Z
0	1	Low impedance	0	1	Low impedance	0
1	0	High impedance	0	1	Low impedance	0
0	1	Low impedance	1	0	High impedance	0
1	0	High impedance	1	0	High impedance	1

Corresponding OR gate configurations are shown in Figs. 24(a) and (b) the corresponding state or 'truth tables' are given in Tables 8 and 9.

Multiple input gates may be formed by putting more than two transistors in series or parallel. The factors which determine the limiting number of inputs are the small voltage drop across the conducting transistors in the case of series configurations and the leakage current in the case of parallel configurations. In the case of the 2N-240 the leakage current is very small so, if possible, multi-input

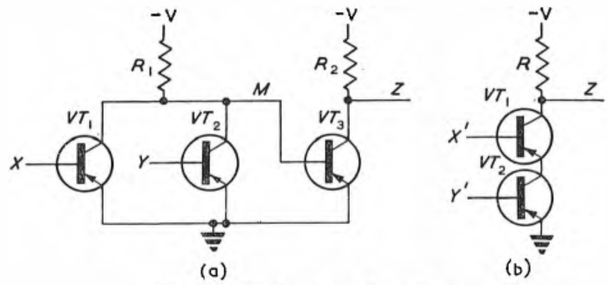


Fig. 24. 'Or' gate configurations

gates should be constructed in terms of parallel configurations rather than series configurations. For example the word generator has been designed so that in no case is there more than three transistors in series, but as many as eight transistors have been used in parallel.

The Simple Binary Counter

It has been shown that if the bases and the collectors of two transistors are cross connected as shown in Fig. 19 a simple bistable element or flip-flop is obtained.

This device may be made to change state by the application of suitable current pulses to the appropriate collector points. For instance, if VT_1 is non-conducting, the only

TABLE 8

X	TRANSISTOR VT_1	Y	TRANSISTOR VT_2	M	TRANSISTOR VT_3	Z
0	High impedance	0	High impedance	1	Low impedance	0
1	Low impedance	0	High impedance	0	High impedance	1
0	High impedance	1	Low impedance	0	High impedance	1
1	Low impedance	1	Low impedance	0	High impedance	1

TABLE 9

X	X'	TRANSISTOR VT_1	Y	Y'	TRANSISTOR VT_2	Z
0	1	Low impedance	0	1	Low impedance	0
1	0	High impedance	0	1	Low impedance	1
0	1	High or low impedance	1	0	High impedance	1
1	0	High impedance	1	0	High impedance	1

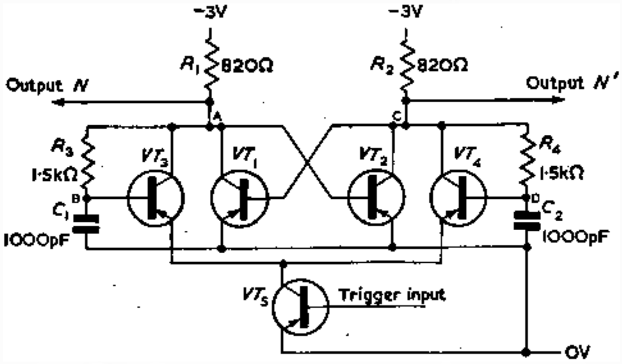


Fig. 25. Direct coupled simple binary counter

current in R_1 is that which comes from the base of VT_2. If now sufficient current is injected into the collector point of VT_1, the additional current in R_1 causes the potential of this collector point and also the base of VT_2 to rise to earth. VT_2 then ceases to conduct and VT_1 starts to conduct, the device thus taking up the other stable state. Subsequent removal of the external current signal has no effect and the device has changed state. A similar argument applies to a signal at the other collector point.

The simple binary counter described earlier is a one input device which changes state on the arrival of a pulse. Such a device may be constructed from the flip-flop shown in Fig. 19 if additional circuits are provided to gate the incoming pulses to the appropriate collector points. A suitable circuit is shown in Fig. 25.

The gating elements consist of three further transistors VT_3, VT_4, VT_5 and two resistor-capacitor delays. Suppose that the flip-flop is in the state where VT_1 is non-conducting and no current is being drawn from the base of VT_3. In the equilibrium state the collector of VT_1 and the base of VT_3 are both at the same negative potential and the collector of VT_2 and the base of VT_4 are both at a potential slightly negative of earth. The arrival of a negative going edge at the base of VT_5 causes VT_5 to conduct. VT_3 also conducts since its base is negative and current flows through VT_3 and R_1 to raise the potential of the collector of VT_1 to earth, thus causing the flip-flop formed by VT_1 and VT_2 to change state. The collector of VT_1 is now slightly negative of earth potential, and the base of VT_3 rises to this potential with a delay determined by the resistance capacitance circuit associated with its base, the current through VT_3 is thus cut off. Meanwhile, the base of VT_4 starts to go negative with a delay determined by the resistance-capacitance circuit. This base, however, does not reach the potential of the collector of VT_2 since a residual charge remains on C_2 until the negative potential is removed from the base of VT_5. Thus by suitable choice of the values of the delay circuits VT_4 does not conduct sufficiently to change the state of the flip-flop back again until the negative pulse has been removed and the next negative pulse has arrived at the base of VT_5.

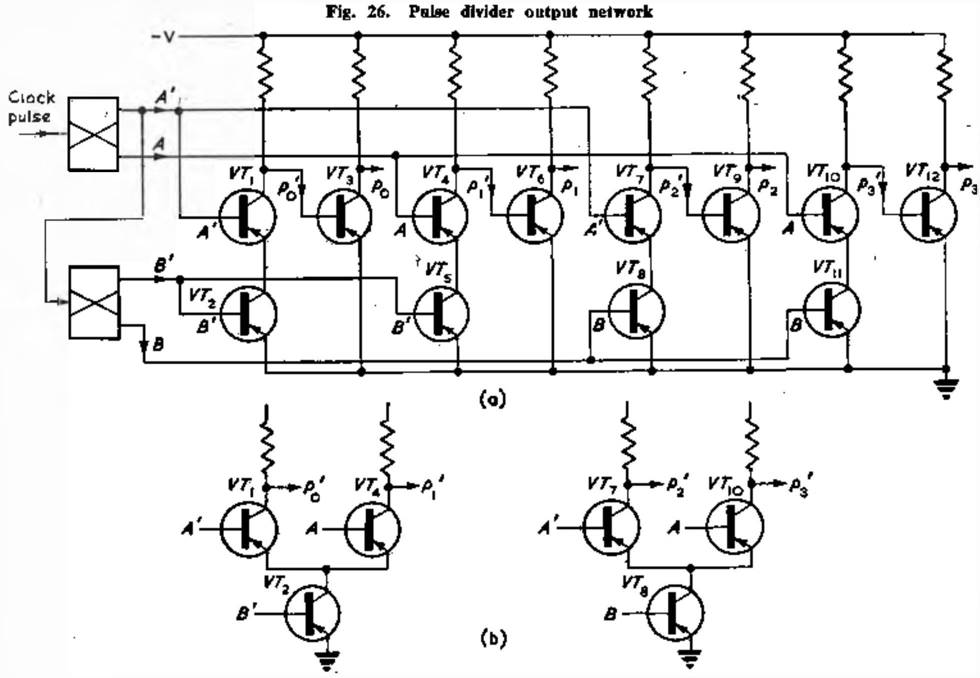
The circuit shown in Fig. 25 thus changes state once only on the arrival of each negative going edge at the base of VT_5 and, by virtue of the convention that a negative potential represents the binary number 1, satisfies the requirements of a binary counter as described previously.

The Development of Tree Circuits

As described previously the outputs of the pulse divider are obtained by taking the outputs from the binary counter and combining them in AND gates. The straightforward way of doing this is to have a separate AND gate for each combination of outputs. Fig. 26(a) illustrates the case for the four output pulse divider. The AND gates are of the type shown in Fig. 22(a) and for each combination of outputs from the binary counters a particular collector point is in the 0 state and all the rest are in the 1 state. For example, when the counter is in the 0,0 state $A=0$ and $B=0$, and VT_1 and VT_2 are both in the low impedance state giving a 0 output from the collector of VT_1. For all the other gates at least one transistor is in the high impedance state and the outputs are 1. If the outputs from the collector points are passed through invertors, one output is 1 and all the rest are 0. The arrival of a clock pulse changes the state of the counter and another output is 1. Thus the outputs from the invertors are those required from a four state pulse divider.

It is possible, however, to simplify the network in the following way. Examination of Fig. 26(a) shows that transistors VT_2 and VT_5 receive the same signal i.e. B' and that both have grounded emitters. These transistors thus have identical behaviour during the operation of the pulse divider and may therefore be replaced by a single transistor. A similar replacement may be made for VT_8 and VT_{11}. The resulting network is shown in Fig. 26(b) and is known as a tree. Examination of this network shows that for any combination of states of the binary counter one, and only one, collector point of transistors VT_1, VT_4, VT_7 VT_{10} is connected to earth via a low impedance path. This collector

Fig. 26. Pulse divider output network



is then in the 0 state while the others are in the 1 state. Inverting these outputs as before gives the required signals for the pulse divider.

By building up the AND gates into a tree two transistors have been saved. The extension of this to more than three inputs achieves even greater economies, and the case of three inputs is described in detail later.

Detailed Description of the Word Generator

Since the word generator is required to produce inputs to two input devices such as adders as well as to one-input devices such as shift registers and complementers, provision was made for the generation of two independent words by duplication of the toggle switches and the circuits to be described. Also the type of circuit developed required the strobe pulses for control purposes and these were brought out as additional outputs. A synchronizing pulse for an oscilloscope was also provided.

The Circuit Elements

The following description takes the same form as the elementary word generator already described but for convenience the strobe generator is described first. A block diagram of the complete equipment is given in Fig. 27.

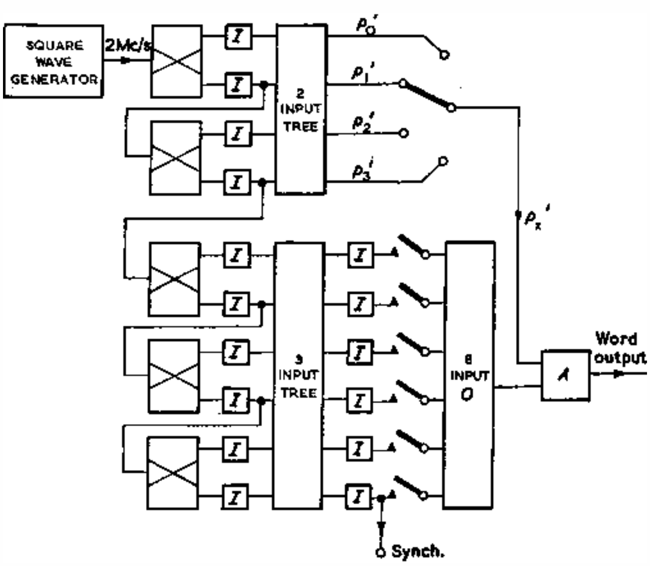
The Strobe Generator

The strobe generator is a two digit binary counter triggered from a 2Mc/s square wave generator. The outputs of the binary counter are connected via invertors to a four output tree. The outputs of the tree are then inverted to form the four strobe signals. The invertors between the counter and the tree are not necessary for logical reasons, but act more as buffers for the counters. It was found that for stable operation of the flip-flops it was essential that the external loading be minimized and this was achieved by isolating the flip-flops by means of the invertors. The collector resistor of each invertor was then adjusted to meet the individual loading conditions within the limit of the maximum collector current permitted, a point dealt with in the section on limiting conditions.

The Pulse Divider

This comprises a three digit binary counter connected to a three input eight output tree via buffer invertors. The trigger input is obtained from one of the flip-flops in the

Fig. 27. Block diagram of word generator



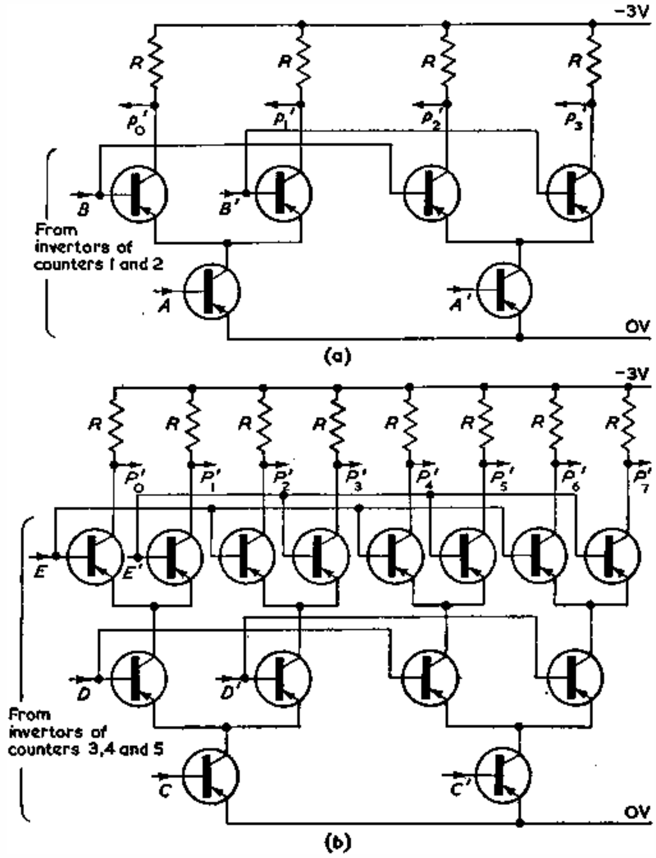


Fig. 28. Direct coupled tree circuits

(a) Direct coupled 2 high tree
(b) Direct coupled 3 high tree

strobe generator. The outputs of the tree are then inverted to produce the required eight outputs. Circuits of these trees are given in Figs. 28(a) and (b).

The Switches and Or Gates

The outputs of the pulse divider were passed through two sets of toggle switches to two eight input OR gates of the parallel transistor type, the collector points of which give the inverses of the words set up on the toggle switches.

Since the inverse signals are available both from the eight input OR gate and from the strobe pulse generator the final gating or strobing is done by an AND gate of the type shown in Fig. 22(b) and the word output appears in the correct sense at the collector point of this final AND gate.

Operating Details

Potentiometers are used as variable resistors for the loads on the OR gates. It was originally anticipated that adjustment for ageing would have to be made but, in fact, no adjustment has been necessary in the first fifteen hundred hours of operation.

Two wafer switches are provided, one for selecting the strobe pulse to gate the desired quadrant of the digit period and one to select waveforms for checking purposes. A synchronizing pulse is taken from digit period seven.

There are no other controls in the equipment. An actual waveform of the word output is reproduced in Fig. 29.

Electrical Details

Each digit is of 0·5μsec in width and lasts one quarter of the digit period. Eight digit periods form one word length which lasts 16μsec, i.e., a word repetition frequency of 62·5kc/s.

At a nominal supply voltage of $-3{\cdot}0$V the current consumption is approximately 100mA. The equipment operates satisfactorily over an input voltage range of $-2{\cdot}8$ to $-3{\cdot}4$V.

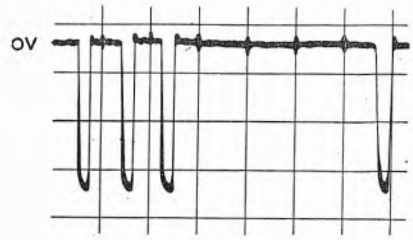


Fig. 29. Waveform of word output
Word 11100001
Digit width: 0·5m/sec
Amplitude −3·0V nominal (off load)
Word repetition frequency 62·5kc/s

The components used are:

Transistors, Type 2N240	83,
Resistors—two values only	50,
Capacitors—one value only	10,
Potentiometers	4.

Limiting Conditions

The ambient temperature of the equipment has been limited to a maximum of 30°C, the average being about 24°C, and no failures* have ocurred in about fifteen hundred hours running time. However, assuming an acceptable failure rate of, say, better than $\frac{1}{4}$ per cent/1 000 hours, it can be estimated from a study of the lifetime curves of the 2N-240 that this failure rate would be exceeded if the ambient temperature rose above 35°C to 40°C depending on the dissipation in the transistor. In applications where a transistor is used as a switch, the dissipation is generally small and it is reasonable to assume that 35°C represents a limiting figure for the word generator. Above this figure the leakage current I_{cbo} and I_{ceo} may have a significant effect on the failure rate.

It is of interest to examine the number of bases which may be connected in parallel to one collector point. Suppose one has five transistors as shown in Fig. 30 each with a collector load of 820Ω with their bases commoned to the

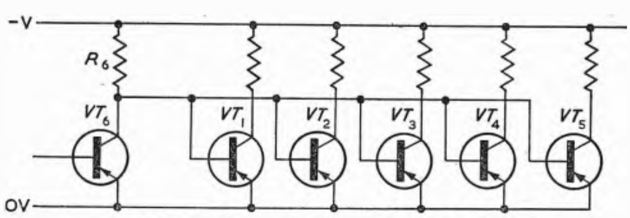


Fig. 30. Circuit for determining maximum loading

collector of a sixth transistor with a collector load of $R_6\Omega$. Assuming identical transistor characteristics then the voltage across each resistor R_1 to R_5 is:

$$(V - V_{ce})$$

where V is the supply voltage and V_{ce} is the saturation voltage drop when a transistor is conducting in the saturation region. The value of V_{ce} is taken in this application to be $-0{\cdot}03$V. The collector current I_c for each transistor VT_1 to VT_5 is given by:

$$I_c = \frac{V - V_{ce}}{R} = \frac{3 - 0{\cdot}3}{820} = 3{\cdot}62\text{mA}$$

It can be seen from Fig. 16(b) that the minimum value of base current to maintain this collector current is 1mA and Fig. 17(a) shows that the base-emitter potential at 1mA is $-0{\cdot}3$V. One has therefore moved from the collector of, say, VT_1 and determined the base-emitter potential. Assuming identical transistors, then if five time-bases are commoned to the collector of VT_6, when VT_6 is non-conducting the value of R_6 is the controlling factor in ensuring that the five commoned transistors can enter the saturation region. Hence R_6 is approximately

$$\frac{3{\cdot}0 - 0{\cdot}3}{5 \times 1\,000 \times 10^{-6}} = 540\Omega.$$

With this value of collector load, VT_6 when conducting only passes 5·5mA, well within the maximum specified limit of 15mA. Rubinoff, Beter *et al*[2] have dealt with this point in more detail considering the case where the transistors are not identical. In the equipment described in this article no collector drives more than four bases and in most cases only two.

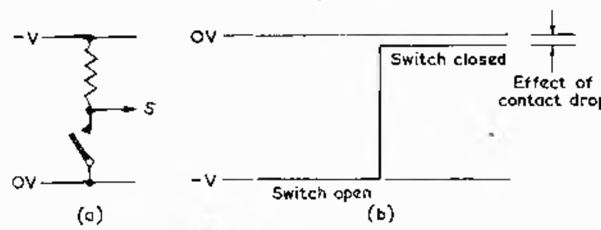


Fig. 31. Simple switch

It is often required to place several transistors in series to form the AND gates described earlier but ultimately a stage would be reached where the collector of the top transistor would never be brought close enough to earth to switch a connected transistor into the 'off' state. The limit to this number has been shown to be five[2] but in the word generator this number is set to a maximum of three in series.

Conclusions

The equipment described in this article has been functioning under laboratory conditions for fifteen hundred hours without the occurence of any fault concerning the transistors. It is concluded that the advantage of direct coupled circuits in permitting circuit simplicity is considerable and that the 2N-240 is suitable for such applications at ambient temperatures between 25°C and 35°C.

Acknowledgments

Acknowledgment is due to Messrs. A. Stratton, A. M. Pegrum and C. S. Fromberg of the Ministry of Aviation for their constructive criticism of the article in the course of preparation.

Appendix

Transistors as Switches

Logical operations can be performed by switches as shown in Figs. 31 and 32. The use of relay switches in this manner is well known and an analysis of the principles was carried out by Shannon[3] in 1938. Brown and Beter[1] have applied the same principles to switching operations with transistors.

Considering the switch in Fig. 31(a) when the contacts

* *It should be noted that a failure is not necessarily a catastrophic fault; the movement of any characteristic outside the limits demanded by the direct coupled technique is a failure even though it may be of a transitory nature.*

are open an open-circuit is presented to current through R and the potential at S is $-V$. When the contacts are closed a small resistance, the value depending on the state of the contacts, is placed in series with R and the potential at S is very nearly zero. If a second switch is put in series with the first, Fig. 32(a), then the output potential will only be near zero when both switch N and switch M are

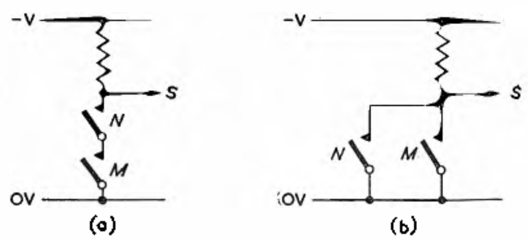


Fig. 32. Switches as logical devices

(a) S indicates a change only when both N and M are closed or when either N or M is opened after both are closed

(b) S indicates a change in potential when either N or M is closed after both are open or when N and M are opened after both are closed

closed. In Fig. 32(b) S is near zero when either switch N or switch M is closed.

Consider Fig. 33(a) where a pnp device is connected in a common emitter configuration. There is a direct analogy to the switch of Fig. 31(a) with the collector representing one pole, the emitter a second pole and the base region acting as a toggle connecting one to the other. The emitter, base and collector currents are related by:

$$I_{\text{emitter}} = I_{\text{base}} + I_{\text{collector}}$$

Referring to Fig. 33(a), in the 'off' condition, where S is open-circuited and no significant current flows through

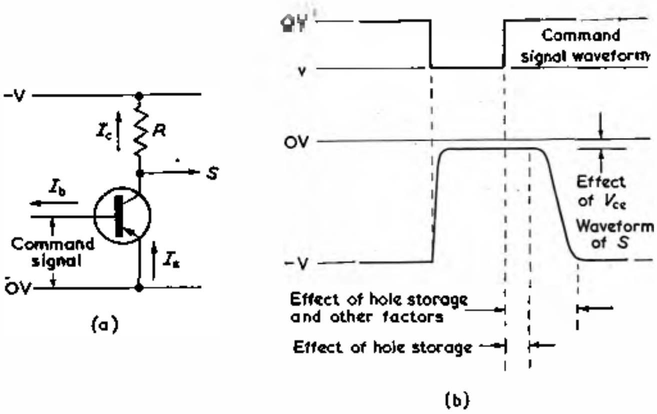


Fig. 33. A pnp transistor in a common emitter configuration

the load resistor, the base is open-circuited and the potential at S is substantially $-V$. To enter the 'on' condition, a negative potential is applied between base and emitter; current flows through R and a voltage drop V_{ce} analogous to the contact drop appears across the transistor. The value of V_{ce} varies with different types of transistor, but for any one type it varies with the collector current and base potential. For use in the switching applications discussed in the article V_{ce} needs to be limited to a small maximum figure and its variation known accurately with changes in collector and base currents.

In applications of the type under consideration the time taken for the transistor to start conducting is extremely short and is not generally governed to a great extent by external conditions. The time taken to switch out of the conducting state, however, is a function both of the physical design of the transistor and the electrical parameters of the circuit. This dependence upon circuit parameters is of direct relevance to the section of this article dealing with direct coupled logic and it is necessary to examine briefly the mechanism of current flow in a pnp transistor. More precise information can be obtained from Bradley[4].

For simplicity a transistor may be considered as two diodes back to back, the emitter-base diode and the base-collector diode. In the circuit of Fig. 33(a), the emitter-base diode is forward biased to a condition of low resistance and the base collector diode reverse biased to a state of high resistance when the base is negative relative to the emitter.

When the base is positive relative to the emitter the transistor will be non-conducting, except for leakage currents which are assumed to be insignificant because both diodes will be in a high resistance state. As the base potential goes negative relative to the emitter, the base-emitter resistance is lowered and the diode proceeds to pass current. This takes the form of holes being injected by the emitter into the base region, a small fraction of which recombine with electrons at the base lead termination. The remainder diffuse across the base region and recombine at the collector forming the collector current.

As the base potential goes more negative the emitter injects more holes into the base region and both base and collector currents increase. Eventually, a point is reached where all the supply potential has been dropped across the load resistor and the base region contains more holes than are necessary to maintain the collector current at its maximum value. The base region is acting as a store of holes.

If now the base is returned to zero potential, the emitter-base diode, again neglecting the leakage current, ceases to conduct but the collector current is maintained near the maximum level until the excess holes have been collected. Thus the total decay time of the negative excursion of the collector signal, shown in Fig. 33(b) is governed not only by the external circuit but by the degree of saturation this, in itself, being governed by the physical construction of the device.

For fast switching, the saturation characteristic must be controlled by limiting the current in the base or emitter with the use of diodes, bias potentials and resistors. However, for the present application, adequate switching speeds can be obtained by the use of a transistor which has a very thin base region. It follows that with a device of this kind no precaution to control the parameters affecting saturation, need be taken.

The hole storage effect in the 2N-240 is minimized by using an electrochemical process in manufacture to form a thin uniform base region a few microns in width. To switch 'on' requires about 0·10μsec, to switch off from saturation about 0·12μsec, one or two orders of magnitude faster than most alloy junction devices operating under the same conditions of saturation.

Correction

In part 1 of this article Tables 2 and 3 were transposed.

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2. Beter, R. H., Bradley, W. E., Brown, R. B., Rubinoff, M. Surface Barrier Transistor Switching Circuits. Semiconductors Ltd. Appl. Note SB-240.
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4. Bradley, W. E. Principles of the Surface Barrier Transistor. *Proc. Instn. Radio Engrs.* 41, 1702 (1953).

Effects of Argon Content on the Characteristics of Glow-Discharge Tubes

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

Measurements have been made to determine the effects of varying the argon content of glow-discharge stabilizer tubes on the striking and running voltages, the running-voltage/temperature curves, the initial drifts and the impedance/frequency and noise characteristics. Work was originally carried out on neon-filled and helium-filled tubes, having cerium cathodes with argon contents varying from zero to 3·5 per cent in 0·5 per cent steps. Later studies were confined to neon-filled tubes with argon contents varying over the relatively wide range of 0·001 to 10 per cent having molybdenum anodes and cathodes and employing high-stability reference-tube manufacturing techniques. The results of the work are presented here and discussed.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

IN recent years the demand for glow-discharge voltage-regulator tubes of high stability has risen considerably. It is well known that such tubes form simple and inexpensive methods of voltage stabilization and are extensively used in electronic stabilizers of the degenerative type for providing a reference voltage[1]. They have also been employed as reference elements in stabilizers employing grid-controlled rectifiers[1,2]. Glow-discharge tubes are frequently used too with linear resistors, or in combination with other non-linear devices, in a bridge circuit for detecting a change in the input or output voltage of an automatic voltage stabilizer and producing a signal to operate a device for correcting the output voltage to a pre-determined value[1,3].

The desirable characteristics of glow-discharge tubes vary with the application; e.g. when used as a voltage stabilizer the tube should present a low impedance to ripple frequencies and have good regulation over a large current range. The performance of a simple stabilizer circuit can easily be calculated assuming ideal characteristics but unless the tube is itself stable and free from drift and erratic voltage jumps the calculation is of little value. With a voltage reference tube, whether it be used in a degenerative-type stabilizer, in a voltage-sensitive bridge, or elsewhere, the temperature coefficient of running voltage and the initial voltage drift should be small and, in addition, the running-voltage/current characteristic should be stable and reproducible to within small limits if the tube is used intermittently. This characteristic should also be free from voltage steps and hysteresis.

For automatic-telephone switching application[4] it is also important to have stable tube characteristics and, in particular, the usefulness of the tube may be severely limited for such purposes unless the impedance presented to a.c. signals over the audio-frequency range and the noise produced are low. Stability and a reasonably flat running-voltage/current curve are again essential if the tube is used as a coupling element between stages of a d.c. amplifier[5,6].

During the last decade a great deal of work has been carried out on glow-discharge tubes. Much of this consisted of comparing the performances of commercial stabilizer and reference tubes[1] and the results indicated that many tubes exhibited characteristics which severely limited their use for the applications just mentioned. The following characteristics have been studied in detail; (1) running-voltage/current curves (particular attention being paid to regulation, voltage steps and hysteresis), (2) initial and long-term drifts, (3) change of running voltage with temperature, (4) impedance-frequency curves[7], (5) noise[8]. Although this work proved useful in assessing some of the limitations of tubes it did not indicate how the characteristics could be improved because the physical parameters of the tubes varied considerably. Later, therefore, tests were carried out on a number of special tubes with different cathode materials, gas fillings, gas pressures and construction to determine the effect of these parameters on the various characteristics[9,10]. Tubes with cathodes of copper, nickel, tantalum, molybdenum, zirconium and cerium, and with neon/argon and helium/argon gas-fillings were used for these investigations. Gas pressures were varied in steps from 30 to 50mm of mercury. The results obtained from these special tubes suggested that another tube parameter, namely the argon content in the gas filling (which had not been held constant in the tubes examined) might have a considerable effect on the characteristics. A small quantity of argon is added to the main filling to reduce both the striking and running voltage of the tube and also to minimize the difference between these two voltages.

Subsequent examinations[11] of similar special tubes in which only the argon content was varied, confirmed that this parameter was important. The tubes made for these investigations had cerium cathodes and main gas fillings of (a) neon (b) helium. Argon content varied from zero to 3·5 per cent in steps of 0·5 per cent. The gas pressures used were 40mm of mercury in the helium-filled tubes and 50mm of mercury in the neon-filled ones. Each tube had a rod anode 1mm in diameter surrounded by a cylindrical cathode 20mm in diameter and 30mm long. Mica sheets at each end of the cathode cylinder enclosed the working portion of the tube. A glass envelope, having a volume of about 45cm³, housed the electrode structure. The tube samples were made under exactly similar conditions and from identical materials. The conclusions which could be drawn from this work were limited, however, by the large spread of results observed for tubes of a given type and by the fact that no impedance/frequency and noise characteristics could be obtained for neon tubes having more than 2·5 per cent of argon because of the presence of internal high-frequency oscillations. The running-voltage/current curves exhibited voltage steps which increased in magnitude with the argon content. The large spread of results, the oscillations and the voltage steps are now thought to be related to non-uniformities of the cerium-coated cathode surfaces which were employed.

It was then decided to make a fuller study of the effect of varying the argon-content in neon-argon filled tubes by changing the argon content in steps over the relatively wide range of 0·001 to 10 per cent[12]. To obtain reliable

* *The University of Sheffield.*

characteristics, tubes with molybdenum electrodes were chosen and high-stability reference-tube manufacturing techniques were employed. The following percentage argon contents were used: 0, 0·001, 0·01, 0·1, 0·25, 0·5, 0·75, 1·0, 3·0, and 10·0. The important dimensions and other features of the tubes are shown in Fig. 1. Running-voltage/current curves, running-voltage/temperature curves, initial drifts, impedance-frequency curves, noise characteristics and running-voltage variations with life were determined for these tubes. The variation of striking voltage with argon content was not recorded because the results varied considerably for tubes of the same type due largely to variations in the pre-breakdown gap length shown on Fig. 1. Striking voltages tend to be higher, however, with low argon contents.

Measurements

Running voltages at various currents and changes in running voltage with time and temperature were determined by comparing a known fraction of the tube voltage with that of a standard cell using a vernier potentiometer[13,14]. The ambient temperature was varied by immersing the tubes in an electrically-heated oil-bath.

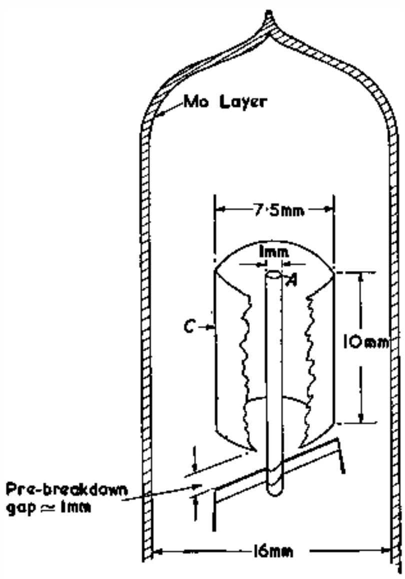


Fig. 1. High-stability tube construction

The impedance-frequency characteristics of the tubes with cerium cathodes were obtained from alternating-current and voltage measurements together with phase-angle determination using an oscillograph technique[11]. The effective a.c. resistances and inductances of the molybdenum-cathode tubes were found in the frequency range 300c/s to 30kc/s using a modified form of Owen bridge network[12]. This method was a good deal more accurate than the previous one, particularly at low frequencies.

All noise measurements were made with a calibrated amplifier-detector unit having a noise bandwidth of 87kc/s.

During the life tests each tube current was chosen so that the glow just covered the inside surface of the cylindrical cathode and the running voltage was measured intermittently at a current corresponding to the minimum running voltage.

Results and Discussion

(a) Cerium Cathodes

Fig. 2 shows mean curves of variations in striking and running voltages with changes in argon content for cerium-cathode tubes. Very small additions of argon are seen to have a great effect on the striking and running voltages. A minimum difference between these voltages may be obtained when the argon content is between 1 and 1·5 per cent for helium-filled tubes and less than 0·5 per cent for neon-filled ones. These results agree fairly well with those obtained by Jurriaanse, Penning and Moubis[15] on molybdenum/neon-argon tubes and by Meissner[16] on tubes with nickel cathodes.

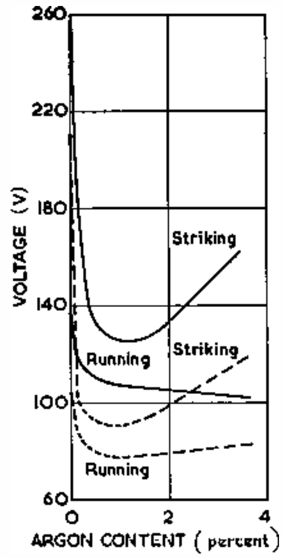


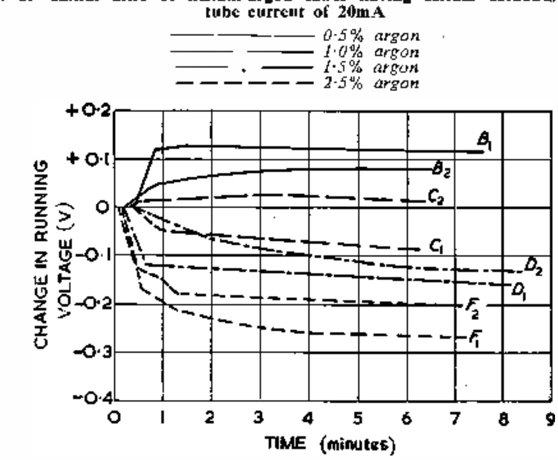
Fig. 2. Striking and running voltages as a function of argon content for cerium-cathode tubes

(Running voltages recorded for a tube current of 20mA)
——— helium-filled tubes
— — — neon-filled tubes

The running voltages of both pure-helium and pure-neon filled tubes fell as the tube current was increased, the regulation being large. Helium-filled tubes also showed considerable hysteresis in the voltage-current curves. Addition of argon had the effect of decreasing the regulation although voltage steps in the characteristics gave some tubes poor regulation figures. Except for tubes containing no argon, the voltage steps increased with increasing amounts of argon and were much larger in helium tubes than neon ones having the same argon content. Helium-argon tubes also showed larger hysteresis effects than neon-argon ones.

Tubes with no argon exhibited a negative temperature coefficient of running voltage very much greater than those which had. With argon present the voltage changes with temperature were very small and the amount of argon used is not very important in this respect. Helium-filled tubes with 1 per cent argon, however, gave the optimum results. All helium-filled tubes, irrespective of the argon content, displayed both positive and negative slopes in the running-voltage/temperature curves. For all neon-filled tubes, however, the temperature coefficient was found to be negative in the range 20 to 40°C and usually up to 80°C and it increased slightly with increasing amounts of argon. Neon-filled tubes with argon contents up to 1·5 per cent generally had almost zero temperature coefficients over the range 40 to 70°C.

Fig. 3. Initial drift of helium-argon tubes having cerium cathodes for a tube current of 20mA



The curves of Fig. 3 indicate the effect of argon content on the initial drifts in running voltage of helium tubes. The initial drifts are small and in all cases drifting was almost complete about 3min after switching on.

Fig. 4 shows the effect of argon content on the impedance of a helium-filled tube. For frequencies above about 10kc/s an increase in argon content reduces the tube impedance and this confirms some earlier work by Benson and Bental[9].

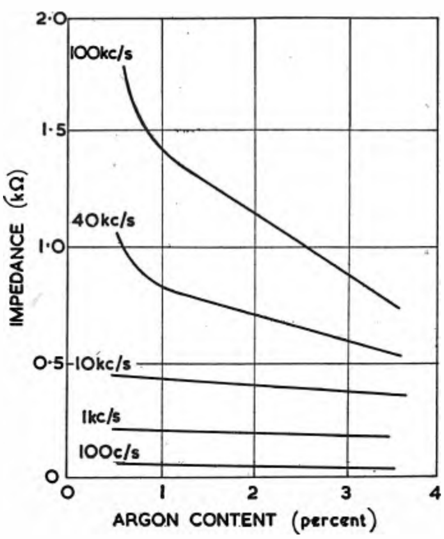


Fig. 4. Variation of impedance with frequency and argon content of helium filled cerium-cathode tubes

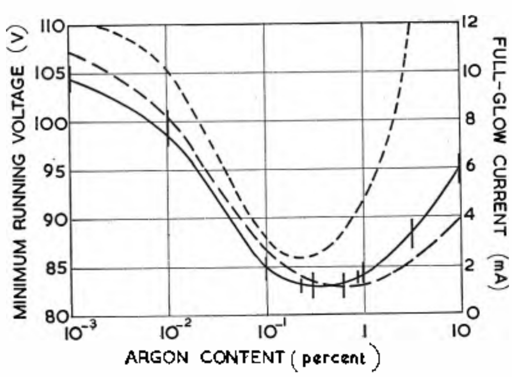


Fig. 5. Minimum running voltage and full-glow current as a function of argon content for neon filled molybdenum cathode tubes

——— *minimum running voltage*
— — — *minimum running voltage according to Jurriaanse, Penning and Moubis (reproduced from reference 15 by permission of Philips Electrical Ltd)*
– – – – *full glow current*

(b) MOLYBDENUM CATHODES

The minimum running voltage is plotted against argon content in Fig. 5 and the variation of full-glow current, which is a measure of the current density, is also shown. The results are seen to agree fairly well with those of Jurriaanse, Penning and Moubis[15] who used a different electrode geometry which could account for the shift between the two running-voltage curves. A minimum running voltage occurs at about 0·3 per cent argon and a minimum spread (as indicated by the short vertical lines on Fig. 5) was observed in the characteristics for tubes having this argon content. This may be due to the difficulty of measuring and controlling the amount of added argon in the case of very low contents or, for argon contents greater than 1 per cent to the inability to age and sputter the tubes adequately.

Typical running-voltage/current characteristics are shown in Fig. 6. Such curves were reproducible from day to day to within 0·05V for tubes having up to 1 per cent argon and to within 0·1V for tubes with higher argon contents.

For tubes with low argon contents, the minimum running voltage occurs at higher currents than for tubes containing 1 per cent argon, and the subnormal-glow region appears to be greatest when 0·01 per cent argon is present although

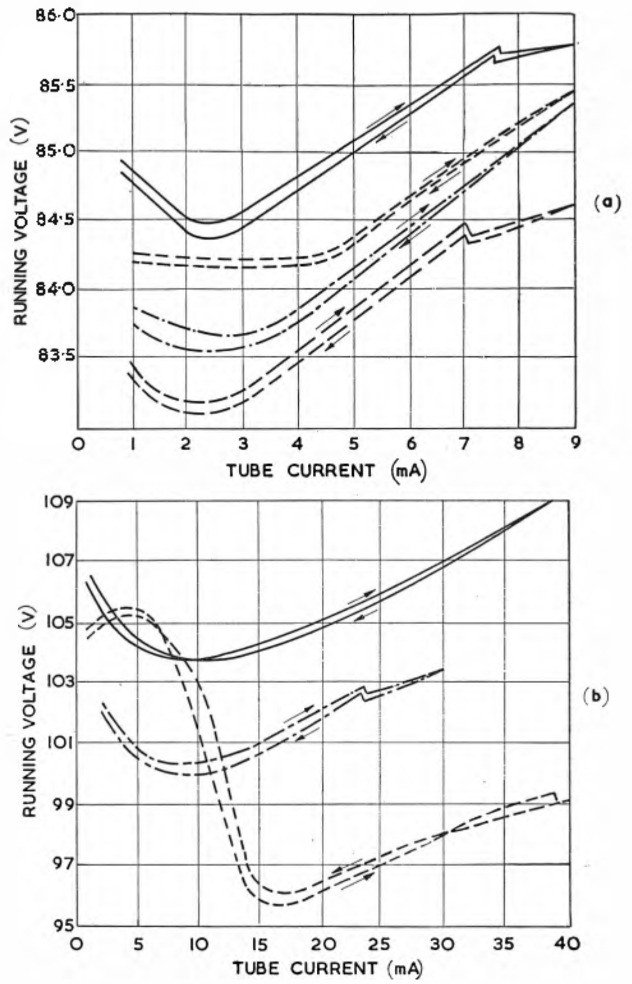


Fig. 6. Running-voltage/current characteristics of neon filled molybdenum-cathode tubes

(a) *0·1% argon*
0·25% argon
0·5% argon
1·0% argon
(b) *10^{-3}% argon*
10^{-2}% argon
10% argon

this is not too clear from the typical curve of Fig. 6(b) This is not attributed to impurities because the characteristics are reproducible and the glow expands uniformly over the cathode surface with increasing current. It was observed that with the 0·01 per cent argon tubes, at currents below 4mA, the glow appeared as a diffused region filling most of the anode-cathode space, and at higher currents when the discharge was almost abnormal, the glow formed a defined negative glow region near the cathode. These phenomena seem to be associated with the ionization process in the gas.

With increasing argon content above 0·25 per cent, the subnormal glow recedes and the normal glow becomes more easily definable, and with 1 per cent argon the running-voltage/current curves are sensibly flat over the current range 1 to 5mA.

The area of cathode covered by glow in tubes with more than 1 per cent argon contracted at lower currents after ageing. This is probably responsible for the observed form of running-voltage/current curves at low currents. The minimum running voltage occurs at much higher values of the tube current, where the discharge has a cleaning influence which offsets the spoiling effect due to impurities.

It is difficult to draw conclusions from the tests on tubes with 3 and 10 per cent argon, because at currents below 15mA the contraction phenomenon was prevalent, although

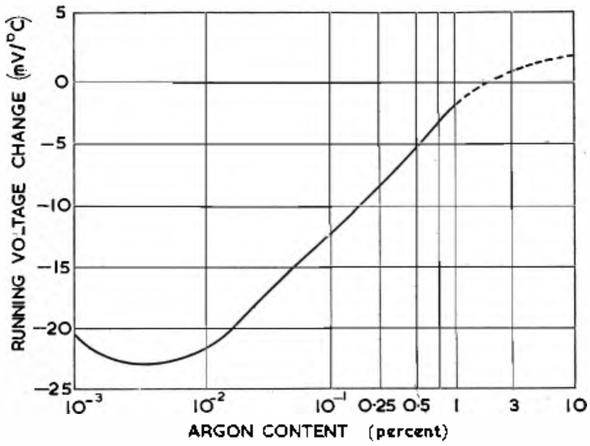


Fig. 7. Average voltage temperature coefficient as a function of argon content for neon-filled molybdenum cathode tubes

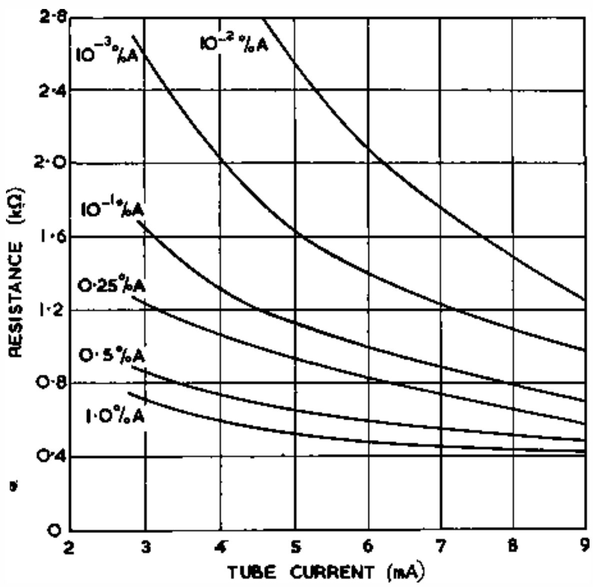


Fig. 8. Variation of resistance at 10kc/s with tube current and argon content

the hysteresis which usually accompanies this was small. No voltage steps or jumps were observed in the running-voltage/current curves in the normal operating range. This may seem surprising in view of steps having been observed in the characteristics of neon-argon tubes with cerium cathodes. These steps although affected by the argon content, are, however, now thought to be due to an unevenly-prepared cathode surface, for which the sputtering and preparation procedures were influenced by the amount of argon to differing extents.

The small steps in running voltage which are observed at high currents occur when the glow jumps to the outer surface of the cathode cylinder. Previous to this the running voltage rises with increasing current owing to the edge effect at the cathode.

The results of the running-voltage/temperature measurements are summarized in Fig. 7 which shows a plot of

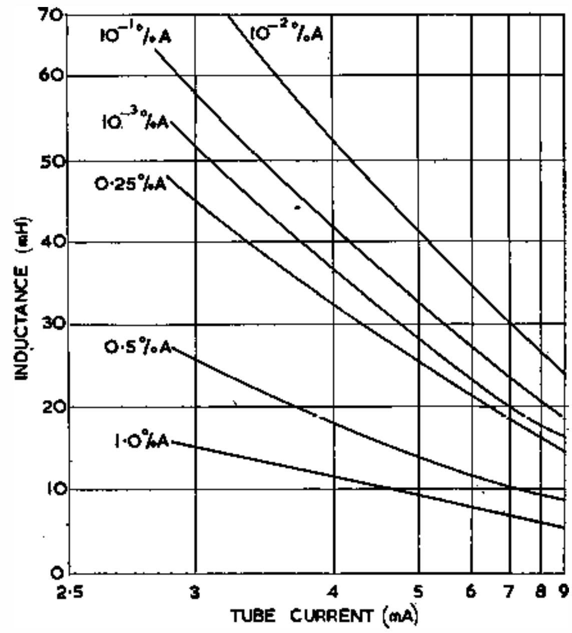


Fig. 9. Variation of inductance at 10kc/s with tube current and argon content

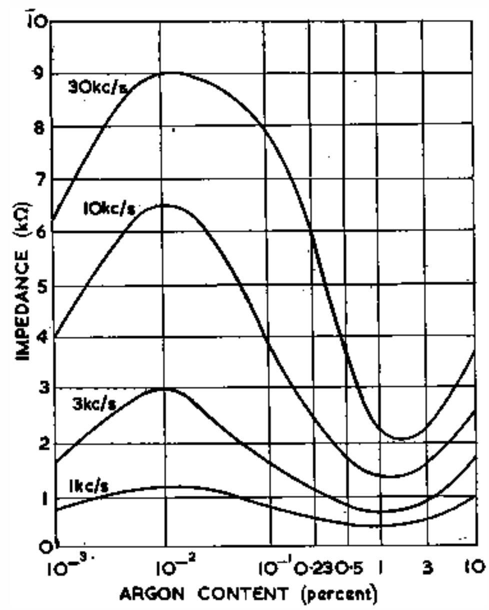


Fig. 10. Impedance as a function of argon content for various frequencies

average voltage temperature coefficient against argon content for the range 20 to 50°C.

Most of the change in voltage between the anode and the cathode in a glow-discharge tube occurs over a short distance from the cathode (the cathode-fall region) and is dependent on the gas density in this region. If the tube envelope temperature increases, heating of the gas near the envelope will cause it to expand and will increase the gas density, near the cathode.

The temperature coefficient of running voltage of a tube has been calculated in the past[1,17] by considering a density change in the cathode region and relating this to the slope of the running-voltage/pressure curve at the initial tube pressure. A modified form of the earlier theory has been developed by the authors where changes in gas pressure, and not density, are estimated for varying tube-envelope temperature. The temperature coefficient of the tube is shown to be proportional to the slope of the running-voltage/pressure curve. Temperature coefficients calculated from the results of the modified theory, for pure-neon and neon +0·5 per cent argon tubes agree closely with the observed values, over the temperature range 30 to 50°C. Further work is being undertaken at present to determine the forms of the equations for the running-voltage/pressure curves of tubes for various gas mixtures so that changes in running voltage can still be estimated even for large temperature variations.

The initial drift of running voltage was found to be negative for most tubes and, for small argon contents, to decrease with increasing amounts of argon. The drift was very small for argon contents between 0·1 and 1 per cent and in most cases was complete after 3min.

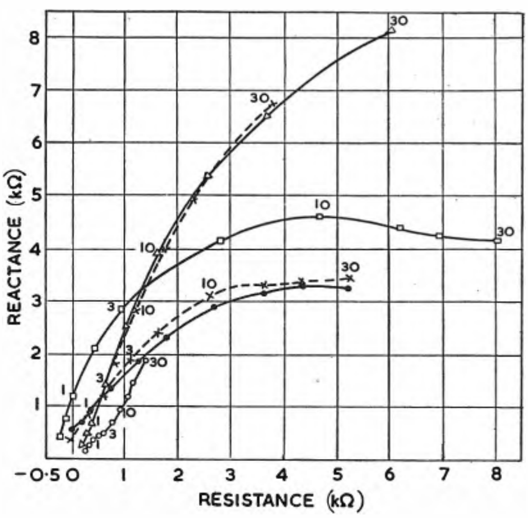


Fig. 12. Equivalent circuit of a glow-discharge tube

pure neon
10^{-3}% argon
10^{-2}% argon
0·1% argon
0·3% argon
1·0% argon

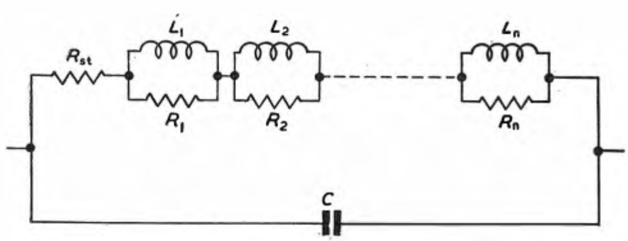


Fig. 12. Equivalent circuit of a glow-discharge tube

It is thought that there is a connexion between initial drift and running-voltage/temperature coefficient as the results tend to suggest. The initial drift can be ascribed to a gradual increase in gas pressure as the gas in the cathode region reaches an equilibrium temperature determined by the heating effect in the cathode region; this increase in gas pressure would result in the lowering of the running voltage as explained previously.

It is difficult to compare the impedance-frequency characteristics in the normal-glow region at a fixed tube current since no one current can be chosen where the discharge is normal in all tubes. Figs. 8 and 9, however, show the general variation of effective resistance and inductance with tube current and argon content at a frequency of 10kc/s. The effects of argon are greatest at low tube currents and high frequencies (Fig. 10). Most of the impedance characteristics are perhaps better illustrated when plotted as vector loci (Fig. 11). The nature of the impedance locus is being analysed by the authors at present and will be the subject of a forthcoming paper. Briefly, the equivalent circuit of the glow-discharge tube (Fig. 12) can be derived theoretically if the delayed effects present in the discharge are taken into account. The time-constants of the parallel inductive and resistive elements of the equivalent circuit are associated with the time factors of each delayed effect.

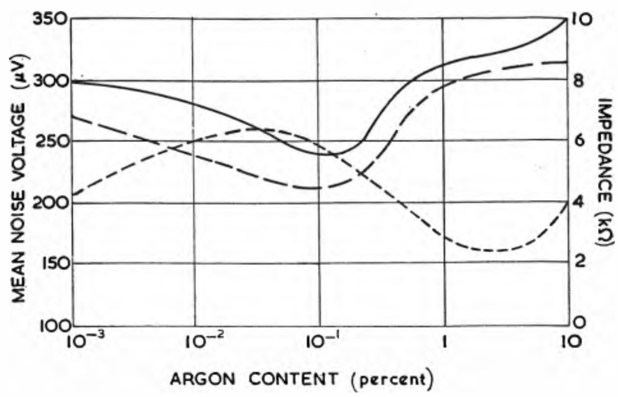


Fig. 13. Mean-noise voltage as a function of argon content for a tube current of 3mA

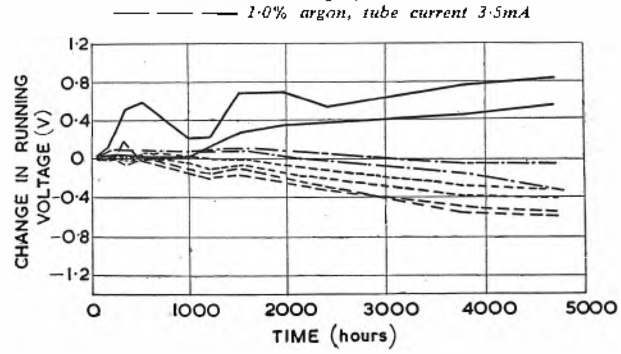
Noise voltage for a load resistance of 37·5kΩ
Equivalent noise voltage e_n
Equivalent noise impedance z_n

In neon-argon gas-filled tubes one such delayed effect is the electron production by neon metastable atoms. The magnitude and time factor of this process will be greatly influenced by the argon concentration.

The noise generated in all glow-discharge tubes is found to decrease with increasing tube current. A comparison of the noise-voltage/argon content curve of Fig. 13 with the current-density/argon content curve of Fig. 5 shows that minimum noise occurs with about 0·1 per cent argon corresponding to the region of minimum current density. Benson and Gillespie[11] have previously stated that the noise is probably due to collisions of electrons with gas atoms and the probability of excitation and ionization by collision is a function of electron energy. Thus the noise generated will also be a function of electron energy and the minimum noise (Fig. 13) coincides with the minimum electron energy which according to Kruithof and Penning[18] occurs when the

Fig. 14. Change of running voltage with life

·001% argon, tube current 12mA
0·1% argon, tube current 2·3mA
0·5% argon, tube current 3·0mA
1·0% argon, tube current 3·5mA



argon content is about 0·1 per cent. A tube may be regarded as a noise-voltage generator of e.m.f. e_n in series with an equivalent noise impedance z_n. Fig. 13 shows that the variation of equivalent noise impedance with argon content has the same form as that of the impedance at high frequencies when plotted against argon content (Fig.10).

The results of the life tests are shown in Fig. 14. Despite the random variations in running voltage, the general trend of the curves suggests that the influence of the argon content is negligible for contents greater than 0·01 per cent. The voltage change over the first 1 000 hours is small and, subsequently, there is a steady decrease of running voltage, except for the lowest argon content which shows an increase. The negative drift of running voltage is difficult to account for precisely but is probably due to the predominance of one of a number of processes taking place in the tube during life. The main physical changes which are possible are:— clean up of neon or argon, release of argon initially trapped in the cathode or sputtered layer during the preparation of the tube (this would lead to a decrease of running voltage in the case of the argon contents less than 1·0 per cent), release and clean up of impurities, in particular oxygen which could account for random fluctuations[19] and a change in cathode surface roughness. The fluctuations observed here may well have been due to current and temperature variations during the running period of the tube. In order to minimize these effects it would be better to stabilize the tube current and record the running-voltage variations continuously with a potentiometric recorder[19].

Conclusions

As stated earlier, the conclusions which could be drawn from the work on tubes with cerium cathodes were limited by the large spread of results obtained for tubes of a given type and by the presence, in many tubes, of high-frequency oscillations. The non-uniformities of the coated-cathode surfaces probably were the cause of the unexpected results.

The most desirable characteristics of neon tubes with molybdenum cathodes are obtained when the argon content is about 1·0 per cent. Minimum impedance and zero-voltage/temperature coefficient, however, can be realized for argon contents lying between 1 and 3 per cent, but it is difficult to prepare tubes which have similar and reproducible characteristics when the argon exceeds 1 per cent.

Some recent work has been carried out substituting first krypton and then xenon for the argon in similar molybdenum-neon reference tubes. Krypton and xenon influence all the characteristics in a similar way to argon.

Acknowledgement

The work recorded has been carried out in the Department of Electrical Engineering at the University of Sheffield. The authors wish to thank Professor A. L. Cullen for facilities afforded in the laboratories of this Department, and for his constant encouragement and advice; also Mr. G. P. Burdett, for carrying out many careful examinations of running-voltage/current and running-voltage/temperature characteristics and initial drifts of tubes. They also wish to acknowledge the kindness of the English Electric Valve Co. Ltd, in supplying all the special tubes for examination, for financial assistance with the work, for the loan of equipment and for many helpful discussions and valuable suggestions.

In preparing this article the authors have made use of some of the information and diagrams in the papers listed as items 11 and 12 under the section References. These two papers were first published by The Institution of Electrical Engineers as Paper No. 2394R and Monograph No. 321R respectively and have been republished in the Proceedings of the Institution.

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An Automatic Grading System for a Granite Quarry

An electronic control system for screening, storage and automatic grading has recently been installed at the Cliffe Hill Granite Company's quarry at Markfield, Leicester by Ericsson Telephones Ltd, Nottingham.

About 1 400 tons of stone are quarried daily and processed on this site for use as coated stone, ready mixed concrete, road chippings, aggregates and railway ballast.

After blasting at the quarry face, the stone is reduced in size by a primary crusher to 6in and then taken by conveyor to a secondary crusher and a screener installation. This screening plant classifies the stone into ⅜, ¾, 1½ and 2in grades and each grade is then fed into separate bins to 100 tons capacity from which they can be directed by conveyor if required to tertiary crushers producing ¼ and ⅛in grades, or direct to a screen contained in the final storage installation.

This latter plant contains eleven bins, seven of 300 tons and four of 150 tons capacity and all being fitted with pneumatically operated discharge gates.

The automatic grading system as now installed enables the plant to carry out the following:

(1) Deliver batches of material to an existing tar-macadam plant with an hourly capacity of 30 to 40 2-ton batches.
(2) Deliver batched material direct into lorries.
(3) Deliver batched or graded material to a ready-mixed concrete plant.
(4) Deliver graded materials to stock.

The control system console is fitted with plug-in programming units so that eleven materials contained in the hoppers can be metered in any desired proportion by constant rate feed systems which have a maximum timing rate of 99 seconds and can be set to an accuracy of one second.

A formula number corresponding to a batch is displayed by Digitron number indicating tubes on the control panel and a batch counter is fitted which will deliver up to ten batches of any desired formula.

THE MEASUREMENT OF TRANSISTOR NOISE FIGURE

By F. J. Hyde*, M.Sc., A.M.I.E.E.

The standard noise diode technique is considered. General formulae are presented which take into account (1) the transmission properties of the coupling circuit between the noise diode generator and the transistor and (2) the effect of subsequent amplifier noise. These formulae contain readily-measurable admittances and voltage ratios.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

At frequencies up to a few hundred megacycles per second a standard noise diode is generally used in the measurement of noise figure. The diode is connected in parallel with the source admittance of the device under test, which is followed by an amplifier and output power indicator. With the diode unenergized a noise output reading is taken. Commonly, the noise diode anode current I_d is then adjusted so that the noise output power is doubled. Provided that certain simplifying assumptions are made then a simple expression for noise figure ensues[1]. This is:

$$F = (20 I_d / G_s) \,.\, 290/T \quad \ldots\ldots\ldots\ldots \quad (1)$$

where G_s is the source conductance and T is the absolute temperature of the source. The assumptions made in deriving this formula are as follows† :—

(1) The frequency of measurement is such that the noise current generator $\langle i_d^2 \rangle$ associated with the noise diode is given by the temperature-limited shot noise formula $\langle i_d^2 \rangle = 2eI_d\Delta f$, where e is the electronic charge and Δf is the bandwidth. This presupposes temperature-limited operation of the diode and that its noise power output is uniform over the frequency range considered. This must be such that flicker noise and transit time effects, at low and high frequencies respectively, are insignificant.

(2) The source conductance G_s may be indentified with the conductance shunting the diode noise current generator terminals.

(3) The contribution to the noise output from the amplifier, which follows the device under test, is negligible under all conditions.

Although the above assumptions are often valid for a limited range of operation, there will arise conditions for which, in particular, (2) and (3) are no longer true. The purpose of this note is to draw attention to modifications to formula (1) which then arise. The modified formulae involve readily-measurable quantities. Although these formulae are of general usefulness, emphasis is placed on the determination of the noise figure of transistors. A suitable schematic representation of a transistor, which includes its noise properties, is presented in the next section. The analysis is restricted to the frequency range within which diode lead inductance can be ignored.

Equivalent Noise Representation of a Four-Pole Device

It has been shown[2,3] that any noisy four-pole device (e.g. a transistor) may be represented by the schematic diagram shown in Fig. 1. This consists of a noiseless device with the input of which a series noise voltage generator e_{nt} and a shunt noise current generator i_{nt} are associated. In general these two generators are not independent. The noiseless device has the same small-signal properties as the practical device. For subsequent use Y_{it} and Y_{ot} are defined as the input and output admittances of the transistor; these depend on the practical terminations used.

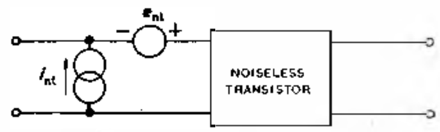

Fig. 1. Schematic representation of a transistor

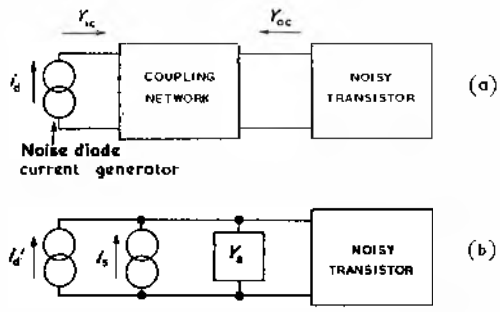

Fig. 2. Schematic representation of noise-diode-to-transistor coupling
(a) Direct (b) Equivalent; $Y_s \equiv Y_{oc}$

The Effect of the Coupling Circuit Between Noise Diode and Transistor

Fig. 2(a) shows schematically the noise diode coupled to the transistor. Here i_d is the shot noise generator in parallel with the anode-cathode terminals of the diode. The coupling network comprises the components associated with the anode circuit of the diode (including anode-cathode terminal capacitance) and the input circuit of the transistor. It also includes stray couplings between these components and to earth. The internal conductance of the diode due to the Schottky effect is assumed to be small compared with the conductance of the external circuit. A_c is defined as the open-circuit voltage gain of the network and its input and output admittances are denoted as Y_{ic} and $Y_{oc} \equiv Y_s$ respectively. The output admittance of the network is clearly the source admittance for the transistor. It readily follows that as a source for the transistor Fig. 2(a) may be redrawn as in Fig. 2(b), where $i_d' = i_d Y_s A_c / Y_{ic}$.

Consider now the experimental determination of noise figure on the basis of Fig. 3, in which the amplifier is assumed to be noiseless. Here i_d' is defined as above and i_s is the thermal noise current generator associated with $Y_s(\langle i_s^2 \rangle = 4kTG_s\Delta f)$, where G_s is the conductive part of Y_s and k is Boltzmann's constant. It is assumed that the standard technique is used and that the noise diode anode current is adjusted so that the mean square noise current in a link aa' is doubled. As is shown in appendix 1 it

* *University College of North Wales, Bangor.*

† *Quantities enclosed in angular brackets are averaged values.*

follows that the transistor noise figure F_t is given by:

$$F_t = 20\,I_d\,.\,(290/T)\,.\,(|A_c|^2/G_s)\,|Y_s/Y_{ic}|^2 \;\ldots\ldots\; (2)$$

Alternatively, if A_c' is defined as the voltage gain of the coupling network when its output is loaded by the input

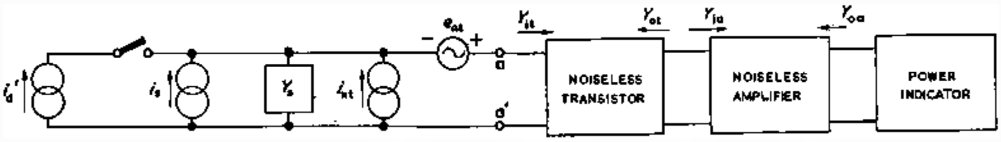

Fig. 3. Schematic representation of apparatus for noise figure measurement: noiseless amplifier

admittance Y_{it} of the transistor, i.e. $A_c' = A_c Y_s/(Y_s + Y_{it})$, then the expression for F_t is:

$$F_t = 20\,I_d\,.\,(290/T)\,.\,(|A_c'|^2/G_s)\left|\frac{Y_s + Y_{it}}{Y_{ic}}\right|^2 \ldots\ldots (3)$$

In equations (2) and (3) the quantities involved may readily be measured. The admittances and conductance are obtained by bridge techniques while A_c or A_c' can be determined using a high input impedance millivoltmeter and attenuator. It is important that the millivoltmeter should not affect the circuit conditions; i.e. its admittance should be small compared with that across which it is connected. A suitable basic type of indicator has been described by Eaglesfield[4]. An absolute calibration of this indicator is not necessary if it is merely used as a constant level indicator.

The Effect of Amplifier Noise

In the previous section it was assumed that the amplifier was noiseless. The more general case when this is not so will be considered now. In Fig. 4 the amplifier is represented as a noiseless fourpole having input and output admittances Y_{ia} and Y_{oa} respectively, with associated noise generators i_{na} and e_{na} at the input.

A_t and A_a are defined as the ratios of the output to input voltages of the transistor and amplifier respectively under loaded conditions. Let F_o be the overall noise figure of the transistor plus amplifier and F_a the noise figure of the amplifier alone when connected to a source admittance Y_{ot} at temperature T. Then it is shown in Appendix 2 that:

$$F_t = F_o - \frac{(F_a - 1)}{A_t^2}\,\frac{|Y_s + Y_{it}|^2}{|Y_{ot} + Y_{ia}|^2}\cdot G_{ot}/G_s \ldots . (4)$$

The measurement of F_o is straightforward. To measure F_a in a separate experiment the amplifier must be fed from a source of admittance Y_{ot}, the conductance of which must be at the same temperature as G_s, namely T. In determining both F_o and F_a the relevance of equations (2) and (3) should be borne in mind. The admittances of equation (4) are once more readily obtained by bridge methods and A_t by using a high impedance millivoltmeter and attenuator.

Discussion

The effects of the transmission properties of the coupling circuit between the diode and transistor and of the noise of the subsequent amplifier may be taken into account by means of equations (2) or (3) and (4) respectively. The quantities which are involved in these equations may be measured in a straightforward manner. In this connexion it may be pointed out that an equivalent form of equation (4) due to Friis[5] is:

$$F_t = F_o - (F_a - 1)/G_{av} \;\ldots\ldots\ldots\ldots\; (5)$$

where G_{av} is defined as the 'available gain' of the transistor. By using equation (4) the necessity for making gain measurements under conjugate-matched conditions, which are required for a direct determination of G_{av}, is avoided.

A quick test to determine whether the amplifier noise is making a significant contribution to the output is to de-energize the transistor and note the subsequent background noise reading. If this is very much lower than the operational reading then amplifier noise is probably insignificant. This is not more than a qualitative test, however, since the de-energized transistor will have an output admittance which differs from the operational output admittance. Furthermore, since the de-energized transistor is in thermal equilibrium with its surroundings, it will be the seat of Johnson noise which is directly related to its output admittance by Nyquist's formula.

Conclusion

In experimenal work at radio frequencies which involves complex source admittances the only satisfactory way of obtaining accurate results is by measuring the source admittance at each setting using a high-grade bridge. As is indicated by equations (2) and (3) it is also desirable to measure the transmission properties of the circuit coupling the diode to the device under test. It should not be assumed that amplifier noise is negligible simply because de-energizing the transistor produces a significant decrease in total noise output. An experimental check based on equation (4) is worthwhile.

Appendix

(1) Noise Figure of Transistor Followed by Noiseless Amplifier

Consider the current i_{aa}' flowing in a shorting link aa' in Fig. 3. With the noise diode unenergized:

$$i_{aa}' = i_{nt} + i_s + e_{nt}Y_s \;\ldots\ldots\ldots\ldots\; (6)$$

With the noise diode energized:

$$i_{aa}' = i_{nt} + i_d' + i_s + e_{nt}Y_s \;\ldots\ldots\ldots\; (7)$$

and as a consequence of the mean square short-circuit current being doubled, in this case:

$$\langle i_d'^2\rangle = \langle(i_{nt} + i_s + e_{nt}Y_s)(i_{nt} + i_s + e_{nt}Y_s)^*\rangle \;\ldots .\; (8)$$

where the asterisk denotes the complex conjugate of the quantity in the brackets. Using as the definition of noise figure:

$$F_t = \frac{\text{Total mean square noise current in link } aa'}{\text{Mean square noise current in link } aa' \text{ due to the source}}$$

it follows that:

$$F_t = \frac{\langle(i_{nt} + i_s + e_{nt}Y_s)(i_{nt} + i_s + e_{nt}Y_s)^*\rangle}{\langle i_s i_s^*\rangle} = \langle i_d'^2\rangle/\langle i_s^2\rangle \;\ldots\ldots\ldots\ldots\; (9)$$

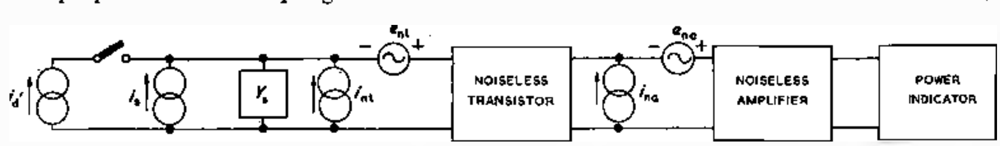

Fig. 4. Schematic representation of apparatus for noise figure measurement: noisy amplifier

or:

$$F_t = 20\,I_d\,(290/T)\,(|A_c|^2/G_s)\,|Y_s/Y_{ic}|^2 \;\ldots\ldots\; (10)$$

(2) The Effect of Amplifier Noise

Fig. (4) refers. By Norton's theorem the source for the noiseless transistor can be represented by the admittance Y_s in parallel with a noise current generator $i_s + i_{nt} + e_{nt}Y_s$. Hence the transistor output terminal voltage is given by $(i_s + i_{nt} + e_{nt}Y_s)\,A_t/(Y_s + Y_{it})$. The total noise input voltage at the terminals of the noiseless amplifier is $(i_s + i_{nt} + e_{nt}Y_s)\,A_t/(Y_s + Y_{it}) + (i_{na} + e_{na}Y_{ot})/(Y_{ot} + Y_{ia})$.

Of this, the noise voltage from the source is $i_sA_t/(Y_s + Y_{it})$, so that the overall noise figure F_o has the form:

$$F_o = 1 + \frac{\langle(i_{nt}+e_{nt}Y_s)\,(i_{nt}+e_{nt}Y_s)^*\rangle}{\langle i_s i_s^*\rangle} + \frac{\langle(i_{na}+e_{na}Y_{ot})\,(i_{na}+e_{na}Y_{ot})^*\rangle}{A_t^2(|Y_{ot}+Y_{ia}|^2/|Y_s+Y_{it}|^2)\,\langle i_s i_s^*\rangle} \quad \ldots\ldots\ldots\ldots (11)$$

When connected to a source admittance Y_{ot} at temperature T, the noise figure of the amplifier alone is:

$$F_a = 1 + \langle(i_{na}+e_{na}Y_{ot})\,(i_{na}+e_{na}Y_{ot})^*\rangle/\langle i_{ot}i_{ot}^*\rangle \quad \ldots. (12)$$

where $\langle i_{ot}i_{ot}^*\rangle \cong 4kTG_{ot}\Delta f$ is the mean square thermal noise current generator associated with G_{ot}. It follows from equations (11) and (12) that:

$$F_t = F_o - \frac{(F_a - 1)}{A_t^2}\,\frac{|Y_s + Y_{it}|^2}{|Y_{ot} + Y_{ia}|^2}\,G_{ot}/G_s \quad \ldots. (13)$$

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Effect of Minority-Carriers on the Dynamic Characteristic of Parametric Diodes

By I. Hefni*

In the course of work on parametric devices using variable capacitance diodes an anomalous V-I characteristic for the pn junction has been observed. The effects of this is shown by oscillograph traces and it is suggested that it is due to the existence of minority carriers.

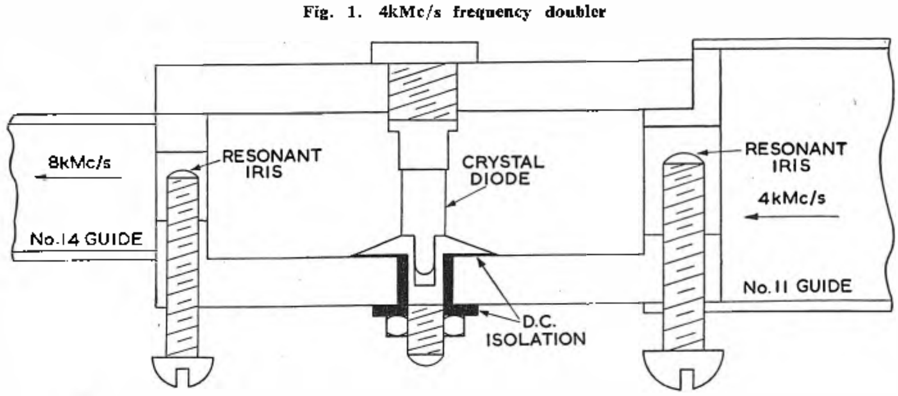
(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

THE theory of the variable capacitance pn junction is based on the assumption of the two majority-carriers being separated by a voltage sensitive space charge region, and neglects the existence of the minority-carriers. This might be true if the signal level is small or if the frequency is low compared with the minority-carriers' lifetime.

In the course of work on parametric devices using variable capacitance diodes, an anomalous dynamic *V-I* characteristic for the pn junction type has been observed which would suggest that the representation of these diodes by a simple circuit, consisting of a variable capacitance shunted by a small leakage conductance both in series with a small spreading resistance, is not valid for large signals at high frequencies, at least under certain operating conditions, as for instance when the diode is self-biased.

When a silicon mesa-type pn junction diode (Microwave Associates MA460C) was used as a harmonic generator in a resonant cavity (Fig. 1), the efficiency of the diode as a doubler (4kMc/s to 8kMc/s) was found to be highest when it was biased to conduct in the reverse direction (about 0·5mA). It was first thought that this reverse current—being much higher than the normal saturation reverse current—was an avalanche breakdown current†; but when the bias voltage was measured it was found to be much less than the breakdown voltage of the diode (1·5V compared with 8·5V breakdown voltage). The conduction current was then plotted against the bias voltage in the presence of an r.f. source (about 50mW). This resulted in the curve of Fig. 2, which showed a 'secondary' breakdown current and a negative d.c. resistance in the region *A*. When the r.f. power was reduced to a certain level (about 10mW), the irregularity in the characteristic disappeared. This is shown in the photographs of Fig. 3 (taken on a Tektronix transistor-curve tracer type 575), where trace (a) represents the normal static *V-I* characteristic with no r.f., and trace (b) shows the increase in both the forward and reverse breakdown (avalanche) currents when a small r.f. voltage was superimposed on the d.c. sweeping voltage. As the r.f. voltage was further increased a 'pip' in the reverse (saturation) current started to appear and to

Fig. 1. 4kMc/s frequency doubler



* *Marconi's Wireless Telegraph Co Ltd.*

† *R. S. Ohl et al. of B.T.L. also reported an increase in the output of the 2nd and 3rd harmonics of millimeter waves, when an ion bombarded silicon diode was operated near the reverse current break.*

grow in amplitude with the increase of the r.f. (trace [c]).

Obviously this irregularity in the dynamic characteristic will not occur for point-contact diodes; this is clear from the traces of Fig. 4, which correspond to those of Fig. 3,

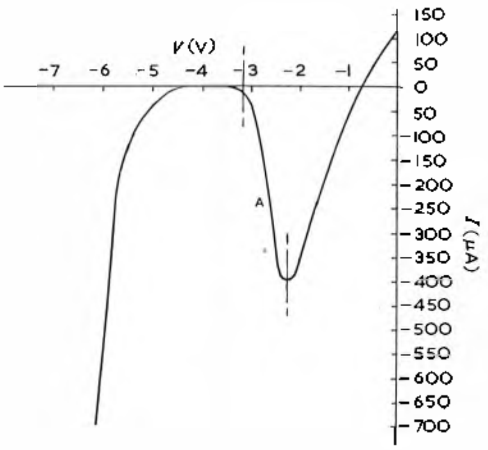


Fig. 2. Dynamic characteristic for pn junction diode at 4kMc/s

when the junction was replaced by a silicon point-contact diode (BTH CS3A).

The fact that the 'secondary' breakdown does not appear in the static characteristic would exclude the possibility that it might be a Zener current produced by internal

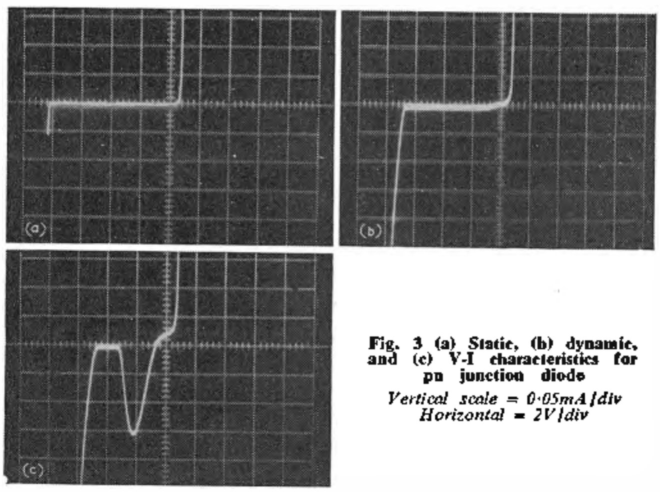


Fig. 3 (a) Static, (b) dynamic, and (c) V-I characteristics for pn junction diode

Vertical scale = 0·05mA/div
Horizontal = 2V/div

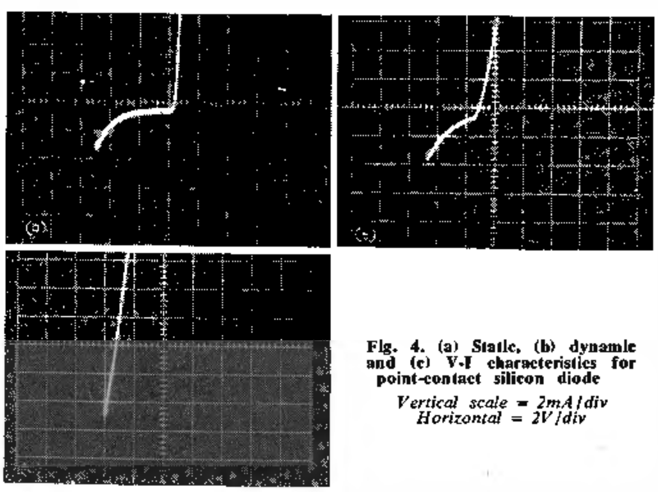


Fig. 4. (a) Static, (b) dynamic and (c) V-I characteristics for point-contact silicon diode

Vertical scale = 2mA/div
Horizontal = 2V/div

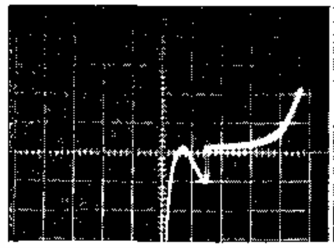


Fig. 5. Dynamic characteristic for a junction diode representing two negative resistance regions

Vertical scale = 0·05mA/div
Horizontal = 2V/div

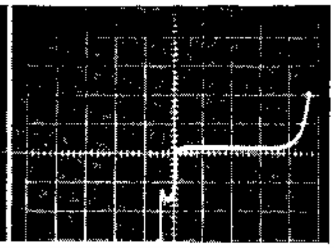


Fig. 6. Dynamic characteristic showing the self oscillations at one of the negative resistance regions

Vertical scale = 0·2mA/div
Horizontal = 10V/div

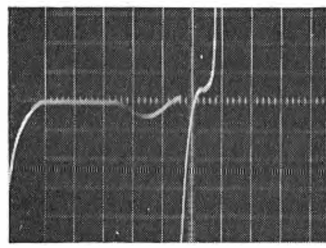


Fig. 7. Irregularity in the forward current characteristic for junction diodes

Vertical scale = 0·05mA/div
Horizontal = 2V/div

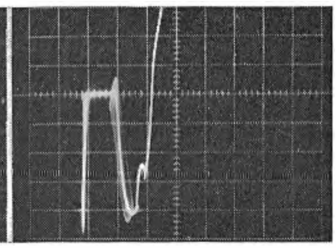


Fig. 8. Dynamic characteristic for welded contact germanium diode

Vertical scale = 0·05mA/div
Horizontal = 5V/div

field emission, and the only obvious explanation is that it might be due to stored minority-carriers which are injected during the positive half-cycle of the r.f. and diffuse back when the voltage is reversed. Since the time required for 'recovery' (which depends on the minority-carrier lifetime) for these diodes may be of the order of a fraction of a microsecond, this phenomenon should not occur at frequencies below a few megacycles. This was verified experimentally and the frequency limit was found to be around 10Mc/s.

Further experiments were carried out on different diodes in resonant and non-resonant circuits and at different frequencies. In these experiments all the junction diodes showed some sort of irregularity in the dynamic characteristic above a few megacycles, and for high r.f. levels the irregularities often resulted in more than one negative-resistance region as shown in Figs. 5 and 6. (Fig. 6 shows the self oscillations due to the negative-resistance when the resistance of the d.c. circuit was very much reduced). The irregularities will also occur at forward currents with relatively small r.f. voltage (Fig. 7). On the other hand, the variable capacitance welded contact germanium diode (Mullard OA47) would not show any irregularity when the bias was swept in the forward direction, but for reverse bias and large r.f. signal the diode did conduct a forward current as shown in Fig. 8.

At present the mechanisms causing all the different irregularities in the dynamic characteristic have not been clearly resolved although it is believed that the minority-carriers are responsible for most of them. However, the implications of these irregularities, particularly the negative-resistance regions, on the use of junction diodes in parametric devices are numerous, and these will be the subject of further studies.

Acknowledgment

The author wishes to thank L. De Tullio, I. G. Cressell, A. Guilliams and Dr. S. Forte for helpful discussions and assistance, and A. E. Sarson for providing the silicon junction diode.

A Circuit for the

Protection of a Stabilized Transistor Power Supply

By H. Kemhadjian* and A. F. Newell*

Higher-power transistors are being used in circuits more and more frequently. They often need to be operated near maximum permissible ratings and so might be rapidly destroyed by a sudden overload on the circuit. There is thus a requirement for a protection circuit which will switch off the stabilized transistor power supply more rapidly than the fastest existing relays.

The overload protection circuit described in this article can switch off the supplies in less than 50μsec. Additionally the current at which the protection circuit operates can be set to much lower currents than the full available output of the power supply. Thus experimental circuits, powered by the stabilized supply, will also be protected against any damage arising from faults within those circuits.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

THE increased use of high-power transistors in recent circuits necessitates power supplies capable of supplying the circuits. These include stabilized power supplies in which it is logical to use transistors and to use them as economically as possible. When transistors are operated near their peak current and voltage ratings, some protection must be provided against overload. The circuit described in this article is capable of providing protection against overload of the power supply and can be easily incorporated in a supply without affecting its normal operation.

In addition, the current, at which protection against overload is provided, can be varied over a wide range so that circuits can be adequately protected. This is especially useful for laboratory power supplies where experimental circuits may need protection against such conditions as thermal runaway.

Principles of Operation

Fig. 1 is a simplified basic circuit for a series stabilizer[1,2]. The emitter-follower, X_4, has the unstabilized voltage applied to its collector and by means of a voltage amplifying transistor, X_3, the output voltage may be stabilized to some value of reference voltage V. Any change in output voltage will be compensated by the application of this change of voltage, suitably amplified, to the base of the emitter-follower X_4.

If the output is short-circuited, the current through the emitter-follower X_4 will be limited only by the internal resistance of the unstabilized supply. If, for other reasons, that resistance must be small, large values of overload current must be allowed for. The protection circuit acts by cutting-off the emitter-follower by means of a small positive voltage applied to its base.

Fig. 2 shows a bistable circuit which performs the operation outlined above. To the left of the broken line is a conventional Eccles-Jordan bistable circuit. The collector of transistor X_2 is connected to the base of the emitter-follower X_4; X_2 is normally in the cut-off condition. The low leakage silicon diode (MR_1) ensures that changes of leakage current through X_2 do not affect the stability of the supply. In normal operation the anode of MR_1 is more negative than the maximum negative voltage on the base of X_4.

In the emitter lead of X_1 is a low value resistor through which the load current flows and across which, therefore, is developed a voltage proportional to the load current. The circuit is arranged so that, when the load current exceeds a certain value, the bistable circuit is triggered into

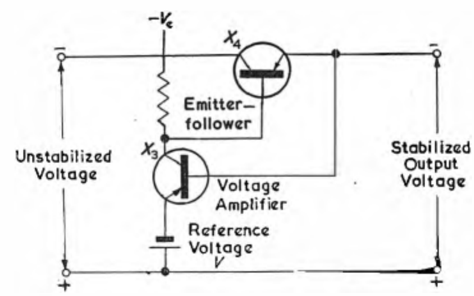


Fig. 1. Basic stabilizer circuit

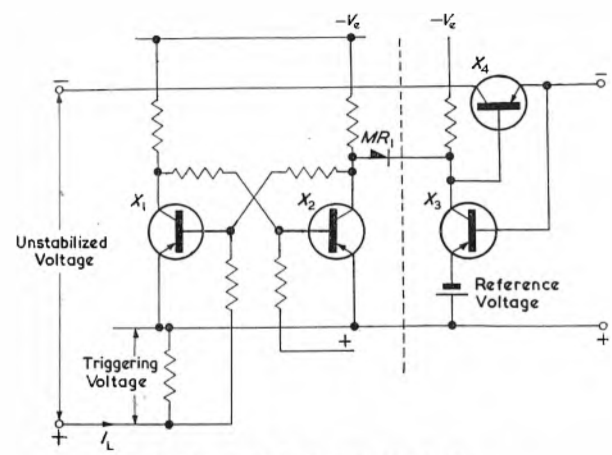


Fig. 2. Eccles-Jordan bistable circuit

its other stable state. X_2 'bottoms', causing MR_1 to conduct and thus the emitter-follower is cut-off.

In the basic circuit (Fig. 3) the emitter of X_2 is connected to a positive voltage to allow for the small voltage drop across X_2 and MR_1. This arrangement ensures that the emitter-follower is definitely cut-off. The base of X_2 is connected to a voltage which is more positive than its emitter, to maintain the transistor in its cut-off state over

* *Mullard Ltd.*

The authors have applied for a patent (British Patent Application No. 12431/59), 'Improvements in or relating to voltage stabilizing circuits'.

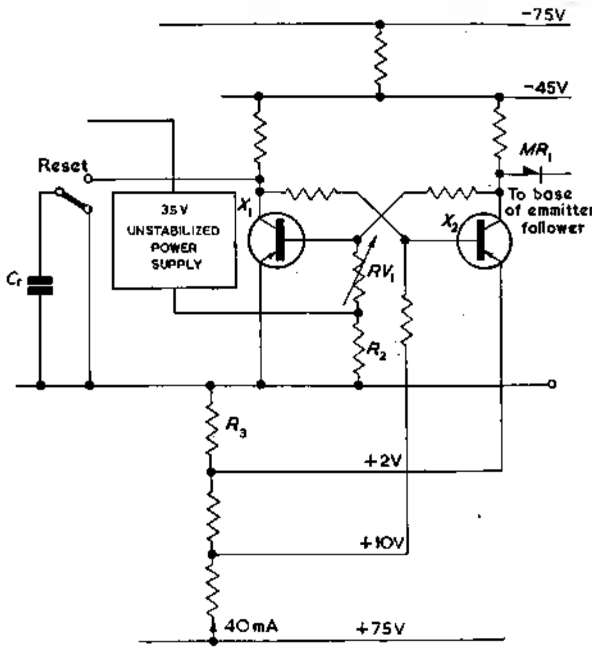


Fig. 3. Basic protection circuit

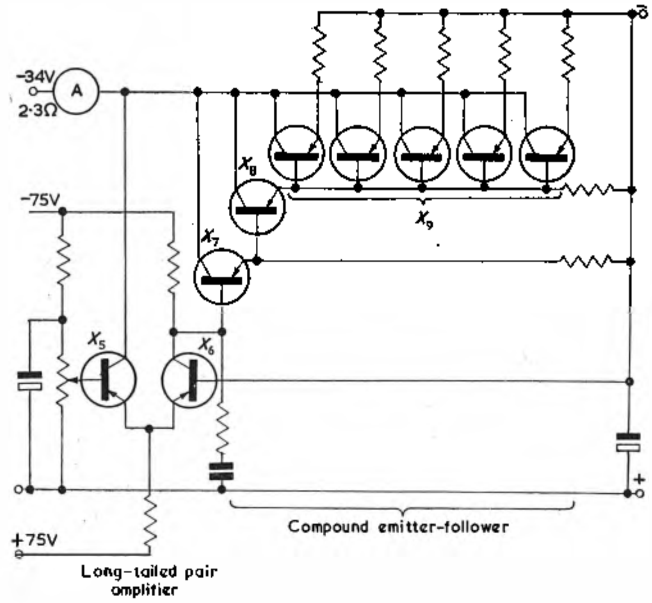


Fig. 4. Circuit for a practical power supply

the whole range of ambient temperature for which the circuit is designed.

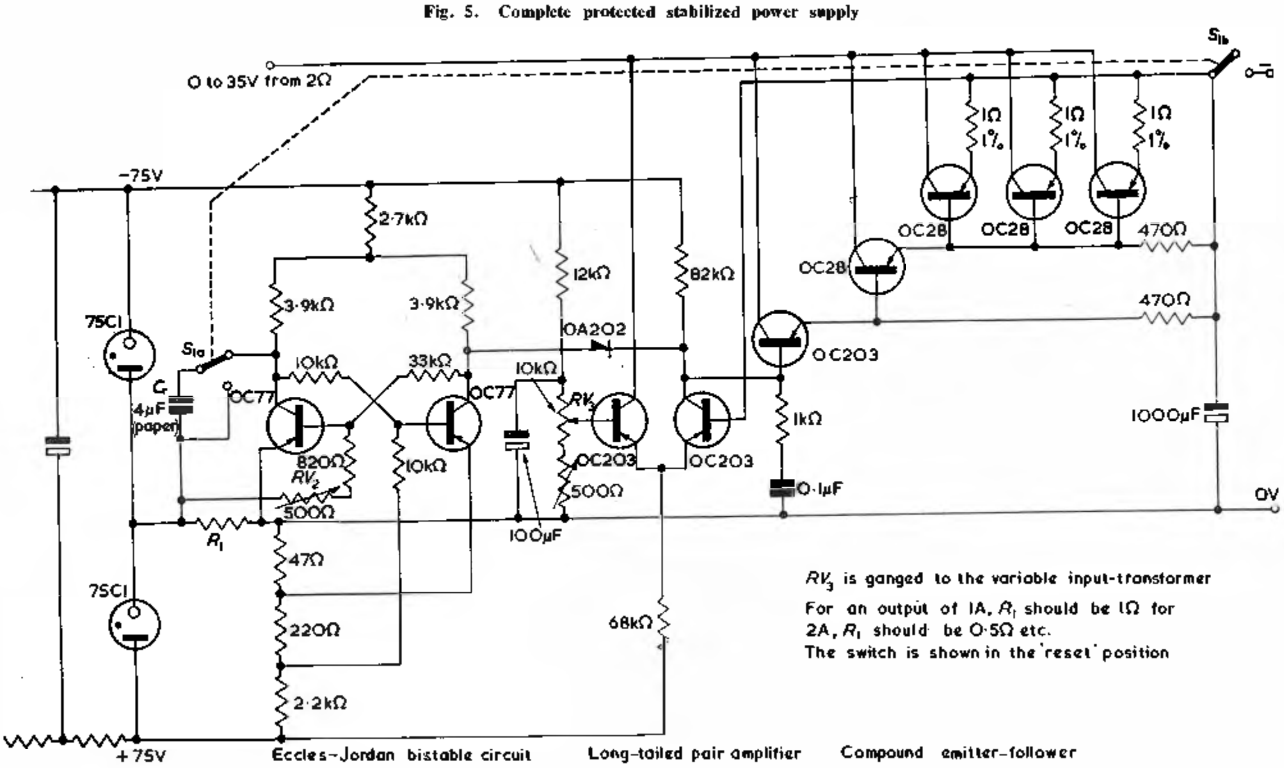
The current through the voltage divider chain must be relatively large (40mA) because, when the circuit triggers to its other stable state, the base current of the emitter-follower will flow through the bottomed emitter-follower X_4 and through R_3 to the common line. If this latter current is comparable with that through the divider chain it can inhibit the changeover action of the bistable circuit after that circuit has been triggered.

After a fault has been cleared, the capacitor C_r can be connected momentarily to the collector of X_1 by a spring loaded switch, to reset the circuit. The charging current, which flows on connexion of C_r, produces a positive triggering voltage at the base of X_2, cutting-off this transistor.

Practical Example

Fig. 4 shows a practical power supply based on the circuits of Fig. 1. The power supply is designed to provide an output current of up to 3A at a stabilized voltage of 27V. In place of the single-transistor emitter-follower there is a compound emitter-follower (X_{7-9}). X_9 itself is composed of five transistors connected in parallel thus providing the high current required, while the transistors all operate within their individual ratings. The voltage-amplifying transistor (X_3 in Fig. 1), is replaced by a long-

Fig. 5. Complete protected stabilized power supply



tailed-pair amplifier, X_5 and X_6. Fig. 5 shows how the protection circuit can be incorporated in this power supply.

This complete circuit (Fig. 5) is given by way of example only; the effects of component tolerances have not yet been considered. The supply is rated for 3A maximum output above which current the protection circuit operates. The output transistor (X_9, Fig. 4) thus only needs three individual OC28 transistors to supply this current, because short-circuit conditions will be prevented.

There are certain important requirements in the design of a protected circuit.

(1) The transistors of the compound emitter-follower must be able to withstand a collector to emitter voltage equal to the maximum unstabilized voltage available. This voltage will appear across the emitter-follower when that transistor combination is cut-off by the operation of the protection circuit.

(2) Additional supply voltages must be provided for the protection circuit.

The protection circuit needs a triggering voltage, developed across R_1, of approximately 1V. Different values of R_1 can be switched in to vary the load current at which triggering occurs. Standard values can be used for R_1, and RV_3 adjusted so that triggering occurs at the exact value of load current required.

If there is a fault in the load the discharge current of the output capacitor (only limited by the resistance of the fault) is added to the current flowing through the output terminals. It is possible that the combined current pulse would damage the circuit in which the fault has occurred. For this reason the output capacitor might need to be reduced to 100μF.

The protection circuit is inoperative until the capacitor C_r has been charged. The value of this capacitor is dependent on the value of the output capacitor, so that 4μF is the minimum for a 1 000μF output capacitor. C_r may, of course, be reduced if the output capacitor is reduced.

If the stabilizer circuit is triggered 'on' with a fault still present the protection circuit is then inoperative for a few milliseconds, and the dissipation in the series transistors could become excessive, because the full unstabilized voltage acts across the output transistors causing a large current to flow through them. To prevent this a double pole switch is used for the reset operation, so that one output terminal is disconnected at the same time as the capacitor C_r is connected across X_1.

Supply voltages for the protection circuits are derived from dropping resistors across the high current stabilizers. An oscillogram was taken across the triggering resistor R_1, with an output voltage of 25V, and the circuit set to trigger at 1A. The peak current was 2A and the time of the pulse about 25μsec.

Performance of the Practical Circuit

The figures quoted below have not been confirmed with maximum spreads of components, but they give some idea of the expected performance.

Changing the value of R_1 adjusts the triggering current between 20mA and 3A. With R_1 equal to 1Ω, the variable resistor was adjusted until the circuit triggered at exactly 1A. The variation of a given triggering current over a temperature range of 20 to 45°C was less than 10 per cent. The value of triggering current was nearly constant for all settings of output voltage down to less than 1V.

Acknowledgment

The authors wish to thank the directors of Mullard Limited for permission to publish this article.

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A Telemetry System for Car Testing

A radio telemetry system with considerable industrial applications has recently been developed by the electronic research department of Sir W. G. Armstrong Whitworth Aircraft Ltd, Coventry.

Known as DIDAS (Dynamic Instrumentation Data Automobile System) it is based on the radio telemetry techniques developed for guided missiles and when applied to the track testing of prototype cars considerable saving in design time can be achieved.

This new method of testing is a great step forward from existing methods which involve carrying either bulky recording equipment in the test car, giving an unwanted weight penalty, or carrying a technician to record readings from meters, with the consequent limitation on the number of readings a human can accurately record, particularly in a bumping, swaying vehicle.

The DIDAS telemetry system enables measurements of various physical factors (strain, pressure, position, vibration, temperature, etc.) to be taken from up to 23 different sources on the car while it is in motion. The signals thus derived are fed into a multiplexing switch which combines them into a single coded output which is fed into a transmitter.

The equipment carried in the vehicle is light (only 20 to 30lb) and is small enough to be placed on the seat beside the driver, who can also switch it on or off when required. All the power requirements for the transmitting equipment can be obtained from the vehicle's battery.

The transmitter, which operates in the u.h.f. band, is connected to a small stub aerial on the roof of the vehicle and the range is about two miles with the present low power output.

At the receiving site the signals, after processing and amplification, are fed into recording apparatus so as to give a permanent record of the performance of the vehicle while under test. They can also be fed into a computer if required.

As they are recorded, the signals are also displayed on meters so that the observer can examine the results of the tests as they are carried out and which can be further displayed on a cathode-ray oscilloscope.

A v.h.f. radio telephone link provides communication between the driver and the observer.

The mobile equipment in a motor-car

Analysis of the Split Load Stage
(Transistorized)

By H. Pfyffer*

In the first part of this article the split load stage is investigated as a single stage with local negative feedback. Expressions for the current and the voltage gain are derived; the output impedance and the output return loss are also considered.

In the second part a three stage feedback amplifier using the split load stage at the output is analysed with respect to the total gain, the open loop gain and the output impedance. Typical numerical values have been introduced in two examples to give an idea of the behaviour of the split load stage.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

THE split load stage is useful as a class-A output stage. By using this configuration it is possible to apply local feedback to the stage without dissipating power in a feedback impedance, since the load itself is used to provide negative feedback[1]. The type of feedback is a mixture of current and voltage feedback. The split load stage is intermediate between the common collector and the common emitter stage.

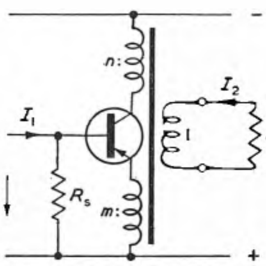

Fig. 1. The split load stage

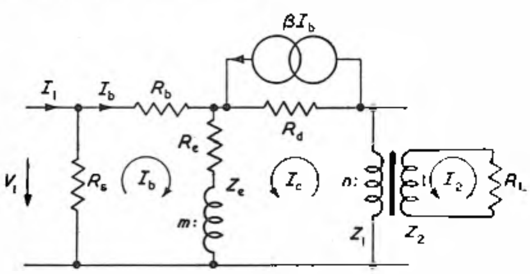

Fig. 2. Equivalent circuit

It can also be used as the output transistor of a multistage feedback amplifier.

Single Stage with Local Negative Feedback

In the first part of the article the split load stage is investigated working between a source of a given impedance R_s and a given load R_L. The source can be a preceding transistor stage.

The Equivalent Circuit and Definitions

The split load stage is shown in Fig. 1 and its equivalent circuit in Fig. 2. The following relations are true for the transformer (Z_1 being the shunt impedance $j\omega L_1$):

$$\left.\begin{aligned} &Z_2 = (Z_1/n^2) \quad Z_e = (m^2/n^2)Z_1 \quad Z_{e1} = (m/n)Z_1 \\ &\qquad\qquad\qquad\qquad\qquad\qquad = \sqrt{(Z_eZ_1)} \\ &Z_{12} = \sqrt{(Z_1Z_2)} = (Z_1/n) \quad Z_{2e} = \sqrt{(Z_2Z_e)} = Z_1(m/n^2) \end{aligned}\right\} \quad \ldots\ldots (1)$$

* *The General Electric Co. Ltd. Coventry.*

Application of mesh analysis yields the following set of equations:

$$\left.\begin{aligned} V_1 &= I_b(R_b+R_e+Z_e) + I_c(R_e + Z_e + Z_{1e}) - I_2Z_{2e} \\ 0 &= I_b(R_e+Z_e-\beta R_d+Z_{e1})+I_c(R_e+Z_e+Z_1+R_d+2Z_{e1}) \\ &\qquad - I_2(Z_{e2} + Z_{21}) \\ 0 &= -I_bZ_{e2} + I_c(-Z_{12} - Z_{e2}) + I_2(Z_2 + R_L) \end{aligned}\right\} \quad . \quad (2)$$

The first equation can be modified by using the relation:

$$V_1 = R_s(I_1 - I_b) \quad \ldots\ldots\ldots\ldots\ldots (3)$$

and the equations can be rewritten in matrix form:

$$\begin{vmatrix} I_1R_s \\ 0 \\ 0 \end{vmatrix} = \begin{vmatrix} R_s+R_b+R_e+Z_e & R_e + Z_e + Z_{1e} & -Z_{2e} \\ R_e+Z_e-\beta R_d+Z_{e1} & R_e+Z_e+Z_1+R_d+2Z_{e1} & Z_{e2}-Z_{21} \\ -Z_{e2} & -Z_{12} - Z_{e2} & Z_2 + R_L \end{vmatrix} \cdot \begin{vmatrix} I_b \\ I_c \\ I_2 \end{vmatrix} \quad \ldots\ldots\ldots\ldots (4)$$

The impedance matrix represents at the same time the circuit determinant Δ.

Principal Symbols

R_e	=	Emitter resistance } of T equivalent circuit
R_b	=	Base resistance } of T equivalent circuit
R_d	=	Collector resistance
β	=	Short-circuit current gain base-collector
Z_1, Z_e, Z_2	=	Shunt impedances of transformer
Z_{1e}, Z_{2e}, Z_{12}	=	Mutual impedances of transformer
R_s	=	Source impedance for split load stage
R_L	=	Load impedance of split load stage
m	=	Number of turns of emitter winding
n	=	Number of turns of collector winding
l	=	Number of turns between feedback tapping point and end of winding away from collector
γ	=	Current gain
μ	=	Voltage gain
K	=	Power gain
Δ	=	Circuit determinant
F	=	Return difference
T	=	Return ratio
ρ	=	Return loss of output impedance
Z_o	=	Output impedance of split load stage
Z_{i1}, Z_{i2}	=	Input impedance of preceding stages
Z_{o1}	=	Output impedance including external collector resistance of 1st and bias resistors of 2nd stage
R_f	=	Current feedback resistor

The Current Gain

Solving the circuit equations by the determinant method, the current gain is found to be

$$\gamma = (I_2/I_1) = \frac{R_s\Delta_{13}}{\Delta} \quad \ldots\ldots\ldots\ldots\ldots (5)$$

where Δ_{13} is the cofactor of Δ by deleting the first row and the third column.

In all practical cases Z_1 is chosen to be very much larger than the other circuit impedances (to minimize the transformer losses). In addition to this assumption, the following inequalities are satisfied in most practical cases:

$$\left.\begin{array}{l}(R_e/R_d) \ll 1 \\ \dfrac{R_s + R_b}{R_d} \ll \beta\end{array}\right\} \quad \ldots\ldots\ldots\ldots\ldots (7)$$

The current gain then reduces to a useful form.

$$\gamma = \frac{m(1+\beta)+n\beta}{1+\dfrac{R_b+(1+\beta)R_e}{R_s}+(R_L/R_s)\left(\dfrac{R_s+R_b+R_e}{R_d}n^2+m^2(1+\beta)+mn\beta\right)} \quad \ldots (8)$$

The Voltage Gain

The voltage gain can be easily obtained from the expression for the current gain. If it is assumed that equation (2) is solved for I_2/V_1, it is obvious that Δ remains the same as for equation (5), except that R_s is missing, hence:

$$I_2/V_1 = (\Delta_{13}/\Delta)_{(R_s=0)} \quad \ldots\ldots\ldots\ldots\ldots (9)$$

It can be further noted that

$$V_2 = I_2R_L \quad \ldots\ldots\ldots\ldots\ldots\ldots (10)$$

The voltage gain is defined as

$$\mu = (V_2/V_1) \quad \ldots\ldots\ldots\ldots\ldots (11)$$

Combining equations (9), (10) and (11) and observing equation (7), the voltage gain is obtained in its simplified form:

$$\mu = \frac{(m(1+\beta)+n\beta)R_L}{R_b+(1+\beta)R_e+R_L\left[\dfrac{R_b+R_e}{R_d}n^2+m^2(1+\beta)+mn\beta\right]} \quad \ldots\ldots\ldots\ldots (12)$$

The Amount of Feedback Applied

Once the circuit determinant is found, it is relatively easy to find the amount of feedback applied to an amplifier. According to Bode[2] the return difference $F = 1 + T$ is given by

$$F = (\Delta/\Delta^\circ) \quad \ldots\ldots\ldots\ldots\ldots (13)$$

Where T is the return ratio, and Δ° is obtained from Δ with the transconductance of the stage under consideration put equal to zero. Δ° in this case is simply obtained by putting β equal to zero. Considering equations (4) and (13) and assuming Z_1 to be very large, F is thus found to be

$$F = 1 + \frac{\beta[R_e + m^2R_L(1+(n/m))]}{R_s+R_b+R_e+R_L\left[n^2\dfrac{R_s+R_b+R_e}{R_d}+m^2\right]} \quad \ldots\ldots\ldots\ldots (14)$$

Output Impedance and Output Return Loss

The output impedance Z_o is determined by V_2/I_2 in the absence of R_L, it can be calculated by solving the circuit equations (2) for I_2 under the assumption that $I_1 = 0$ and $V_2 = I_2R_L$ with V_2 being considered as the driving source. Thus

$$Z_o = (\Delta_{(R_L=0)})/\Delta_{33} \quad \ldots\ldots\ldots\ldots (15)$$

$$Z_o = R_d\frac{R_s+R_b+(1+\beta)R_e}{(m^2+2mn+n^2)(R_s+R_b)+R_d[(1+\beta)m^2+\beta mn]} \quad \ldots\ldots\ldots\ldots (16)$$

The output return loss ρ with respect to R_L

$$\rho = \frac{R_L + Z_o}{-R_L + Z_o} \quad \ldots\ldots\ldots\ldots (17)$$

can be obtained again by using the circuit determinant.

Combining equations (4) and (17), it follows that

$$\rho = \frac{\Delta_{(R_L=0)} + R_L\Delta_{33}}{\Delta_{(R_L=0)} - R_L\Delta_{33}} = \Delta/\Delta_{-R_L} \quad \ldots (18)$$

where Δ_{-R_L} is equal to Δ with the exception that R_L is replaced with its negative value. For the split load stage ρ is calculated to be

$$\rho = 1 + \frac{2}{\dfrac{R_s+R_b+(1+\beta)R_e}{R_L[n^2[(R_s+R_b+R_e)/R_d]+m^2(1+\beta)+mn\beta]} - 1} \quad \ldots\ldots\ldots\ldots (19)$$

If ρ is brought into relation with F, the following expression is obtained:

$$\rho = \frac{1+T}{1+AT} \quad \ldots\ldots\ldots\ldots\ldots (20)$$

where

$$A = \frac{1-(R_L/R_e)\left[m^2+mn+(2/\beta)\left(m^2+n^2\dfrac{R_s+R_b+R_e}{R_d}\right)\right]}{1+(R_L/R_e)(m^2+mn)} \quad \ldots\ldots\ldots\ldots (21)$$

If it can be assumed that

$$m^2R_L \gg R_e \quad \ldots\ldots\ldots\ldots\ldots (21a)$$

and

$$R_e \ll R_s \quad \ldots\ldots\ldots\ldots\ldots (21b)$$

then A simplifies to

$$A = -\frac{1+(n/m)+(2/\beta)\left(1+(n^2/m^2)\cdot\dfrac{R_s+R_b}{R_d}\right)}{1+(n/m)} \quad \ldots\ldots\ldots\ldots (22)$$

The return loss ρ is a function of m and n. If, for example, n is specified, the value of m required to yield a maximum value of ρ, i.e. $\rho = \infty$, can be determined.

The following equation is obtained:

$$m_{\rho(\max)} = \frac{n\beta}{2(1+\beta)}\left[\sqrt{\left[1+4\frac{1+\beta}{\beta^2}\left(\frac{R_s+R_b+(1+\beta)R_e}{n^2R_L}-\frac{R_s+R_b+R_e}{R_d}\right)\right]}-1\right] \quad \ldots\ldots\ldots\ldots (23)$$

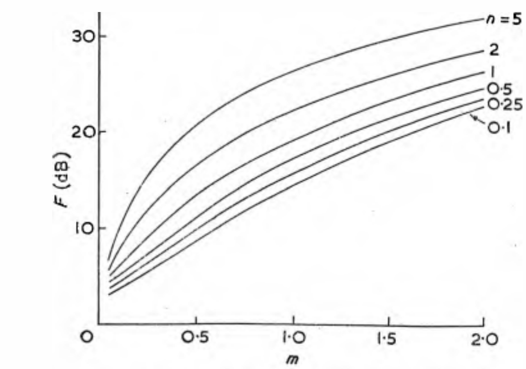

Fig. 3. (a) F in function of m

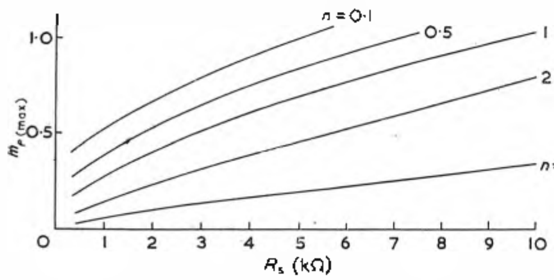

Fig. 3. (b) $m_{\rho(max)}$ in function of R_s

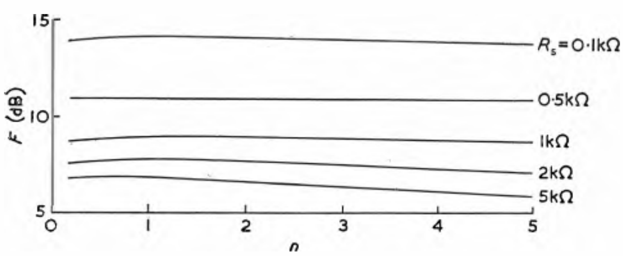

Fig. 3. (c) F in function of n, m equal to corresponding $m_{\rho(max)}$

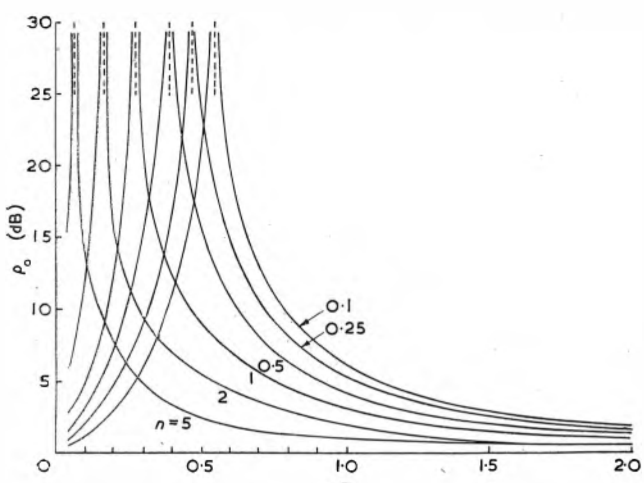

Fig. 3. (d) Return loss ρ_o in function of m

Example

To give an idea of the behaviour of the split load stage, the following set of curves (Fig. 3) have been calculated for a typical case:

(a) F as a function of m, with n as a running parameter.

(b) $m_{\rho(max)}$ as a function of R_s with n as a running parameter.

(c) F as a function of n/R_s with n as a running parameter, $m = m_{\rho(max)}$.

(d) ρ_o as a function of m, with n as a running parameter.

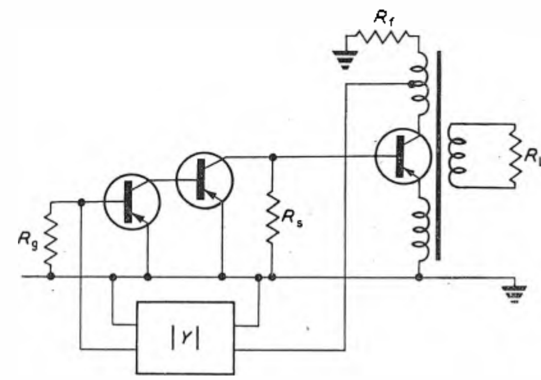

Fig. 4. The three stage feedback amplifier

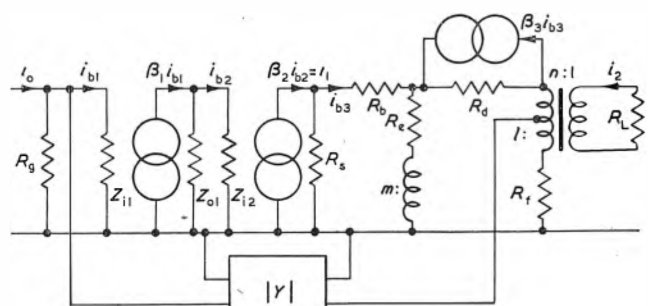

Fig. 5. Equivalent circuit

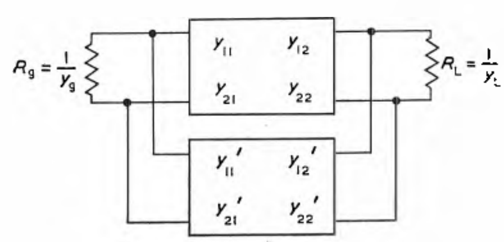

Fig. 6. Four-terminal network representation of feedback amplifier

Although these curves are not generally applicable, nevertheless they indicate the circuit behaviour of a split load stage.

The transistor is assumed to have the following parameters:

$$\beta = 50,\ R_e = 10\Omega,\ R_b = 200\Omega,\ R_d = 10\text{k}\Omega$$

load and source are:

$$R_s = 1\text{k}\Omega,\ R_L = 100\Omega$$

It can be seen that equations (7) and (21b) are satisfied.

Note

It is of interest to note that all the expressions obtained can be converted into the formulae for either the common collector stage or the common emitter stage by letting n or m go to zero respectively.

Three Stage Feedback Amplifier

In this part the split load stage is assumed to be the output stage of a feedback amplifier. The two preceding transistors are of the common emitter configuration. The feedback is of the bridge type at the output and of the shunt type at the input, i.e. it is connected to the base of the first transistor.

Expressions for the open loop gain and the output impedance with feedback are derived.

The Equivalent Circuit, Gain Without Feedback

The circuit under consideration is represented in Fig. 4 and the equivalent circuit in Fig. 5.

It will be assumed that the feedback network is given in terms of the Y matrix parameters.

The current gain without feedback can easily be obtained by the following manipulations:

$$\gamma_1 = (i_{b1}/i_o) = \frac{R_g}{R_g + Z_{i1}},\ \gamma_2 = (i_{b2}/i_{b1}) = \frac{\beta_1 Z_{o1}}{Z_{o1} + Z_{i2}}\ \ \gamma = (i_2/i_1),\ \gamma_o = \gamma_1\gamma_2\gamma$$

(see equation (8)).

$$\gamma_o = \frac{R_g\beta_1 \,.\, Z_{o1} \,.\, \beta_2[m(1+\beta_3)+n\beta_3]}{(R_g+Z_{i1})(Z_{o1}+Z_{i2})\left\{\left[1+\dfrac{R_b+(1+\beta_3)R_e}{R_s}\right]+R_L/R_s\left[\dfrac{R_s+R_b+R_e}{R_d}n^2+m^2(1+\beta_3)+mn\beta_3\right]\right\}} \quad \ldots\ldots\ldots \quad (24)$$

The voltage gain of the amplifier can be obtained using the same method as for μ of the split load stage. μ_o is obtained as:

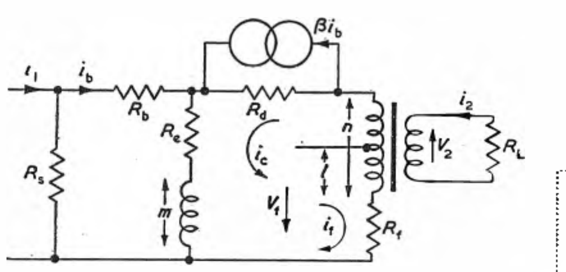

Fig. 7. Equivalent circuit with feedback bridge

$$\mu_o = \frac{Z_{o1}\beta_1\beta_2[m(1+\beta_3)+n\beta_3] \,.\, R_L}{Z_{i1}(Z_{o1}+Z_{i2})\left\{R_b+(1+\beta_3)R_e+R_L\left[\dfrac{R_b+R_e}{R_d}n^2+m^2(1+\beta_3)+mn\beta_3\right]\right\}} \quad \ldots\ldots\ldots\ldots \quad (25)$$

The power gain K_o is given as the product

$$K_o = \gamma_o\mu_o \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (26)$$

The Return Difference of the Amplifier

As mentioned previously, the return difference is given by

$$F = (\Delta/\Delta_o)$$

In terms of the Y matrix the circuit determinant is obtained from Fig. 6, as follows:

$$|Y| = \begin{vmatrix} y_{11} + y_{11}' + y_g & y_{12} + y_{12}' \\ y_{21} + y_{21}' & y_{22} + y_L + y_{22}' \end{vmatrix} \quad \ldots\ldots\ldots\ldots \quad (27)$$

Δ is the determinant of this matrix, and

Δ^o is obtained when $y_{21} = 0$

The return difference yields to

$$F = 1 + \frac{y_{21}y_{21}'}{(y_{11}'+y_{11}+y_g)(y_{22}'+y_{22}+y_L) - (y_{21}')^2} \quad \ldots \quad (28)$$

For the bridge type of feedback the total admittance at the l-tapping of the output transformer has to be found, also the transfer admittance y_m from the input to the l-tapping, which has to replace y_{21} in equation (28). Once all the values are found with respect to the l-tapping, the circuit can be dealt with in the normal way by matrix analysis.

The short-circuit gain of the split load stage from the base to the l-tapping will be calculated first (Fig. 7).

A set of circuit equations, representing the meshes, can be written down. In most cases the following inequality is satisfied:

$$R_t \ll R_d \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (29)$$

To obtain $\gamma_L = i_t/i_1$ for $V_t = 0$, the circuit equations can be solved, and considering equations (7) and (29) in addition to the assumption that m and n are in the vicinity of one, γ_L yields:

$$\gamma_L \simeq \frac{\beta(nl+(R_t/R_L))+(1+\beta)ml}{\left\{1+\dfrac{R_b+(1+\beta)R_e}{R_s}\right\}\{l^2+(R_t/R_L)\}+R_t/R_s\,\{m^2(1+\beta)+mn\beta\}} \quad \ldots\ldots\ldots\ldots \quad (30)$$

The transfer admittance of the whole amplifier from input to l-tapping is then given by

$$y_m = \frac{Z_{o1} \,.\, \beta_1\beta_2\gamma_L}{Z_{i1}\,(Z_{i2} + Z_{o1})} \quad \ldots\ldots\ldots\ldots \quad (31)$$

The total admittance between the l-tapping and ground is obtained as follows (using again the same simplifications as for γ_L)

$$y_{22l} \simeq \frac{(m+n)^2\left(\dfrac{R_s+R_b+R_e}{R_d}\right)+m^2(1+\beta)+mn\beta+\dfrac{R_b+(1+\beta)R_e+R_s}{R_L}}{(R_b+(1+\beta)R_e+R_s)(l^2+(R_t/R_L))+R_t\{m^2(1+\beta)+mn\beta\}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (32)$$

Using equations (28) and (32), the open loop gain T can be rewritten as follows:

$$T = \frac{y_m \,.\, y_{21}'}{(y_g+y_{11}+y_{11}')(y_{22l}+y_{22}') - (y_{21}')^2} \quad \ldots\ldots\ldots \quad (33)$$

The Output Impedance with Feedback

The output impedance with feedback can be calculated in a convenient way by using Blackman's formula[2]

$$Z_{out}' = (F_{so}/F_{oo})\,[Z_o]_{y_m=0}$$

F_{so} is the return difference for $R_L = 0$ and F_{oc} the return difference for $R_L = \infty$, and $[Z_o]_{y_m=0}$ can be interpreted as the output impedance with no forward transmission. To satisfy this condition, β of the split load stage has to be zero.

$$[Z_o]_{y_m=0} = \frac{R_d(R_s + R_b + R_e)}{(m+n)^2\,(R_s+R_b+R_e) + m^2R_d} \quad \ldots\ldots\ldots\ldots \quad (34)$$

which is easily obtained from equation (16).

The return differences for open and for short-circuit conditions at the output have to be determined. In order to obtain expressions which are not too complicated, the following simplifications are assumed:

$$\left.\begin{array}{l} y_{22}' \ll (1/R_f),\ (y_{21}')^2 \ll (1/R_f)(y_{11}'+y_{11}+y_g) \\ \beta \gg 1,\ R_d \gg R_e + R_s + R_b \\ m \text{ and } n \text{ in the vicinity of one} \end{array}\right\} \quad \ldots\ldots (35)$$

These assumptions are justified for the major part of all applications. For T_{so} and T_{oc} the following formulae are then obtained:

$$T_{so} \simeq \frac{Z_{o1}\beta_1\beta_2\beta_3 y_{21}'R_s}{Z_{i1}(Z_{i2}+Z_{o1})\{(R_s+R_b+(1+\beta)R_e)(y_{11}+y_{11}'+y_g)l/R_f\}} \quad \ldots\ldots\ldots\ldots (36)$$

$$T_{oc} \simeq \frac{Z_{o1}\beta_1\beta_2\beta_3(nl+ml)y_{21}'R_s}{Z_{i1}(Z_{i2}+Z_{o1})\left\{\left(y_{11}'+y_{11}+y_g\right)\left(m^2(1+\beta_3)+mn\beta_3+(m+n)^2\dfrac{R_s+R_b+R_e}{R_d}\right)\right\}} \quad \ldots\ldots\ldots\ldots (37)$$

Introducing these expressions into Blackman's formula, the output impedance yields (under the assumption, that both T_{sc} and T_{oc} are large with respect to unity, which is usually true for the bridge feedback).

$$Z_o \simeq \frac{R_d}{\left\{(m+n)^2+m^2\dfrac{R_d}{R_s+R_b+R_e}\right\}} \cdot \frac{R_f\left\{m^2(1+\beta_3)+mn\beta_3+(m+n)^2\dfrac{R_s+R_b+R_e}{R_d}\right\}}{l(m+n)\{R_s+R_b+(1+\beta_3)R_e\}} \quad \ldots\ldots\ldots\ldots (38)$$

EXAMPLE

To give an idea of possible values of components which may occur in a three stage amplifier with the split load stage at the output and bridge feedback over the three transistors, the following typical parameters for the transistors are assumed:

Output stage (split load stage)

$\beta = 50,\ R_e = 10\Omega,\ R_b = 200\Omega,\ R_d = 10\text{k}\Omega,$

$R_s = 1\text{k}\Omega,\ R_L = 10\Omega,\ m = 0{\cdot}25,\ n = 2.$

The two preceding stages may have identical parameters:

$\beta_1 = \beta_2 = 50,\ Z_{i1} = Z_{i2} = 2\text{k}\Omega,\ Z_{o1} = 30\text{k}\Omega,$

$R_g = 1\text{k}\Omega.$

The gain, the feedback bridge at the output and the feedback network will be considered next.

Using formula (8), the current gain is obtained as $\gamma = 21{\cdot}5 \times 10^3$, and equation (12) gives a voltage gain of $\mu = 3{\cdot}2 \times 10^3$. Combining the current gain with the voltage gain yields the power gain which in this case is $K = 6{\cdot}9 \times 10^7$ or 78dB.

The output impedance with feedback should be 100Ω in order to match it to the load. In the formula (38) for Z_o all the values except R_f and l are known. (It is assumed that equation (35) is true, and therefore, y_{11}', y_{12}', y_{21}' and y_{22}' do not appear as unknowns in equation (38)). The ratio of the two unknowns can be determined as $R_f/l = 7{\cdot}35\Omega$. To keep the losses in R_f small it is assumed to be 10Ω, and l will then be 1·36.

To calculate the required feedback network for an assumed return difference, γ_L and y_{22l} have to be obtained first by using equations (30) and (32) respectively.

These values are obtained as

$$\gamma_L = 58 \text{ and } y_{22l} = 12{\cdot}8\text{m mho}$$

Under the assumption that the feedback network consists of a single resistor only, which admittance is represented by $-y_{21}'$ (in this case $y_{11}' = y_{22}' = y_{21}' = -y_{12}'$) this is the only unknown in the expression for F or T. If T is 30dB, y_{21}' is then obtained to 9×10^{-3}m mho (by solving equation (33) for y_{21}'). This value represents a resistor of 110kΩ, and the inequalities (35) are therefore satisfied.

Conclusion

The split load stage has been analysed and the results show that, for a careful choice of the transformer turns ratios, a high return loss with respect to the load can be obtained. In addition 6 to 10dB of local negative feedback can be applied without introducing losses in available power.

The investigation of a three stage amplifier yielded formulae for the amount of overall feedback and for the output impedance. Some simplifications had to be considered, otherwise the obtained expressions would have been beyond practical interest.

It is interesting to note that the expressions (30), (32), (36) and (37) degenerate into the formulae for either pure current feedback or pure voltage feedback by letting l or R_f go to zero respectively. This can be done because the bridge feedback is intermediate between the two above

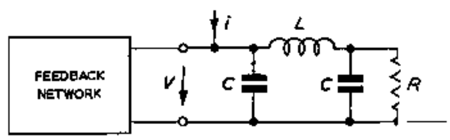

Fig. 8. Feedback from transformer tapping

mentioned types. Care has to be taken when the output impedance is considered. Formula (38) has been derived under the assumption that T_{oc} and T_{sc} are large compared with unity, which is not necessarily true for the current or the voltage feedback. Blackman's formula would have to be used in its complete form.

Similar methods of analysis may be carried out for the case where the overall feedback in a multi-stage amplifier is taken from the emitter or a tapping in the emitter winding of the transformer, instead of taking it from the collector side.

There is a further advantage of the split load stage as far as overall feedback, connected to the output transformer, is concerned. It is well known that a transformer can be represented by the equivalent circuit of Fig. 8 at high frequencies.

L represents the leakage inductance and C the winding capacitance. If the transformer is built such that $j\omega L \ll 1/j\omega C$ up to frequencies which have to be specified for each application, then the voltage which is applied to the feedback network is given by

$$V = i\,(j\omega L + R)$$

i.e. there is a phase advance. This is very desirable in order to help to stabilize a multi-stage feedback loop.

Acknowledgment

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Data on Ferrite Materials when Used as Rod Cores in Solenoidal Inductors and Transformers

By A. J. Baden Fuller*, M.A.

Charts are presented which permit comparison of permeability and loss for various ferrites over the frequency range 2Mc/s to 80Mc/s when used as rod cores for solenoidal inductors and transformers. Various rod specimens of ferrite material were used as the core of a coil whose impedance was measured. The results are presented in terms of a 'low level Q' and 'insertion factor' of the material.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

THE permeability of a ferrite core material may be considered as a complex quantity. Then the Q of the material is the ratio of the real (inductive) component to the imaginary (lossy) component of the complex permeability, where a series representation is used: or the reciprocal of this ratio where a parallel representation is used. The Q of the material is also the Q of a coil wound on a closed core of the material except to the extent that it is reduced by the losses in the coil.

Useful design charts have already been presented[1] where Q and the real component of series or parallel permeability are shown in terms of frequency for the Mullard Ferroxcube grades of ferrites. The curves were calculated from typical data on permeability and loss published[2] by Mullard Ltd. The manufacturers measured these properties on ring specimens but it is expected that many applications of ferrite materials in the megacycle range of frequencies will be in the form of cylindrical slugs inside solenoidal coils and further results, other than those on ring specimens, were needed to aid design of these coils.

The accompanying figures permit comparison of permeability and loss for some Ferroxcube grades of ferrite over the frequency range 2Mc/s to 80Mc/s when used as rod cores for solenoidal inductors and transformers. The results are presented in terms of Q and 'insertion factor', which are shown for different frequencies in the figures. The results from which these curves have been taken were obtained on short rods of ferrite where there is a large and indeterminate air gap, so that these results will be different from those obtained on ring specimens by the manufacturers. These tests are valuable in giving a direct comparison between different grades of ferrite under the same conditions and the results may be used directly to help in the design of solenoidally wound coils.

The test coil consisted of six turns of 20 s.w.g. copper wire on a coil former of $\frac{3}{8}$in outside diameter with an axial hole $\frac{1}{4}$in diameter for the core specimen. The impedance of this coil with different magnetic cores was measured on a Wayne Kerr v.h.f. admittance bridge, type B701 at various frequencies over the range 2 to 80Mc/s. The bridge measures admittance, so giving the parallel inductance and resistance of the coil. The inductive range of the bridge was extended by adding capacitors in parallel with the coil under test. At the low frequencies the capacitance needed was so large that the losses in the capacitors became appreciable and because of this the tests were discontinued at 2Mc/s.

As it was required to compare the properties of the various core materials, all the tests were carried out on the same coil throughout the frequency range. The cores consist of rods of ferrite material 2in long by $\frac{1}{4}$in diameter. The increase in inductance of the coil when a core is inserted is taken as the 'insertion factor' of the core material which is defined by

$$\text{Insertion factor} = \frac{\text{Inductance of coil with a core}}{\text{Inductance of coil without a core}}$$

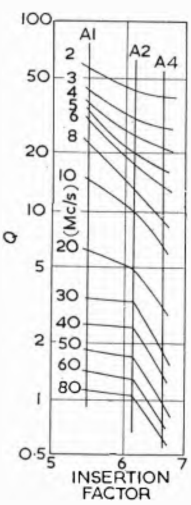

Fig. 1. Manganese-zinc ferrite

The Continental and North American equivalent names for the Mullard Ferroxcube ferrites are: A.1—IIB; A.2—IIC.2; A.4—IIIA; B.2—IVB; B.3—IVC; B.4—IVD; B.5—IVE

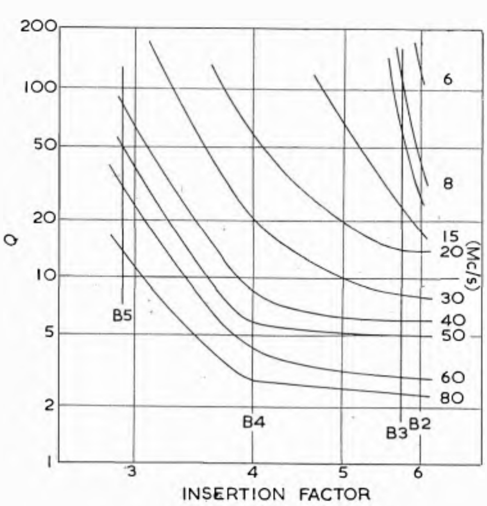

Fig. 2. Nickel-zinc ferrite

The measured Q of the coil with its ferrite core is taken to be the Q of the material of the core at that frequency.

The accuracy of measurement on the Wayne Kerr bridge is 2 per cent. There are some errors introduced by using the bridge on the limits of its range and by any loss in the

* *A.E.I. (Rugby) Ltd.*

extra capacitors used to extend the range of the bridge. The true value of the Q of the core material is estimated in the Appendix to be

$$1/Q = 1/Q_2 - 1/Q_1\mu$$

where Q_2 and Q_1 are the measured Q's of the ferrite cored coil and the air cored coil respectively, and μ is the insertion factor. Q_1 was measured to be about 200 so the error in assuming

$$Q = Q_2$$

is 10 per cent or more for a Q of 100, or more. All the errors tend to make the measured Q too small, so there is an estimated error of about 10 per cent on the high values of Q. The results of these tests are more accurate than 10 per cent except that there may be larger error in low loss materials especially at low frequencies. This error will always give the Q too small.

Ferroxcube material may vary ±20 per cent in magnetic properties between different samples. One sample of each grade of ferrite was measured to provide the results from which these design curves were derived, although another sample of each grade of nickel—zinc ferrite of slightly different dimensions was measured at the same time. There was an average variation of 20 per cent in Q values between the different samples of the same material but this variation appears small compared with the overall variation of Q through the frequency range.

There may be an appreciable difference between these measured values and the properties of any other particular sample of ferrite.

Acknowledgment

The author wishes to express his gratitude to the Directors of Associated Electrical Industries (Rugby) Ltd for permission to publish this article.

Appendix

An analysis of the measured properties of a magnetic cored coil.

For the air coil

Impedance $Z = R + j\omega L$

and $Q_1 = \omega L/R$

For the magnetic core inside a coil

Field $H = Ki$ (definition of K)

$\therefore \ B = \mu_1 Ki$

and back e.m.f. $= j\omega N\mu_1 Ki$, (where N is the number of turns on the coil).

If the magnetic material is lossy μ_1 will have real and imaginary parts.

Let $\mu_1 = \mu(1 - 1/Q)$

where Q is the qualify factor of the core material.

The measured impedance of the coil

$$Z = R + j\omega N\mu K + \frac{\omega N\mu K}{Q}$$

The measured Q of the coil

$$Q_2 = \frac{\omega N\mu K}{R + \omega N\mu K/Q}$$

But from a consideration of the air coil

$$\omega L = \omega NK$$

$$R = \omega NK/Q_1$$

$$Q_2 = \frac{\omega N\mu K}{\omega NK/Q_1 + \omega NK\mu/Q}$$

$$\therefore \quad 1/Q_2 = 1/Q_1\mu + 1/Q$$

$$\text{or} \quad 1/Q = 1/Q_2 - 1/Q_1\mu$$

μ may be taken as the measured increase in inductance due to the core material. This is the insertion factor.

For low loss specimens with measured $Q = 100$ and coil $Q = 200$ and an insertion factor of 5. The measured Q is too small by 10 per cent.

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The Use of Gas in Television Tube Manufacture

The electronic, or electrical engineer tends to regard gas as an outmoded form of lighting or heating. In fact it plays an increasingly important part in industrial processes and this is exemplified by its use in modern cathode-ray tube and valve factories.

The Simonstone Plant of Mullard Ltd, which is probably the most modern of its kind in Europe, produces over 1¼ million television cathode-ray tubes a year and employs some 1 700 people on continuous shift work.

The Blackburn Plant of Mullard Ltd, covers 46 acres of ground and employs nearly 5 000 people. It is responsible for most of the Groups production of radio and television valves, turning out about a million a week.

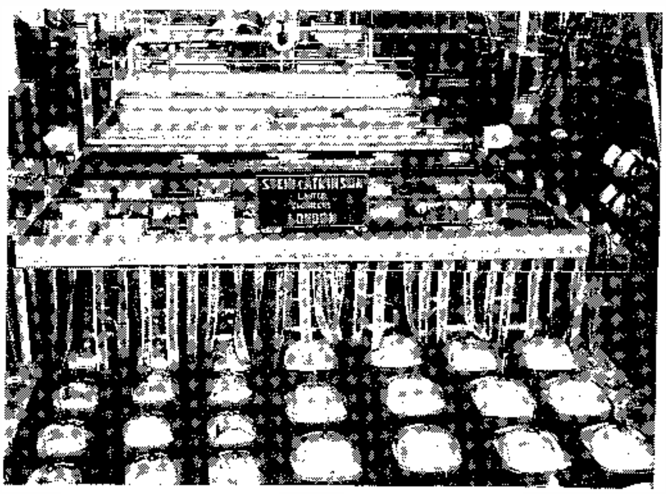
Between them the Mullard factories in Lancashire use about 1·8 million therms ($400 \times 10^6 ft^3$) of town gas a year which is 1·6 per cent of the total amount supplied by the North Western Gas Board for industrial use.

The units having the largest gas consumption in the Simonstone television tube factory are the two Lehr furnaces. Each of these is 100ft long and has a starting consumption of 6 000ft^3/h of gas and an average running load of 4 000ft^3/h. The tubes are passed through one of these two furnaces on a conveyor belt immediately before the gun is fitted to drive off all moisture and to bake out the unwanted constituents of the various coatings, which would otherwise make it impossible to maintain the necessary vacuum in the finished tube. This baking process also fixes the fluorescent screen firmly to the glass face.

But this is only one of the various uses of gas in this plant. It is also widely used in the assembly of glass component parts into the complete bulb; in the assembly of the electron guns and for sealing off the tubes from the vacuum pumps.

Gas is also used extensively at the Blackburn works in the glass making departments for furnaces, glass welding and shaping operations and for annealing; while in the fine wire factory it is used for furnaces for processing in natural ores and as a source of heat for swaging and wire drawing operations.

One of the two 100ft long Lehr furnaces



A System of Stereophonic Broadcasting

AMONG the systems of stereophonic broadcasting recently under consideration by the European Broadcasting Union was one developed by G. D. Browne of the Mullard Research Laboratories*.

This system for use in v.h.f., f.m. broadcasting is based on a time multiplexing technique similar to that used in some point-to-point communications systems and requires only one v.h.f. carrier wave to transmit the stereophonic signals.

The system's chief advantage, however, is that it would enable stereophonic receivers to be produced that are little more complex than conventional monophonic sets: apart from the second loudspeaker and audio stage necessary for stereophony, the only addition would be a circuit using at most two valves, or possibly one transistor and two crystal diodes. Existing v.h.f., f.m. radiograms with facilities for stereophonic record reproduction would, in many cases, be convertible to receiving stereo broadcasts on the system by the addition of only one valve and its associated circuit.

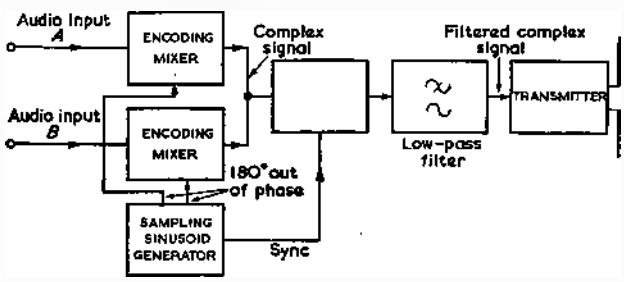


Fig. 1. General arrangement of the transmitter

The system meets all the generally accepted performance requirements for stereophonic broadcasting. It is compatible with the existing v.h.f. sound transmissions, so that conventional v.h.f. sets will accept the stereo transmissions and reproduce them monophonically in the normal way; it is also reversely compatible, which means that the listener with a stereophonic receiver designed for the system will hear the monophonic signal from both channels of his receiver when tuned to a monophonic transmission.

Modification to Existing F.M. Transmitters

The assumed requirement is to provide a broadcast service for normal two-signal stereophony, i.e. stereophony in which two signals (A and B) are derived, the sum of which is an acceptable representation of the equivalent monophonic sound, while their difference is related to the lateral positions of the sound sources involved.

The system is a twin-channel application of the principles of pulse amplitude time multiplexing under conditions in which the bandwidth, after multiplexing but before modulation of the transmitter, is restricted to a logical minimum. The system is essentially symmetrical in character as regards its treatment of the signals A and B.

The A and B stereophonic signals are fed into an 'encoder' whose output passes, in place of the normal monophonic signal, to the transmitter frequency-modulation equipment. The remainder of the transmitter is of standard form.

A sampling generator operating at the multiplexing frequency produces two sinusoids in anti-phase. These are fed into two mixing or multiplying devices to which are also applied respectively the two stereophonic signals A and B, suitably pre-emphasized. The sampling sinusoids are half-wave rectified by the circuits to produce two time interlaced pulse trains, one amplitude modulated by A, the other by B (see Fig. 1). The resultant output waveforms from the encoding mixers are shown in Figs. 2(a) and 2(b). Although for clarity, pulses with an angle of flow of about 90° are shown, calculation and performance indicate that half-wave rectified pulses (i.e. an angle of flow of 180°) produce the best results from the point of view of optimizing the system performance.

Adding the outputs from the encoding mixers gives the complex signal shown in Fig. 2(c). This signal alone does not contain any information to resolve the A, B ambiguity in a subsequent receiver. In order to provide for correct synchronization, a small amplitude component at the sampling or multiplexing frequency in phase quadrature

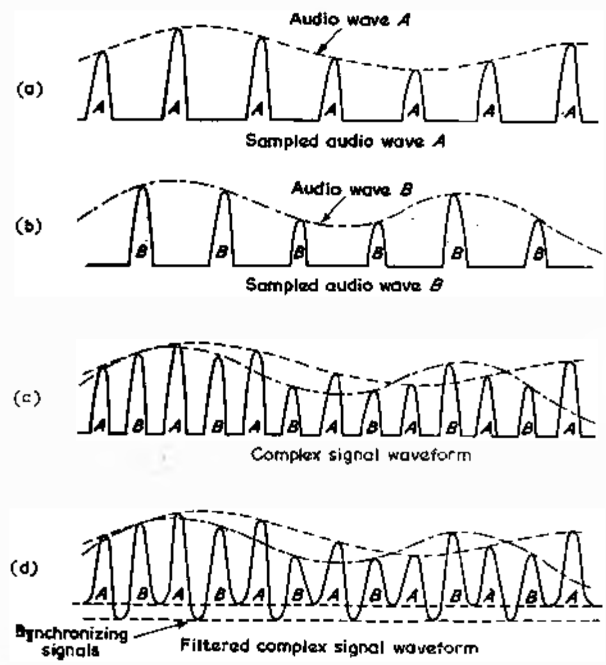


Fig. 2. The transmitter waveforms

with both sampling pulse trains is introduced, to give the asymmetry illustrated in Fig. 2(d). The rounded shape of the pulses is due to the bandwidth limitation effect produced by passing the complex signal through a low-pass filter (Fig. 1) just before entry into the frequency modulator of the transmitter. Thereafter a normal transmitter is used.

Modification to Existing F.M. Receivers

A normal receiver is used up to the output circuit of the frequency discriminator, which is maintained at adequate bandwidth to recover the complex signal shown in Fig. 2(d). The synchronizing separator circuit requires the use of either a transistor or a triode in the oscillator. In some existing normal receiver designs such a triode is already available for use in this way.

The circuit of a valve synchronizing separator and decoder, shown in Fig. 4, although uninvolved, does not represent the ultimate in simplicity which should finally be achieved.

The negative synchronizing pulses are separated in a synchronizing separation circuit (see Fig. 3) and are used

* *British Provisional Patent Application No. 8407/59.*

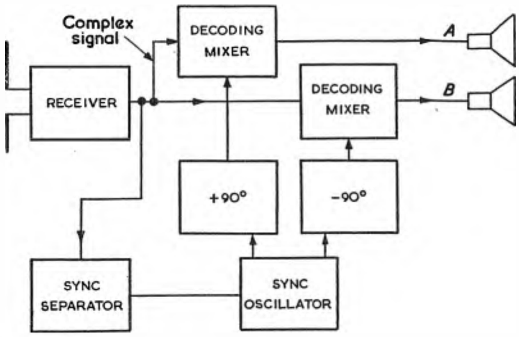


Fig. 3. The receiver arrangement

to phase lock an oscillator at the multiplexing frequency. The output from this oscillator is rephased by $\pm 90°$, to obtain the correct in-phase, anti-phase relationship for operating the decoding mixers. These are thereby synchronized with the encoding mixers in the transmitter, and thus reproduce the signals *A* and *B* respectively. After de-emphasis these signals are then directed into the output amplifiers *A* and *B*.

In the case where the stereo receiver is receiving a monophonic transmission, no synchronizing pulses are available, and the monophonic signal appears in both audio output circuits, irrespective of whether the synchronizing oscillator free runs or stops under these conditions. If the latter course is adopted an advantage in output signal-to-noise ratio is obtained.

Frequency Spectrum

In view of the bandwith limitations due to radio frequency channel allocations and to presently adopted techniques in typical monophonic receivers, it has been necessary to confirm that the proposed system can be operated within these restrictions.

The filtered complex wave fed to the frequency modulator of the transmitter has a frequency spectrum consisting of $A + B$ at audio frequency, $A - B$, d.s.b., a.m. on a suppressed subcarrier at the sampling frequency, and $A + B$, d.s.b., a.m. on a subcarrier at the second harmonic of the sampling frequency. Audio bandwidth is usually 15kc/s and the sampling frequency 32·5kc/s. The bandwidth of the spectrum shown therefore is 80kc/s.

Such a modulating spectrum is acceptable in f.m. transmitters without exceeding the radio frequency signal bandwidths normally employed, for example, in Band II. By retaining a complex signal bandwidth of 80kc/s, ease of receiver synchronization is achieved without either radiating a special high power synchronization signal from the transmitter, or complicating the receiver by the inclusion of high quality synchronizing filters.

Compatibility

The listener with a monophonic receiver will continue to hear an acceptable monophonic signal when receiving a stereophonic transmission. The level of sound at the receiver may change by a few decibels, when the transmitter is switched from mono to stereo operation, but this effect may be obviated, if desired, by suitably restricting the transmitter deviation on mono transmission.

Reverse Compatibility

The listener with a stereophonic receiver will hear a monophonic signal from both audio outputs when receiving a monophonic transmission. The level of sound at the receiver may change by a few decibels, as before, but this effect can also be avoided by similar means.

Crosstalk

The theoretical limit of crosstalk of the *A* input into the *B* output and vice versa (for half-wave rectified and decoding pulses) is −45dB. This is well below the maximum permissible for stereophonic broadcasting, and will also be acceptable for many bilingual and other two-signal transmissions.

Signal-to-Noise Ratio

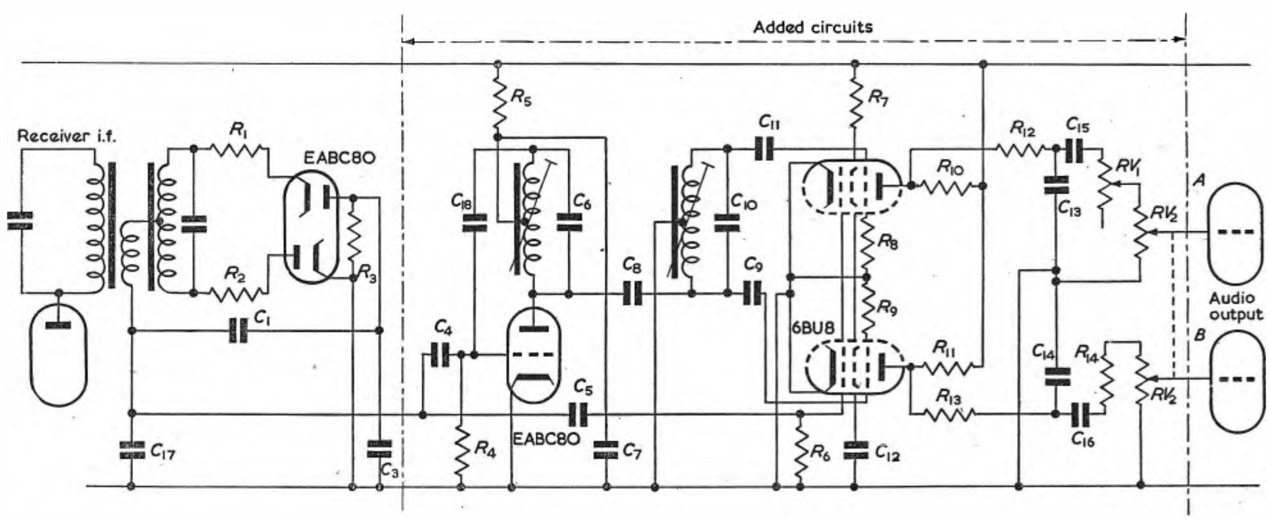
The results of calculations and, in some cases, experiments show that the following performance is achieved.

RECEIVER	MONOPHONIC		STEREOPHONIC	
TYPE OF TRANSMISSION	MONO	STEREO	MONO	STEREO
Change of output S/N	0*	−5dB	−15dB*†	−20dB
Change of receiver output level	0*	−5dB	0*	−5dB

* *This figure of −15dB will be improved to 0dB if the synchronizing oscillator in the receiver stops.*

† *These figures will be degraded by −5dB if the deviation on mono transmission is suitably restricted. Under these circumstances receiver output levels remain constant throughout.*

Fig. 4. The sync separator and decoder



A Television Focus Indicator Unit

By J. B. Potter*

A transistorized television focus indicator is described which allows an objective criterion to be applied to the electrical and optical focusing of any television picture signal source. The unit is suitable for bridging into 75Ω coaxial cable and for its operation relies on the fact that with correct electrical and optical focus of the image conversion equipment the high frequency content in the picture signal is a relative maximum.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

IN a television system it is necessary to obtain correct adjustment of optical and electrical focus of the optical-electrical transducing equipment if optimum quality of the transmitted picture is to be obtained. Under normal studio conditions, this adjustment is dependent on the subjective assessment of the picture displayed on the studio monitor and consistent results with different operating personnel may not be achieved, particularly if close control of the monitor focus is not obtainable.

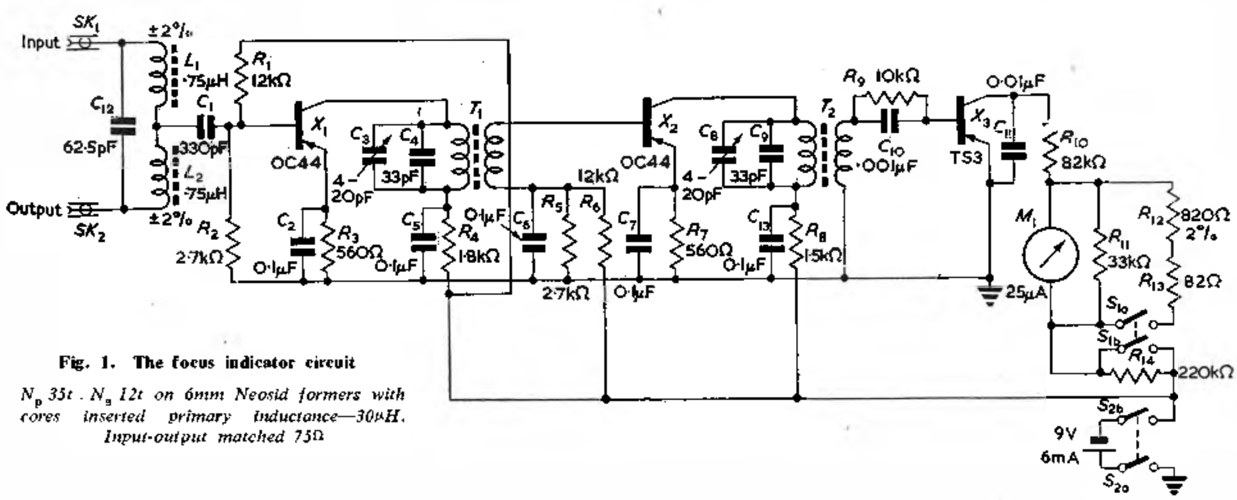


Fig. 1. The focus indicator circuit

N_p 35t . N_s 12t on 6mm Neosid formers with cores inserted primary inductance—30μH. Input-output matched 75Ω

Hence, it is desirable to provide some objective criterion for the adjustment of the focus of the image conversion equipment. The unit described here serves this purpose. It is suitable for bridging into 75Ω coaxial cable at any one point in the television system and indication is obtained visually by means of a meter.

Principle of Operation

The television focus indicator relies for its operation on the fact that with correct electrical and optical focus of the image conversion equipment, the high frequency content in the picture signal is a relative maximum; i.e. sharp edges in the picture correspond to the high frequency components in the signal. Hence as the picture is focused the amplitude of the high frequency components is varied and a maximum is reached at the true focus point. The focus indicator measures the amplitude of one such band of high frequency components, the output device indicating this amplitude and thus allowing the true focus point to be determined.

* *Commonwealth of Australia Postmaster-General's Department.*

Use of Focus Indicator

From the above it will be seen that the indicator can, in principle, be used in conjunction with any picture signal source. However, some care is necessary if the unit is used to control the optical focus of the studio cameras, when normal studio pictures are being shot. In such a case, it is possible for more detail to appear in the background than in the object of interest which may result in only the background being focused. Under normal conditions it is unlikely that the indicator would be used for 'live' material however, its main purpose, as far as cameras are concerned, is to assist in obtaining a correct electrical focus adjustment and in optical focus control for test and maintenance purposes where test charts are used. No such difficulties occur with any of the other picture signal sources and hence full use can be made of the indicator in all other cases.

Circuit Description

The circuit diagram is shown in Fig. 1. From this it will be seen that the focus indicator consists essentially of two r.f. stages in cascade followed by a detector and d.c. amplifier stage.

The input capacitance of the first stage is incorporated in a low-pass filter section to prevent capacitive reflections when the unit is bridged into 75Ω coaxial cable. This filter section, which appears in the signal path when the unit is bridged into a cable, has a cut-off frequency of 10Mc/s and, as it is of the M derived type (M = 0·5), very little frequency distortion occurs at normal operating frequencies due to the presence of this circuit.

The input signal is *RC* coupled to the base of the first transistor X_1 which is an OC44 type transistor. D.C. bias to the base is supplied via R_1 and R_2 from the -9V supply and results in a base potential of approx. 1·5V. The transistor emitter circuit contains a resistor to ground, bypassed by a 0·1μF capacitor, which automatically adjusts the operating point of the transistor to give the correct base current bias for 3mA emitter current. The collector circuit contains the interstage coupling transformer T_1 the primary being tuned to approximately 3Mc/s. Decoupling is provided by R_4 and C_5 so that the collector-emitter voltage is approximately 2V. The transformer provides an impedance transformation of 9 to 1, and matches the output impedance of X_1 to the input impedance of the OC44 transistor used for X_2. Similar arrangements are made for the biasing and decoupling of this stage and the interstage coupling transformer T_2 is identical to T_1 and tuned to the same frequency.

The secondary of T_2 is coupled into the base circuit of X_3, a type TS3 transistor being used as detector and d.c. amplifier. Here the 3Mc/s signal is detected in the base-emitter circuit and the resulting base current is amplified to produce a d.c. current in the collector circuit proportional to the amplitude of the input voltage. This current is filtered by an *RC* filter to remove residual 3Mc/s components and the d.c. output is indicated on a meter. Self bias of this transistor is used and R_9 should be adjusted to give maximum sensitivity of the detector circuit for minimum input signal.

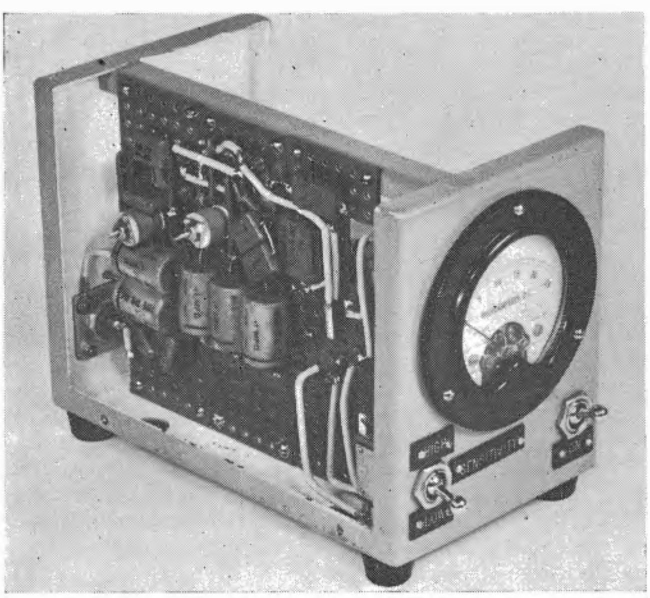
Two current ranges are used to cover the amplitude range of the 3Mc/s component found in normal television pictures. These two ranges are achieved by varying the shunt of the meter and correspond to:—

(1) High sensitivity range, 25μA full scale deflexion.

(2) Low sensitivity range, 100μA full scale deflexion.

It will be noticed that the resistor R_{11} is permanently shunted across the meter, this being necessary to prevent excessive swinging of the meter with adjustment of the focus controls. The meter is automatically protected from damage by means of R_{10} and R_{14}. These resistors ensure that the maximum current which can flow in the meter

Fig. 2. View of indicator unit showing construction details



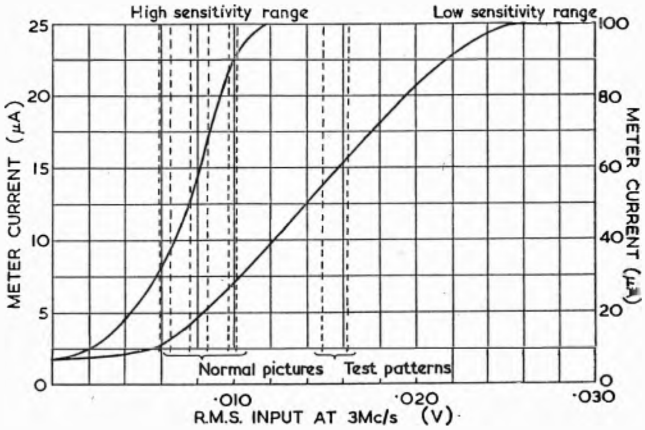


Fig. 3. Television focus indicator calibration curves

Vertical dotted lines are values obtained for slides used in test and show expected operating range. Curve may be displaced vertically due to temperature effects on VT_3 characteristics

circuit is equal to the full scale deflexion of the meter for that range. Such current limiting is achieved due to the constant current nature of the transistor. This ensures that no current limiting occurs until full scale deflexion of the meter is reached.

Because of the biasing arrangements used for X_1 and X_2 the stability of these two stages with temperature change is good. Such stabilization cannot be applied to X_3 but the main effect of temperature on this stage is to change the standing current for zero signal input which is not important. Since this unit has been designed for the Australian 625 line 25 frame system, the frequency of operation and the amplitude ranges to be handled may need adjustment for use with other television systems.

The Physical construction of the unit can be seen from Fig. 2. The size of the unit can be judged from the size of the 2¼in meter which is incorporated in it.

Results

The focus indicator unit was tested using studio slide and film scanning equipment. Twenty slides of varying degrees of detail ranging from test charts to very low contrast snow scenes were taken as representing the normal range of television pictures. Over the range represented by these pictures it was found that a satisfactory indication could be obtained at all times, sufficient indication being given on the low sensitivity range for 50 per cent of the slides used.

A motion picture film was also used for test purposes, the scenes of which again varied widely in contrast and detail. At all times in this case it was possible to obtain a satisfactory reading on the low sensitivity range of the meter.

The calibration curves for the two meter ranges are shown in Fig. 3 superimposed on which are the readings obtained with some of the slides discussed above. By means of these two curves it is possible to ascertain the magnitude of the 3Mc/s component in the pictures used. It will be seen that this varies from 0·006V r.m.s. to approximately 0·016V r.m.s. at 3Mc/s. In normal operation, however, this information is not required and the device is used qualitatively only.

Acknowledgment

The author is indebted to the Engineer in Chief of the Postmaster General's Department, Australia, for permission to publish the information contained in this article.

THE ANODE-CATHODE FOLLOWER

By C. H. Vincent*, M.Sc., Ph.D., F.Inst.P., A.M.I.E.E., and D. Kaine*, B.Sc.

A simple new circuit is described, referred to as an anode-cathode follower, which inverts negative pulses and enables the resulting positive pulses to be fed into a low impedance load with a gain very close to unity. A unit incorporating this circuit is described. It is designed to transmit pulses of up to 50V amplitude into a 100Ω load and includes modifications which enable either positive or negative inputs and positive or negative outputs to be selected by a switch.

(Voir page 261 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 266)

A FREQUENT source of difficulty in electronic work with fast pulses is the distortion and attenuation of the pulses due to the loading of the output of a unit by the capacitance of a cable connecting it to another unit. This problem is liable to be particularly acute in nuclear physics, where long runs of cable may be needed to carry pulses from their source in an experimental area to a remote control-room. Output circuits designed to cope with normal lengths of cable connecting between chassis in adjacent racks may well give difficulty under such conditions. This capacitive loading problem may be overcome by using a cable with a matched termination, so that there is no longer any capacitive loading effect, and the maximum usable length is then limited only by considerations of dispersion and attenuation in the cable itself. However, there still remains the difficulty of providing a suitable source of pulses to work into the low impedance of the cable and, in particular, to do this with an accurately constant gain, preferably of unity.

An ordinary cathode-follower provides a low enough output impedance to work into a cable characteristic impedance, but its gain then falls under this load to a value of the order of 0·5 or so, in typical cases. Moreover, this value is dependent on the transconductance of the valve, and may therefore vary appreciably with changes in the latter. A double cathode-follower[1] may be used for transmitting pulses into a low impedance with a gain approaching unity, but this circuit is not suitable for transmitting large positive pulses, as this would require a correspondingly large standing current in the lower valve of the two in the chain. Typical amplitudes of the pulses to be transmitted to a pulse height analyser, for example, are of the order of some tens of volts, so that the peak currents required are of the order of some hundreds of milliamperes. One possible solution is to attenuate the pulses before transmitting them and to amplify them again at the other end, but this is not a satisfactory solution, since it is less convenient and requires an extra amplifier and, in particular, it makes the whole system much more susceptible to interference pick-up.

The present article describes a new circuit, referred to as an anode-cathode-follower, which has been incorporated into a suitable unit for transmitting pulses of up to 50V amplitude along a 100Ω cable with a gain which is nearly independent of valve characteristic changes and which can be set very close to unity. The basic circuit accepts negative pulses and gives out positive pulses of the same amplitude from a low impedance output. It has also been adapted in the unit mentioned to give out negative pulses (from a high impedance or constant current outlet) into a 100Ω load impedance. An optional anode-follower invertor for the input pulses is also incorporated into the unit, so that pulses of either sign may be both accepted and transmitted, according to the switch combination selected by the operator.

* *United Kingdom Atomic Energy Authority, A.W.R.E., Aldermaston.*

Circuit Description

The basic circuit used is shown in Fig. 1. It is somewhat similar to an anode-follower (V_1) followed by a cathode-follower (V_2). (It is for this reason, of course, that the name 'anode-cathode follower' is suggested.) However, the feedback is taken, not from the anode of V_1, but from the cathode of V_2. In this way, the cathode of V_2 is made to follow a voltage which is the negative of that appearing at the input

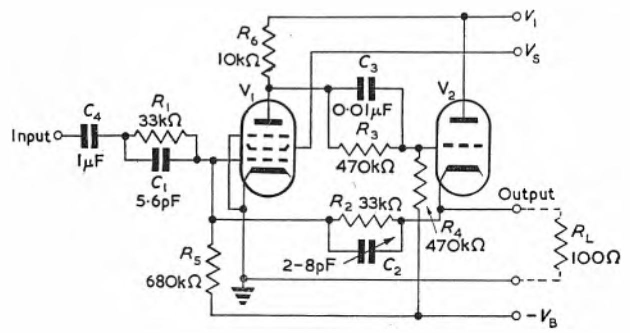


Fig. 1. Basic circuit of the anode-cathode follower with typical component values

terminal, and the varying voltage difference between the grid and cathode of the cathode-follower V_2, due to the low load impedance, is compensated for. Apart from the feedback connexion from its cathode, V_2 in itself is simply a normal cathode-follower. An auxiliary feature of the circuit shown is that the feedback loop is completed for direct current by the resistor R_3, in parallel with the coupling capacitor C_3, and stabilizes the mean currents in the valves with the aid of the resistors R_4 and R_5 to the negative bias supply line. The capacitors C_1 and C_2 are for correction of the high-frequency response of the feedback circuit.

The gain of this circuit with the load R_L may be calculated by normal methods, and is given by

$$G=(R_2/R_1)[1+(1/R_6'g_1)\ \ (1+(1/R_Lg_2)+(1/R_2g_2)+(1/\mu_2))\ (1+(R_2/R_1)+(R_2/R_5))]^{-1} \quad \ldots\ldots\ldots (1)$$

where $1/R_6' = 1/R_6 + 1/R_4$ and the output impedance of the pentode V_1 is taken as infinite, and also where g_1 is the transconductance of V_1 and g_2 and μ_2 are the transconductance and amplification factor, respectively, of V_2. The only approximation made in the derivation of this formula is the assumption that the loading of V_2 by R_2 and the attached input circuit may be represented by a resistance R_2 only. This is a very close approximation. The factor $R_6'g_1$ is the gain of V_1 as a simple amplifier, and this can be made large, so that the second term in the square brackets is correspondingly small. (The remaining factors

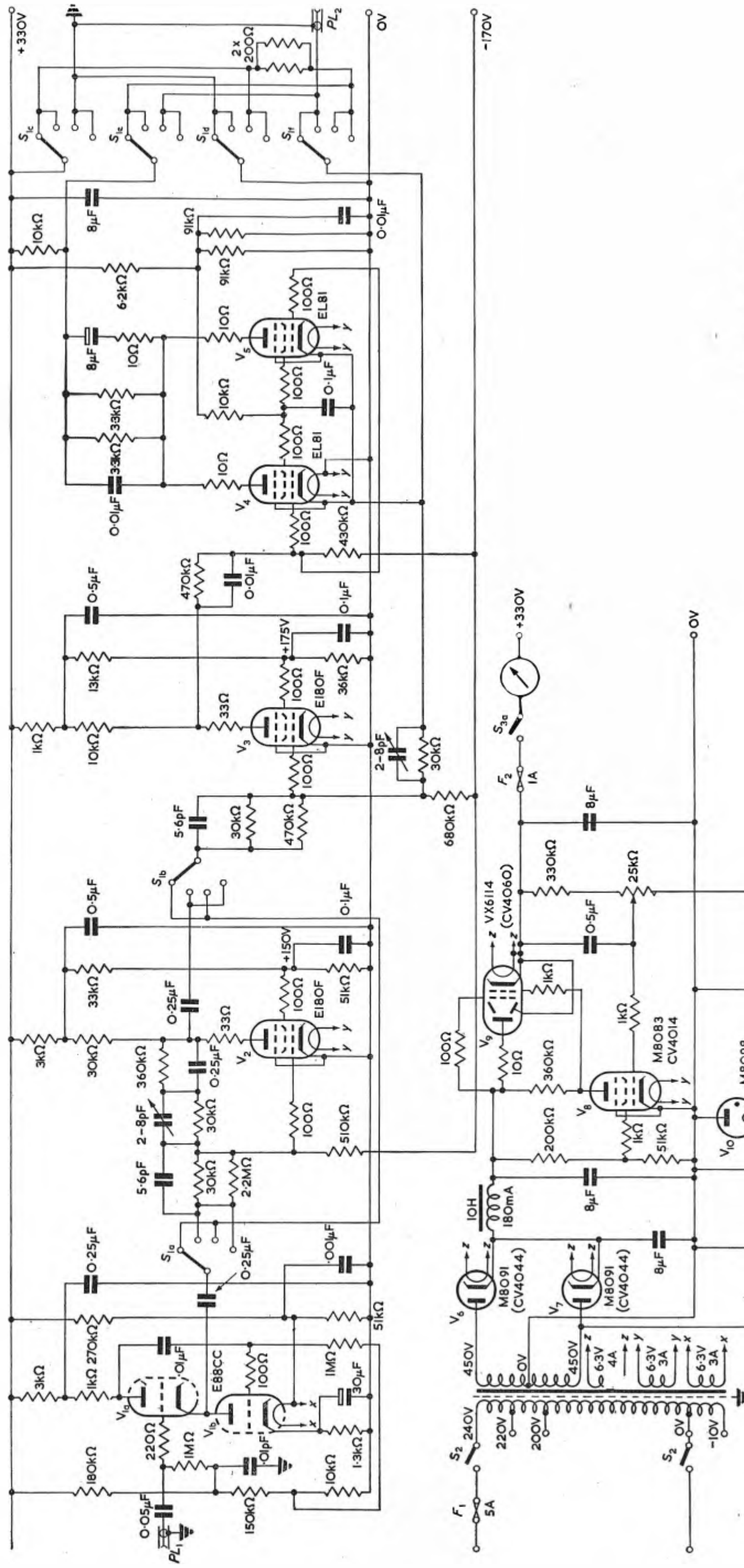


Fig. 2. Pulse transmitter

in that term may be of the order of 2, in each case, but need not be larger.) The dependance of the gain on g_1, g_2 and μ_2, that is, on the valve characteristics, is therefore small. The ratio of R_2 to R_1 may be made slightly greater than unity, to make the overall gain as close to unity as possible. The output impedance of the circuit may be deduced from the above formula and, neglecting small terms in $1/R_6'g_1$, it is

$$\frac{1 + R_2/R_1 + R_2/R_5}{R_6'g_1g_2}$$

Since $R_2 \sim R_1$, and since R_5 may be made relatively large by using a sufficiently high bias supply voltage, this impedance may be further simplified to, approximately,

$$\frac{2}{R_6'g_1g_2}$$

This is the output impedance of the cathode-follower V_2, divided by the approximate loop gain ($R_6'g_1/2$).

The conditions for stability are very similar to those for an anode-follower, as the cathode-follower V_2 makes little effect by its insertion. The loop appears to be very stable under all conditions that are likely to be encountered in practice and no signs of oscillation have been observed.

The Pulse Transmitter Unit

The circuit diagram of the whole unit is shown in Fig. 2. The valve V_1 forms a double cathode-follower, which presents a high impedance to the input terminal and feeds the signal into the relatively low impedance input to V_2 or V_3. The valve V_2 is the anode-follower, which is provided to enable positive input pulses to be accepted and inverted before they are passed on to V_3. The circuit of V_2 includes small capacitors for correcting the high-frequency response and a direct-current feedback connexion for stabilizing the anode current. Sections a and b of switch S_1 make the necessary connexions to include V_2 in the circuit or by-pass it as required. The small loss of signal which would otherwise occur in the circuit of V_2 is compensated for as closely as possible by a suitable high value resistor (2·2MΩ is shown) connected in parallel with the appropriate 30kΩ resistor in the feedback network at the grid. This resistor modifies the feedback ratio to make the overall gain of V_2 as close to unity as possible.

The valve V_3 and the paralleled valves V_4 and V_5 form the first and second stages, respectively, of the output circuit already described. This circuit is also provided with capacitors for high-frequency response correction and with direct-current feedback. The latter serves to stabilize the currents in the valves, particularly V_4 and V_5. It is desirable that the mean current in these valves should not change excessively due to valve characteristic changes or non-linear effects when the circuit is passing large pulses at a high mark-to-space ratio. The unit is primarily intended for use at reasonably low mark-to-space ratios, but is nevertheless capable of dealing with pulses of full amplitude up to a mark-to-space ratio of about 1 to 10. The small loss of signal which occurs in the double cathode-follower V_1, together with the similar loss which would otherwise occur in the circuit of V_3, V_4 and V_5, are compensated for by a high value resistor (470kΩ is shown) in parallel with the appropriate 30kΩ feedback network resistor, as in the circuit of V_2.

When the circuit is giving out positive pulses, the output terminal is connected to the cathodes of V_4 and V_5, so that the 100Ω load is connected to these cathodes, as described in the previous section. Two parallel resistors forming another (internal) load of 100Ω are connected into the anode circuit of V_4 and V_5, for reasons which will become apparent. This resistance does not have any major effect on the operation of the cathode-follower as such. If negative output pulses are required, the switch S_1 is set to a position which connects the output terminal to the anode circuit of V_4 and V_5 and the 100Ω internal resistor to the cathodes of these valves. Sections c, d, e and f of S_1 are involved in these connexions. If the external load is 100Ω, the same as the resistance of the internal load, conditions are unaltered, except that negative pulses appear at the output instead of positive. The power supply used is a floating one, and provision is made to earth one side of the connexion to the external load, in either case. Screen-grid current is returned to the cathodes via the screen by-pass capacitor, so that this current does not affect the equality of the positive and negative pulses obtainable. On the other hand, as it is clear that the negative pulses are obtained from a high impedance or current-controlled source, their voltage amplitude will be proportional to the external load resistance, if the latter departs from 100Ω. The 10kΩ screen dropping resistor is an extra load on the cathodes, which is therefore compensated for by a similar resistor from the anodes to the positive h.t. line.

It would be possible to use triodes for V_4 and V_5 instead of pentodes. Neither the change from triodes to pentodes, nor the addition of the extra 100Ω load at the anodes, substantially affects the operation of the circuit, although they do modify the formulae for the gain and the output impedance in minor detail. The resistance capacitance network shown immediately above the 10Ω parasitic suppression resistors in the anode circuit of V_4 and V_5 is simply for the purpose of reducing the anode voltage to these valves and thus keeping their power dissipation within permissible limits. It is to avoid the necessity of providing a separate lower voltage supply to this stage.

Conclusion

The anode-cathode-follower appears to be a simple and useful circuit for feeding positive pulses of moderate voltage into a low-impedance load with a (negative) gain very close to unity. The units described can provide both positive and negative pulse outputs, and have been found to be of considerable use in a nuclear physics research laboratory.

REFERENCE

1. Hammack, C. M. M.I.T. Radiation Laboratory Report, 469 (1943).

A Nuclear Reactor Trainer

A special purpose computer capable of setting up an electronic analogue of a complete power station is to be used by Calder Hall as a trainer for reactor operators. This 200-amplifier instrument is being built by Short Brothers & Harland Ltd.

The computor will be installed at the Calder Operations School and will be used for simulation of large nuclear power stations. It will enable trainee operators to gain experience of new stations now being built for the Central Electricity Board and the South of Scotland Electricity Board.

The new Short computer is of particular value for training purposes in that it can simulate unstable conditions in the reactor core and give operators experience of quickly bringing them under control. A representation of the steam raising and electrical generation sections of the power station will also be included in order to demonstrate their various effects on the reactor.

A mock reactor control desk is being designed by the Research and Development Branch of the Authority and for this the simulator will provide the following information: average flux, coolant flow, the core inlet gas temperature and the three zone outlet gas temperatures, h.p. and l.p. steam pressures and flows, control rod positions, turbine speeds, and power output.

Work has already begun on the simulator, which has been designed by the Special Products Department of Short's Precision Engineering Division at Castlereagh.

The Tunnel Diode

The negative resistance or 'tunnel' diode originally developed by Dr. Leo Esaki of the Sony Corporation of Japan, is to be manufactured in the United Kingdom by the Transistor Division of Standard Telephones and Cables Ltd.

The samples offered are suitable for operation at a maximum frequency of about 5Mc/s but it is expected to improve on this shortly.

Two types of tunnel diode designated JK10A and JK11A are available with characteristics as follows:

Voltage (V_1) at current maximum (I_1) 50mV

Voltage (V_2) at current minimum (I_2) 200mV

Maximum current (I_1) JK10A 3-10mA

JK11A 10-30mA

Current ratio $I_1/I_2 = 2\text{-}5$

Rating—100mA maximum

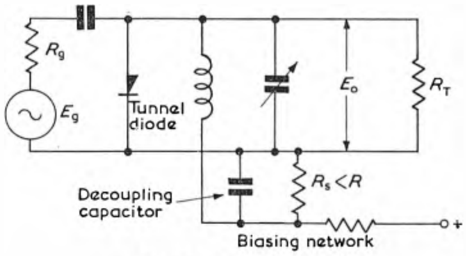


Fig. 1. Tuned amplifier

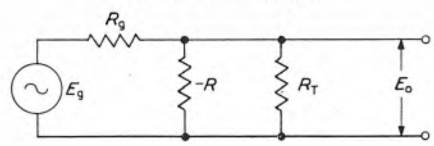


Fig. 2. Equivalent circuit

Typical Circuit Applications of the Tunnel Diode

Tuned Amplifier Circuit (Figs. 1 and 2)

If the tunnel diode is biased to operate at a point where its incremental slope resistance is $-R$, at resonance the output voltage E_o is given by:

$$E_o = \frac{-E_g R\, R_T}{R_g\, R_T - R(R_g + R_T)} \quad \ldots\ldots\ldots\ldots \quad (1)$$

where R_T is the shunt loss of the tuned circuit combined with the load resistance.

For stability the denominator of equation (1) must be negative

$$\text{i.e. } R(R_g + R_T) > R_g\, R_T$$

$$\text{whence } R > (R_g R_T)/(R_g + R_T)$$

For voltage amplification to occur E_o must be greater than E_g, and this means that the numerator of equation (1) must be greater than the denominator:

$$R\, R_T > R\,(R_g + R_T) - R_g\, R_T$$

$$\therefore \quad R_g\,(R - R_T) < 0 \qquad \therefore \quad R < R_T$$

Hence R must be between limits:

$$R_T > R > (R_g R_T)/(R_g + R_T)$$

to ensure stable amplification.

Oscillator Circuit

A suitable circuit is shown in Fig. 3.

The condition for oscillation is:

$$R_S < L/C\, R_D$$

where R_D is the modulus of the slope resistance of the tunnel diode and R_S represents the total resistance in series with the diode including coil resistance, bias supply network resistance and the series resistance of the diode itself.

The frequency of oscillation, f, is approximately:

$$f = 1/2\pi\sqrt{[L(C + C_D)]}$$

where C_D is the junction capacitance of the diode.

Bistable Circuit

When operated as a switch, the tunnel diode will be biased to either of two stable operating points, indicated by A and B in Fig. 4 and in going from A to B and back to A, the current and voltage will follow the path indicated by the arrows.

In terms of voltage, it is possible to make the point A nearer the negative resistance region than the point B. Hence the diode will trigger from A to B on a positive voltage pulse but will not trigger from B to A on a negative pulse of the same amplitude.

Fig. 5 shows a bistable, or divide-by-two, circuit. The resistors, R_S, are chosen to be much less than R_D and the voltage developed across each is about 100mV, which is

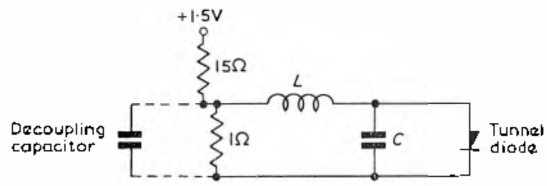


Fig. 3 (above). Oscillator circuit

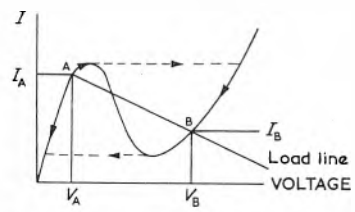
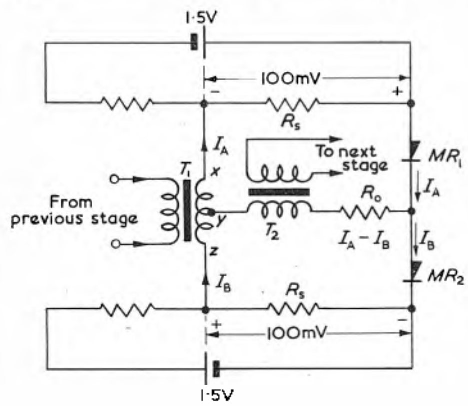


Fig. 4 (right). Voltage/current locus

Fig. 5 (below). Binary circuit



the correct amount to bias the diode to the middle of the negative resistance region. Neglecting, for the moment, the transformers, it is desired to bias the tunnel diode so that MR_1 is operating at point A on Fig. 4 and MR_2 at point B. Then it may be shown that R_o, the resistor in the centre arm, is given by (referring to Fig. 4):

$$R_o = \frac{V_B - V_A}{2\,(I_A - I_B)} - R_S/2$$

R_o is drawn as a load line on Fig. 4.

A positive pulse appearing across the terminals $X - Y$ of transformer T_1 will switch MR_1 from the bias point A towards the bias point B. The addition of an inductance in series with R_D will aid the switching action and cause MR_2 to switch from B to A because the back e.m.f. due to the fall in current through MR_1 is in the right sense to bring MR_2 into the negative resistance region.

The inductance in the centre arm is provided by the leakage inductance of transformer T_2, and the secondary winding of T_2 is connected to trigger the next stage. A second positive input pulse across terminals $Y - Z$ will cause MR_2 to switch from A back to B. Hence the circuit divides by two.

Short News Items

Six technical premiums of 25 guineas each have been awarded by The Radio Industry Council and the Electronic Engineering Association for the following articles published during 1959.

A. E. Karbowiak "Waveguide as a Long Distance Communication Medium" *Electronic Engineering*, September 1959.

T. G. Thorne and J. A. Billings, "The Performance of Doppler Navigation Systems" *British Communications and Electronics*, March 1959.

C. M. Cade "Infra-red Navigation Aids—Some Approaches to Increased Safety in Band Visibility" *British Communications and Electronics*, August/September 1959.

R. Rowland "Printed Circuits Applied to Broad-Band Microwave Links" *Electronic Engineering*, May 1959.

D. A. Wright "Compound Semiconductors" *Electronic Engineering*, November 1959.

R. J. D. Reeves "The Recording and Collocation of Waveforms" *Electronic Engineering*, March/April 1959.

A total of 63 articles published during 1959 were submitted. Mr. Norman Collins will be the chief speaker when the awards are presented at a luncheon at the Cafe Royal on Wednesday, 6 April.

The U.K.A.E.A. Development and Engineering Group with its main establishment at Risley is responsible for the development of new reactor systems and nuclear processing plant. Within the Development and Engineering Group is the Research and Development Branch whose main effort is concentrated on experimental work at Capenhurst, Culcheth, Springfields and Windscale. A Central Technical Services stationed at Risley is part of the Branch and is divided into five sections; two are concerned with neutronic problems and two deal with mechanical and chemical aspects. The fifth section is the Computer Section which has recently been equipped with a Ferranti 'Mercury' and an IBM 704 digital computer together with an Electronics Associated Inc. PACE analogue computer. With these facilities this section is now the largest general purpose computing centre in Europe. An automatic data-transmission link between Risley and the Atomic Energy Research Establishment is in operation so that the IBM 704 can be used more effectively by the Harwell staff. A similar link will be established shortly between Risley and Winfrith Heath. The data-transmission link enables punch card information to pass between these groups extremely rapidly, and special use is made of the link to modify machine programmes and to transmit results.

British Standard G.100, General Requirements for Aircraft Electrical Equipment, is being revised to take account of the requirements of modern high-performance aircraft operation, and in the light of experience of working to the provisions of the first edition, which was published in 1952. The British Standards Institution's technical committee is anxious to obtain from industry as many comments as possible on the drafts of the proposed new standard. Copies of the first two parts of the draft standard can be obtained, free of charge to subscribing members, from the British Standards Institution, 2 Park Street, London W.1, or at 3s. 6d. for Part 1 and 5s. for Part 2 to non-members.

A Second Electronic Computer Exhibition has been announced by the Joint Committee of the Electronic Engineering Association (E.E.A.) and the Office Appliance and Business Equipment Trades Association (O.A.B.E.T.A.) and is to be held in the National Hall, Olympia, London, from Wednesday, 4 October, to Thursday, 12 October 1961. A second Business Computer Symposium will be held concurrently. In the Grand Hall, Olympia, adjoining the Computer Exhibition, will be the Business Efficiency Exhibition, from 2 to 11 October 1961, in which computers and ancillary equipment will be shown. The first Computer Exhibition was held in 1958, but since then much new equipment has been announced. There has been a considerable increase in production and sale of British computers to government departments, local government authorities, industry, commerce, university and research establishments at home. Export sales have become more pronounced, and their value since the last exhibition was held is estimated at £1 750 000. The secretary of the Joint Committee is D. C. Scoones of Messrs. Peat, Marwick & Mitchell & Co., 94-98 Petty France, London S.W.1, and the Exhibition organizer is Mrs. S. S. Elliott, M.B.E., 64 Cannon Street, London E.C.4.

John Bass Ltd, an associate company of The British Manufactured Bearings Co. Ltd of Crawley, Sussex, has introduced a range of clean dry air cabinets. The design of these cabinets is based on the experience gained by the parent company in the course of manufacture of miniaturized ball bearings. Included in the associate company's products is the Bassaire 'Clean Room', a self-contained unit providing filtration down to ½ micron and constant temperatures controlled to an effective differential of 2°F. To eliminate contamination introduced by operators entering such clean air areas a 'Man Cleaner' cleansing chamber has been designed. The chamber, slightly larger than the conventional telephone kiosk, has a door at both ends, enabling the operator to enter at one end for cleaning and to emerge at the other into the clean area.

The tenth international symposium organized by the Polytechnic Institute of Brooklyn's Microwave Research Institute will be held on 19-21 April 1960 in the Auditorium of the Engineering Societies Building (33 W.39 Street in N.Y.C.). This Symposium on Active Networks and Feedback Systems is sponsored by the Professional Group on Circuit Theory of the Institute of Radio Engineers with the support of the Air Force Office of Scientific Research, the Office of Naval Research and the U.S. Army Signal Research and Development Laboratory. Abstracts of all papers, a detailed programme and registration form are available from The Microwave Research Institute, Polytechnic Institute of Brooklyn, 55 Johnson Street, Brooklyn 1, New York.

Pye Ltd of Cambridge are manufacturing forty-five 1kW tropospheric amplifiers for installation at the Supreme Headquarters Allied Powers in Europe. These amplifiers are for use in a NATO telecommunications network covering Turkey, Greece, Italy, France, Germany, Britain and Norway.

Eleven inland waterway tankers owned by Esso Tankschiff Reederei, GmbH, Hamburg, are to be fitted with the latest Decca River Radar 215 which was introduced at the end of 1959. In Belgium and the Netherlands, as well as in Germany, inland waterway tankers owned by the Esso group have been using the original Decca River Radar for some years to enable navigation to be maintained both at night and in low visibility. The Decca River Radar 215 is a very high definition radar equipment specially designed for use in narrow congested waters. Some 300 Decca River Radars are in use in the inland waterways of Europe and the U.S.A.

Drayton-Southern Limited is in course of joint formation by Southern Instruments Limited of Camberley and The Drayton Regulator and Instrument Company Limited for the purpose of expanding the instrumentation and control interests of both companies, whose research, production and sales resources will be available to it. These resources include The Drayton Regulator and Instrument Company's technical sales network and machine shops, and Southern Instrument's research staff and design and production facilities backing their established business in electronic equipment in the data handling, precision measuring, analytical and communications fields. Drayton-Southern Limited, whose authorized capital will be £100 000, will have the benefit of licensing agreements with American companies and will initially offer integrated monitoring and warning systems for the industrial measurement and control of temperature, pressure, strain, vibration, acceleration, load and displacement, based on well tried and widely used principles. The company will also offer instrumentation derived from specialized knowledge in such fields as textiles, chemical manufacture, paper-making, shipping and electricity generation.

A new Division has been formed at Harwell, to be known as the Research Reactors Division. It will have responsibility for the operation, maintenance, safety and development of Harwell research reactors and their facilities, including rigs, loops and auxiliary equipment. In addition the Division will undertake physics experiments necessary for the solution of design and operational problems and studies on possible future research reactor systems. The Division has been formed from the Research Reactor Group of the former Reactor Division, now mostly transferred to the Atomic Energy Establishment, Winfrith, and the Reactor Services Group, previously part of the Engineering Division. The Research Reactor Division will be responsible for the reactors, DIDO, PLUTO, LIDO, BEPO, GLEEP and for operation only of DIMPLE.

A new 186-mile over-the-horizon radio communications link between the United States and the Bahamas has now been put into operation. This link will provide additional telephone and other communication facilities between Nassau and Florida City, Florida. Equipment for the Nassau terminal was engineered and manufactured by Standard Telephones and Cables Limited, London, and works with a similar terminal at Florida City supplied by Kellogg Switchboard and Supply Company of Chicago, Illinois, and ITT Laboratories, Nutley, New Jersey, both STC associates and U.S. divisions of International Telephone and Telegraph Corporation. The Nassau terminal uses two 10kW transmitters, operating in the 2 000Mc/s band, feeding into two 30ft diameter parabolic reflector antennae. Two dual-diversity receivers, utilizing the same antennae as the transmitters, will be used to provide quadruple diversity reception. STC has also supplied and installed a 7 400Mc/s line-of-sight short-haul microwave link to transmit the entire baseband signal between the terminal and the Telecommunications Centre in Nassau City.

A new type of echo-sounder is being fitted to vessels of the Shell tanker fleet. It will measure shallow clearances of water beneath the keel of a ship to within two feet. Kelvin & Hughes Ltd, in conjunction with Shell Tankers Ltd, have developed a model which is being fitted to new 18 000 d.w.t. tankers and the Marconi Marine Company Ltd have designed a comparable unit, one of which has been fitted to a 32 000 d.w.t. tanker. The new type of echo-sounder incorporates a pierced-hull projector with the transmitting and receiving units housed in one small tank. This allows the transmitting wave and the returning echo to travel almost vertically, and thus, for all practical purposes, no parallax error is experienced. To facilitate the reading of depths of less than 30ft, an electronic magnifier has been developed using a cathode-ray tube about 6in in diameter. This unit is mounted in the wheelhouse, and is additional to the paper record in the chartroom.

The Annual Dinner of The Institution of Electrical Engineers was held at Grosvenor House on Thursday, 25 February 1960, when the President, Sir Willis Jackson, D.Sc., F.R.S., was in the chair. The principal guest was Sir George Thomson, M.A., LL.D., Sc.D., F.R.S., Nobel Laureate (Faraday Medallist 1960; President, The Institute of Physics), who proposed the toast of The Institution.

The British Radio Valve Manufacturers' Association (B.V.A.) and the Electronic Valve and Semiconductor Manufacturers' Association (V.A.S.C.A.) have now moved to new offices at Mappin House, 156-162 Oxford Street, London W.1 (telephone: Langham 8562).

The fifth Amateur Television Convention, arranged by the British Amateur Television Club, will be held on Saturday, 10 September 1960, in the Conway Hall, London W.C.1. Further information can be obtained from the Secretary, 149 Ongar Road, Brentwood, Essex.

Marconi's Wireless Telegraph Company Ltd of England, and Hermes Electronics Co. of the U.S.A. (formerly Hycon Eastern, Inc.) have concluded an agreement providing for general technical collaboration between the two companies in the field of point-to-point communications. Hermes are well known as communications engineers and have planned and installed substantial communications systems. Among other activities Hermes are consultants to the Ace High project for S.H.A.P.E. and to the Ministry of Defence of the Royal Norwegian Government. The two companies intend to collaborate specifically in planning and supplying complete systems. Hermes will also become agents and licensees for Marconi point-to-point radio communications equipment in the United States. Under the agreement, it is also intended that the Marconi Company will become licensees and agents for certain Hermes specialized components and equipment.

The Fairbanks Whitney Corporation of New York, whose Group manufactures products ranging from machine tools, diesel electric locomotives and electric power generators to aircraft engine accessories and missile components under an agreement recently signed in London and New York, undertakes to market most of the range of electronic equipment designed and manufactured by E.M.I. Electronics Ltd. The Corporation will also manufacture some of the British company's equipment under licence. Two EMIDEC 1100 computers are to be installed at the Beloit, Wisconsin, plant of Fairbanks Morse & Co., and the Pratt and Whitney's plant at West Hartford, Connecticut. The computers will be used to serve these two industrial installations for both engineering and business data processing and will also provide demonstration units for the United States market.

A Conference on Components and Materials used in Electronic Engineering is being organized by The Electronics and Communications Section of The Institution of Electrical Engineers, with the support of the Council, at The Central Hall, Westminster, London, from 12 to 17 June 1961. The scope of the Conference falls under three main headings:

(a) Materials. The preparation, and the physical and electrical properties of resistive, dielectric, magnetic, piezoelectric, ferro-electric, magnetostrictive, photosensitive, etc., materials and their application to modern components. Constructional materials such as boards for printed wiring, soldering and encapsulating materials.

(b) Components. Fixed and variable resistors, potentiometers, capacitors and inductors, transformers, transductors, and switches, relays, plugs, sockets and contacts. Small motors and synchros. Equipment wiring, r.f. connecting cables and fuses. Potted and printed circuit components and modular structures. It is intended to put special emphasis on new techniques of measurement, reliability and the effects of extreme operating conditions.

(c) Assembly Techniques. Miniaturization and micro-miniaturization. Methods of automatic assembly. Wrapped and crimped connexions.

The Conference will not cover thermionic devices, transistors and other semiconductor devices, neither will applications be included, except when a reference thereto is essential to a description of the nature of components or materials. The Institution invites the submission of papers for consideration. Further information can be obtained from the Secretary, The Institution of Electrical Engineers, Savoy Place, London W.C.2.

Correction

In the article 'Fast Counting Circuits Using E1T Tubes' by V. Radeka, in the February issue, equations (5), (6) and (7) on page 93 should read:

$$\psi_1 = \ldots\ldots\ldots\ldots \quad (5)$$
$$\psi_2 = \ldots\ldots\ldots\ldots \quad (6)$$
$$\psi_3 = \ldots\ldots\ldots\ldots \quad (7)$$

LETTERS TO THE EDITOR

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

Local Feedback in Transistor Amplifiers

DEAR SIR,—I was interested in the article in the September 1959 issue, by Mr. Pfyffer, in which expressions for the behaviour of transistor amplifier stages are given. While it is almost essential to introduce some approximations to reduce such expressions to manageable proportions this must be done with caution and it seems to me that two of the approximations used by Mr. Pfyffer can lead to excessively large errors.

These approximations are the omission of the exponential term $\exp[-j(mf/f_\alpha)]$ in the expression for α, and the omission of the collector junction capacitance. The exponential term is normally dealt with by taking the first two terms of the power series expansion, giving

$$\alpha = \frac{\alpha_0\,[1 - j(mf/f_\alpha)]}{1 + j(f/f_\alpha)}$$

This is found[1] to give satisfactory representation of α up to the cut-off frequency f_α and leads to the well-known relation

$$f_\beta \simeq \frac{f_\alpha}{(1+m)\;\beta_0}$$

Using this approximation for α and ignoring base width modulation, it will be found that the current gain cut-off frequency of a stage employing both shunt and series feedback is given by (in Pfyffer's notation)

$$f_c = f_\alpha \frac{(R+R_e) + (R_s+R_b)(1-\alpha_0) + \dfrac{(R_s+R_L)\,[R+R_e+R_b(1-\alpha_0)] + R_sR_L}{R_F}}{\left\{(R+R_e) + (R_s+R_b)\,(1+m) + \dfrac{(R_s+R_L)\,[R+R_b(1+m)] + R_sR_L}{R_F} + \dfrac{(R+R_e)[R_s+R_L+R_b+(R_s+R_L)(R_b/R_F)] + R_L[R_s+R_b+R_s(R_b/R_F)]}{X_c}\right\}}$$

where X_c is the reactance of C_c at f_α. Provided R_s and R_L are large compared with R, R_b and R_e this may be further simplified to give

$$f_c = f_\alpha \frac{R + (R_sR_L/R_F) + R_e + (R_s+R_b)\,(1-\alpha_0)}{R + (R_sR_L/R_F) + R_e + (R_s+R_b)\,(1+m) + (R_sR_L/X_c)}$$

It is seen that shunt and series feedback have similar effects. The same expression also defines the voltage gain cut-off frequency, if the source is taken to be a voltage generator of internal impedance R_s, which I suggest is a more realistic case than the zero impedance voltage source considered in the article. Hence the various cases need not be dealt with separately.

The example given by Mr. Pfyffer probably refers to a uniform base transistor for which one would expect to find an m factor of about 0·2 and C_c about 10pF. Using these values in conjunction with Mr. Pfyffer's figures it is found that

$$f_c = 2\times10^4 \left\{\frac{(25+16{\cdot}5) + \dfrac{700}{41}}{(25+16{\cdot}5) + 1{\cdot}2\times700 + \dfrac{600\times600}{800}}\right\}\ \text{kc/s.}$$

$$= 880\text{kc/s}$$

It is seen that omission of m and C_c leads to a value of f_c about 80 per cent too high. Higher values of R_L would produce even larger errors. If drift transistors are used the capacitance effects are less but this is offset by the increased m value.

It can be seen that while feedback substantially reduces the effects of variation in $(1-\alpha_0)$ it does not reduce the effects of variations in f_α. Hence deviations of f_α are usually the most important cause of variation in the complete amplifier stage. The errors in Mr. Pfyffer's expressions may exceed the normal spread of f_α for the transistors used and would not therefore be adequate for estimating the effects of transistor parameter variations on the complete circuit.

Yours faithfully,

D. G. W. INGRAM,

Harrow, Middx.

REFERENCE

1. MIDDLEBROOK, R. D. An Introduction to Junction Transistor Theory, Chapter 11, (Chapman & Hall Ltd. 1957).

The Author replies:

DEAR SIR, — The author agrees in principle with Mr. Ingram's statement, that the collector capacitance has to be taken into account in certain cases. On the other hand it can be shown by comparing experimental results with analytical results of formula (40) and Mr. Ingram's formula for the cut-off frequency that the accuracy for both types of analysis is of the same order. Bearing in mind that the spread in f_α of well-known transistors on the market is as high as 4 (e.g. f_α specified as 7·5 to 30Mc/s) the error introduced by the analysis is not excessive.

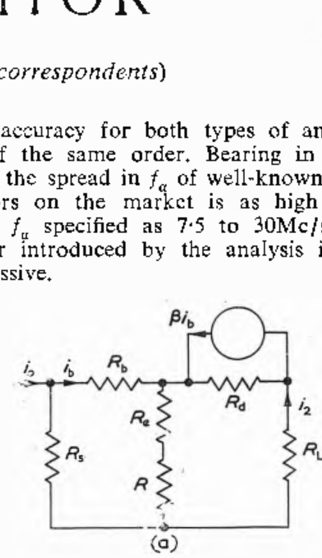


Fig. A. Equivalent circuit used in article

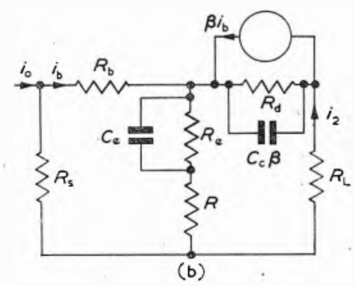


Fig. B. Circuit which Mr. Ingram probably used

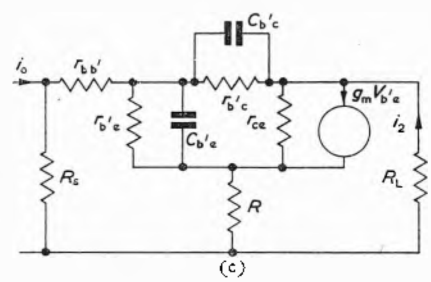


Fig. C. The Hybrid π-equivalent circuit

It is well known that the hybrid π equivalent circuit (see Fig. C) represents the transistor at high frequencies much better, though its application is not as simple as the T-circuit (this has been mentioned in the article). For a quick assessment using average values for the parameters, the T-circuit gives satisfactory results.

The parameter m has to be taken into account, if its value is large, e.g. for drift

TABLE A.

R (ohm)	MEASURED f_x/f_β	FORMULA A f_c/f_β		FORMULA B f_c/f_β		FORMULA C f_c/f_β	
0	2·14	1·58	−26%	1·36	−36%	1·13	−47%
20	2·75	2·50	− 9%	2·06	−25%	1·80	−34%
50	3·34	3·80	+14%	3·18	− 5%	2·75	−17%
100	3·82	5·85	+53%	4·90	+28%	4·30	+13%

transistors (which came on the market after the article was written), where $m = 0{\cdot}7$ to $1{\cdot}0$, and its neglection would lead to an error in the order of 60 per cent.

For comparison, Figs. A, B and C may be considered.

It is obvious from the figures that the application of the *T*-circuit, especially as far as emitter feedback is concerned, is much simpler.

To illustrate the relative accuracy of the analysis with respect to the measured results, the following example may be given.

Transistor: R.F.—junction transistor.

Operating point: $I_c = 1$mA, $V_{ce} = 5$V.

Measured parameters: $f_\alpha = 15$Mc/s; $\beta = 68$; $Z_{IN_0} = 1{\cdot}7$kΩ (INPUT-IMPEDANCE AT $R = 0$); $C_c = 10$pF.

Parameters calculated from above values:

$R_e = 12{\cdot}5\Omega$ $R_b = 840\Omega$ $f_\beta = f_\alpha/(1+\beta) = 215$kc/s.

The cut-off frequency f_x (3dB point)

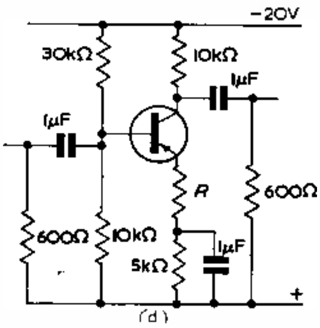


Fig. D. Measuring circuit

was measured in the circuit of Fig. D for various emitter feedback resistors.

In Table A the measured results are compared with the analytical results. (The percentage figures show the deviation from the measured values.)

Used formula:

A (Formula (40) in article)

$$f\gamma = f_\beta \frac{R_s + R_b + (1+\beta)(R+R_e)}{R_s + R_b + R_e + R}$$

B (Mr. Ingram's formula, without using *m*)

$$f_c = f_\beta \frac{R_s + R_b + (1+\beta)(R+R_e)}{R_s + R_b + R_e + R + (R_s R_L/X_e)}$$

C (As B, but considering *m*)

$$f_c = f_\beta \frac{R_s + R_b + (1+\beta)(R+R_e)}{(1+m)(R_s + R_b) + R_e + R + (R_s R_L/X_e)}$$

Yours faithfully,

H. PFYFFER,

The General Electric Co. Ltd, Coventry.

The Bootstrapped Multivibrator

DEAR SIR,—In a multivibrator, it is normal practice to return the grid leak resistors to the positive supply rail. This causes the grid voltage to approach the cut-off voltage of the valve at a fairly fast rate, not much less than the initial rate of rise, and ensures that the period of the multivibrator is stable. However, the width of the pulse produced by this method of connexion is only about 1/6 of the time-constant of the coupling components. If a circuit is to be built with a number of switched ranges giving long pulses, up to a second, say, the capacitors to be used are rather bulky, unless very high resistance values are used which may make the circuit unduly susceptible to insulation defects. Moreover, the highest value resistor covered by a Radio Components Standardization Committee Specification is 10MΩ and those high value resistors available are relatively costly. The circuit to be described was developed to give the same stability as the conventional circuit and to use smaller components.

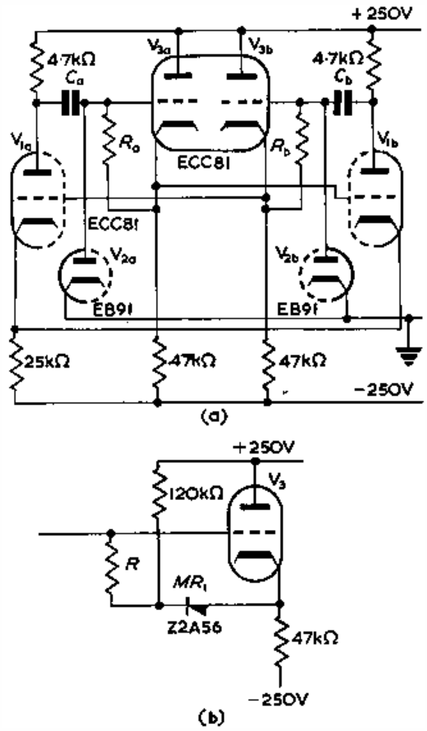


Fig. 1(a). Normal circuit
(b). Alternative connexion for grid leak

The circuit of the bootstrapped multivibrator is shown in Fig. 1(a). The coupling between the anode of one half of V_1 and the grid of the other half is taken through a cathode-follower V_{3a} or V_{3b}. The resistor R_a or R_b which forms part of the timing network, is connected between the grid and cathode of this cathode-follower. Thus, when the anode of V_{1a} goes negative, the capacitor C_a discharges through R_a and because of the bootstrap action of V_{3a}, the rate of charge is nearly constant. Thus the grid potential of V_{1b} approaches cut-off at a nearly constant rate. When the multivibrator turns over, the diode V_{2a} holds the grid of V_{3a} at earth potential.

The voltage across R_a is the standing bias on the cathode-follower V_{3a} and is about 3V. This is very much smaller than the voltage across the grid leak in the conventional circuit, which is the positive supply voltage, and thus the time-constants of the timing networks C_aR_a and C_bR_b can be much smaller. The switching speed of the circuit is also improved, as the smaller values of C_a and C_b used can be charged more quickly through the anode loads.

The stability of the pulse width generated depends on the stability of the standing bias on the cathode-follower V_3. This has been found to change by about 3 per cent for a 10 per cent change in heater voltage. An alternative connexion for the grid leaks, shown in Fig. 1(b), improves the stability of the circuit. The voltage across *R* is increased by the use of the Zener diode MR_1, thus making the variation of grid bias on V_3 a smaller percentage of the total voltage. Using a 5·6V Zener diode in this circuit, the time-constant *CR* had to be increased, but was still much smaller than in a conventional circuit, and the stability was improved to 1 per cent for a 10 per cent change in heater voltage.

A conventional multivibrator circuit, and both alternatives of the bootstrapped multivibrator circuit were built and supplied with stabilized h.t. voltages. The ratio of the time-constant *CR* to the pulse width generated and the stability of the pulse width with ±10 per cent variation of heater voltage were measured. The results are given in Table 1, and from this it may be concluded that a multivibrator circuit has been developed that enables long pulses to be generated using relatively small value timing components, with a slight sacrifice of stability compared with the conventional circuit.

The authors' thanks are due to the Director of the Atomic Weapons Research Establishment, Aldermaston, for permission to publish this note.

Yours faithfully,

T. RING and W. G. GORE,

Atomic Weapons Research Establishment, Aldermaston.

TABLE 1.

	RATIO *CR* : PULSE WIDTH	STABILITY (per cent)
Conventional Multivibrator	6·7	<±1
Bootstrapped Multivibrator (Normal)	0·11	±3·1
Bootstrapped Multivibrator (Alternative)	0·25	±1

Correction

In Fig. 1 of the letter from M. Pacák on page 172 of the March issue, the output voltage should be shown as V_2 = 260-2 300V and the voltage across RV_1 RV_2 as 0-2 060V. The value of RV_1 should be 9 × 100kΩ and the voltage at the secondary of the mains transformer should be 240 + 9 × 90V.

BOOK REVIEWS

Programming Business Computers

By D. McCracken. 487 pp. 69 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 82s.

IN this book the authors have attempted the formidable task of giving the reader a detailed understanding of the concepts and techniques of electronic data-processing as applied to business and commercial applications.

After a brief introduction to the basic requirements of any manual or automatic data processing system, the fundamental concept of the file is treated at some length. Chapter 3 stresses the importance of an organized approach to the mechanization of a system, leading to the use of flow charts. The notations adopted in this section allow lucid charts to be made for any complex processing operation, and are worthy of serious consideration by any group which has not evolved a satisfactory scheme of similar merit.

A brief introduction to the internal structure of a data processor follows, with a very condensed history of computers from the ubiquitous Babbage. The chapter also introduces DATAC, '. . . a hypothetical computer which is a compilation of the features of many machines.' This is not the usual illustrative computer on which a few trivial operations can be demonstrated, but has a fully developed structure which is truly representative of much modern practice. The next ten chapters are, in effect, a programming manual for this machine. It may be argued that the effort necessary to absorb this material (especially if the conscientious reader also tackles the exercises appended to each chapter) could be better devoted to an available system. However, a newcomer to the use of computing systems who has studied the programming of DATAC should find no difficulty in applying the techniques to any other system. Throughout this part of the book the relative merits of various alternative facilities are discussed (for example, fixed versus variable block recording on magnetic tape) although naturally the facilities not provided in DATAC are not so fully treated.

A significant omission from this section is any discussion of 'parallel programming,' provided in some systems so that, for example, a high priority file updating operation may proceed essentially simultaneously with a lower priority operation such as a punched card to magnetic tape conversion. A number of variations of this basic proposal are incorporated in recently announced systems, but their operational value and the price in complexity of programming remain doubtful quantities. A discussion of this topic, even if the authors found themselves unable to make positive recommendations, would have been welcome. They have, with an assurance which may surprise some, asserted that independent buffering of peripheral devices is obsolescent and that time-sharing of the main store for this purpose will soon be universal.

The remaining chapters, including part of the last one of the DATAC group, devoted to machine aided coding, review a number of topics important not only to those involved in programming and coding, but to others whose interests are in overall systems design, auditing, etc. Among the subjects covered are methods of sorting, arrangement of systems to minimize loss of information and time in the event of operators' errors or machine failures, the uses of random access files and the problems of auditing and data protection.

Appendices, apart from a collected summary of the DATAC system's characteristics, cover number representation and a 'data-processing diary.' This last is a verbatim reproduction of daily jottings during the introduction of a data-processing system, but is so concerned with meaningless detail that its omission would not be felt. A glossary of terms and an extensive and valuable, but perhaps uncritical, bibliography complete the book.

The authors have achieved a uniform and very readable style, and although their examples are drawn from American practice sufficient detail is given for the processes to be readily understood. Presentation is up to the usual high standards of the publisher and the book can be confidently recommended to anyone concerned with operation or organization of a commercial data-processing system, or to engineers requiring an insight into the uses to which these systems are applied.

A. C. D. HALEY

Basic Audio. Volumes I, II and III

By N. H. Crowhurst. 349 pp. 371 figs. Demy 8vo. John F. Rider Publisher Inc., New York. Chapman & Hall Ltd, London. 1959. Price 69s. per set.

THIS elementary course on the principles of apparatus intended for sound reproduction appears in three volumes totalling 349 pages. Each deals with a group of subjects, more or less according to their complexity, although there is no attempt at sequential treatment. Therefore, since one must have the set to complete the picture, the inference is that subdivision into different volumes is to ease the burden of paying for all at once. The price is scandalously high, each book costing over a pound, paper backed. At the same time, Mr. Crowhurst has covered an immense amount of ground in an able and ingenious manner, and does indeed explain the function of everything connected with reproduction, except transducers or pick-ups. Nor are the contents written down to the curiously childish level of the many illustrations, some of which are much harder to understand than the text.

If one is prepared to overlook the typically 'popular American' pictorial presentation, the books must certainly rank as the most comprehensive so far written for the layman. Mr. Crowhurst has of course made some condensations, but against this, a great many design problems are fully treated and plenty of numerical examples are given. Indeed, on careful reading, there is an answer to every problem likely to perplex the amateur. The treatment is ingenious, quite complex calculations being simplified in a manner only possible by one long experienced in these matters.

This book is a very real attempt to enlighten the amateur, serviceman or knowledgeable musician, and in this respect it succeeds admirably. The stamp of Mr. Crowhurst's experience is everywhere. The contents are recommended as a mine of information, particularly on amplifier design.

A. DOUGLAS

Electronic Switching, Timing and Pulse Circuits

By J. M. Pettit. 259 pp. 294 figs. Medium 8vo. McGraw-Hill Publishing Co. Ltd, New York, London. 1959. Price 58s.

As the title states, this work is about circuits and what they do. The emphasis is on non-linear circuits rather than linear, and steady state sine waves give place to transient conditions throughout. The treatment is fairly elementary although the circuits dealt with are 'real' in the sense that they are not over idealized. The mathematical analysis is kept to a minimum, and the fundamental formulae, such as the voltage decay of a series *RC* circuit, are assumed known, although the reader is invited to derive many of them in the problems given at the end of the chapters. The methods of solution are mostly graphical, and it is not until Chapter 8 on pulse forming networks that fourier series appear. This, of course, is made inevitable by the presence of Gibbs' Phenomenon, and methods, such as 'Guillemin's network' for avoiding its effect, which are discussed very lucidly in this chapter.

To give some idea of the scope of the book it might be mentioned that the first six chapters are devoted to lumped circuits employing *R* and *C* with a transistor or thermionic valve. Thus, after an introductory chapter comes one on circuit fundamentals. This includes a section on some useful circuit theorems where the author has been careful to draw distinction between those theorems which are applicable to non-linear as well as linear circuits, and those applicable to linear circuits only.

The third chapter deals with electron devices as switching elements, and has sections dealing with the characteristics of semiconductor diodes, as well as thermionic diodes. Also their equivalent circuits are examined.

The fourth chapter gives some circuit examples as an illustration of the techniques developed in the preceding chapters.

The fifth chapter is concerned with multivibrators, stable states and regenerative switching, and the sixth chapter deals with feedback circuits for generating linear voltage drops.

The inductance L does not put in an appearance until chapter 7, where the characteristics and uses of pulse transformers are gone into.

All the circuits in the first seven chapters are lumped, and distributed networks are left to the final chapter 8.

This is a good book which aims at giving an insight into the behaviour of non-linear circuits, and it succeeds.

If one wishes to find fault one could mention that the problems set at the end of the chapter might prove more valuable if some at least, were followed up by a short discussion of the solution. Again, at the end of chapter 6, the author states that the objective of the chapter was to trace the similarities of the various circuits discussed, rather than weigh up their relative merits. One wonders whether a discussion of the differences following after a discussion of the similarities might not have enhanced the value of the chapter. However, these are small points, and not everyone will agree with them.

The book is admirably laid out and the figures are clear. There is an index and many references.

R. H. C. Newton

Electronic Circuit Theory

By H. J. Zimmermann and S. J. Mason. 558 pp. 619 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd. London. 1959. Price 86s.

UNIVERSITY teaching in electronics can be broadly divided into two parts, the first deals with linear circuits and the second with non-linear circuits. In most textbooks the linear work, starting from the idea of a single equivalent circuit for the thermionic valve, is adequately covered, but the treatment of the non-linear case is less satisfactory and tends to be restricted to the operation of one or two applications, such as the multivibrator. Within the last year or two the electrical engineering course at M.I.T. has been remodelled so that the earlier part of the syllabus is concerned only with fundamentals—the core curriculum—and applications are dealt with in the final year. The book under review covers the core material in electronic circuits, with emphasis on the non-linear mode of operation. The book recognizes that both valve and transistor action in non-linear circuits can be understood by considering equivalent circuits made up of an energy source associated with one or more diodes together with passive elements. The method of attack is known in the United States as 'piecewise linear analysis', although it is essentially the same as the step-by-step methods normally used by engineers.

The book deals very briefly with physical ideas, such as the equation representing the behaviour of a pn junction, and the reader is encouraged to think wholly in terms of the equivalent circuit. The result is that the student may try to get more out of the equivalent circuit than is justified by the initial data. This is particularly so in the case of transistor circuits and it is fairly clear that a teaching course based on this book would need adequate laboratory work to ensure that the student kept the whole field in proper perspective.

V. H. Attree

Circuit Theory of Linear Noisy Networks

By H. A. Haus and R. B. Adler. 79 pp. 21 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd. London. 1959. Price 36s.

OVER the past few years a great deal of work has been carried out, mainly by the authors, concerned with the rationalization and the unification of the subject of the noise performance of linear amplifiers. The present small volume, which forms one of a new series of research monographs from the Massachusets Institute of Technology, collates and extends this material.

One of the concepts brought out by the authors' work has been that of exchangeable power, i.e. the stationary value of the power which can be given out or absorbed by the source. By the use of this concept negative resistance devices, such as are now commonly being encountered in the parametric amplifier and maser fields, can be treated in a similar manner to the more usual positive resistance devices. A further innovation has been that of 'noise measure,' which involves both noise factor and gain, and which forms a much better criterion of the noise performance of an amplifier than noise factor alone. The present work shows that in the general case, the noise measure of any number of interconnected amplifiers cannot be better than the noise measure of the best of these amplifiers.

Being a research monograph, little concession is made to the general reader, and a sound knowledge of matrix algebra and its application to network theory is assumed. With its help, the first half of the book is devoted to the lossless transformations that can be made to the multiple terminal pair noisy network, and from consideration of the quantities which remain invariant, a general characteristic noise matrix is developed. The remainder of the book applies the general results to the discussion of the noise measure of two terminal pair amplifiers and the realization of the optimum noise performance.

Those concerned with work on low noise devices will find much to stimulate them in the general logical treatment of amplifiers given in this book.

W. H. Aldous

Handbook of Electronic Control Circuits

By J. Markus. 326 pp. 478 Figs. Demy 4to. McGraw-Hill Publishing Co. Ltd. New York, London. 1959. Price 66s.

THIS comprehensive compilation of over 250 electronic circuits, each complete with values of components, will be useful to industrial electronic and design engineers. From it, the salient points of a number of circuits of a given type can be quickly obtained, as a guide for choosing the most promising circuit for a particular problem. Each discussion covers a circuit based on equipment now in actual operation, and this proves the practical usefulness of any circuit being considered for adaptation.

Electronic Flash Photography

By R. L. Aspden. 179 pp. 100 figs. Demy 8vo. Temple Press Limited. 1959. Price 37s. 6d.

THIS is a survey of principles and practical techniques in industry, research and radiography. The author, who is responsible for high-speed photographic research and development at the Royal Aircraft Establishment, in the opening chapters deals with radiation, fundamental discharge-tube processes, flash characteristics, tube design, control and high-voltage circuits. A consideration of special cameras and photographic recording is followed by a description of specialized techniques and their practical application in specific research projects. A separate chapter covers the principles of flash radiography.

ELECTRONIC EQUIPMENT

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

(Voir page 257 pour la traduction en français: Deutsche Übersetzung Seite 262)

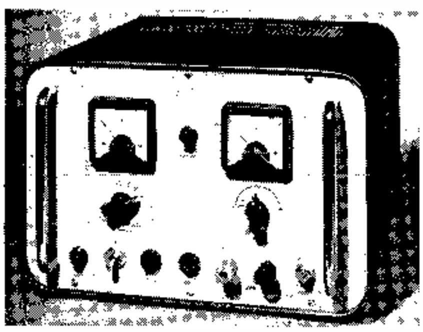
TRANSISTORIZED POWER UNITS

(Illustrated below)

Metrix Instruments Ltd, 54 Victoria Road, Surbiton, Surrey

A comprehensive series of cabinet models has now been added to this firm's range of high stability transistorized power supply units.

Six models cover the range 0 to 10V at 0·5A and 0 to 50V at 1A, 2A, 6A or 10A. The output voltage, which is 'fully floating', is accurately adjusted by a coarse and fine control, and two meters indicate the load voltage and current. The stability ratio with respect to mains is better than 2 000:1 and the source impedance is less than 5mΩ. The peak-to-peak residual ripple is less than 2mV on the 10A model and less than 0·5mV on models with lower current ratings.



The 2A, 6A and 10A models are fitted with two additional 'line loss' output terminals. By connecting these terminals directly to the actual load, the true load voltage, not the voltage at the output terminals, is indicated on the meter and stabilized by the control amplifier. The effect of loss in the connecting leads is therefore completely compensated.

All models are fully protected against overload and component failure. A microsecond trigger device biases the supply down to zero volts in the event of an overload surge, and a relay augments this function and cuts off the supply. The overload cut-out cannot be reset until the fault has been cleared. In addition fuses are included in the mains input and rectifier circuit.

EE 20 751 *for further details*

ULTRA LOW FREQUENCY OSCILLATOR

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

The type 445 u.l.f. oscillator is continuously variable over the frequency range from 0·01 to 1kc/s by resistance-capacitance tuning. The well-established advantages of this form of tuning are combined with a novel selective amplifier circuit with controlled positive feedback to maintain continuous oscillation.

The amplifier consists of a phase invertor followed by two integrating stages and is made selective by the use of overall negative feedback.

The positive feedback circuit includes a symmetrical diode limiter. The constant-amplitude rectangular waveform from this limiter is applied to the input of the amplifier. The high sensitivity of the amplifier reduces the harmonic content of the output signal to negligible proportions.

The diode limiter has the advantage over comonly employed thermal control circuits of providing rapid amplitude control right down to the bottom limit of frequency. After a sudden transient, as for example, when changing ranges, the amplitude of oscillation is stabilized within about three cycles at all frequencies.

At the frequency of oscillation, each of the two integrators gives a phase shift of very nearly 90°. A quadrature output is therefore available after the first integrator stage, which is particularly useful for testing servo systems. The output impedance of each integrating stage is very low, so that the specified load impedance can be connected directly without affecting the performance of the oscillator.

On the 0° output, the output level is continuously variable up to a maximum of 15V into a minimum load of 10KΩ, this output being constant within ±4 per cent over the entire frequency range. The distortion is less than 0·5 per cent total over the measured range from 20c/s to 1kc/s. Hum and noise are less than 0·1 per cent.

The 90° output has a phase accuracy of 90°±0·5° (lagging) referred to the 0° output measured at 400c/s. The maximum output level at 90° is within ±3 per cent of the 0° maximum output. Distortion at 90° is less than 1·5 per cent total over the measured range from 20c/s to 1kc/s. Hum and noise are again less than 0·1 per cent.

This instrument is a further development of a patented circuit design originated by the Royal Radar Establishment.

EE 20 752 *for further details*

HANOVER FAIR

'Electronic Engineering' will occupy Stand 2 in Hall 13 at the Hanover Fair from 24 April to 3 May and visitors will be welcome.

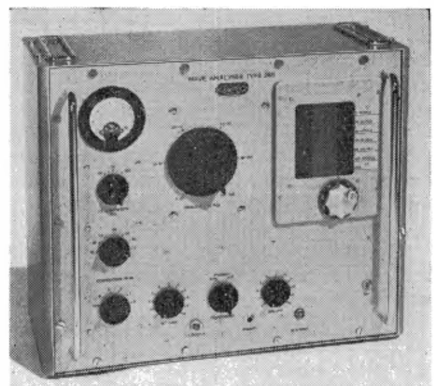
V.H.F. WAVE ANALYSER

(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire

The wave analyser type 248 is a selective measuring set designed for a wide variety of quantitative measurements in the frequency range 5 to 300 Mc/s. It is particularly suitable for the evaluation of harmonics and components of a complex waveform, and may also be used for field strength and interference measurements, indication of insertion gain and loss, and as a bridge detector or heterodyne wavemeter.

Direct reading of the signal amplitude is provided by reference to the settings of two step attenuators and a



panel meter. Second harmonics up to 55dB and higher harmonics up to 70dB below the fundamental may be measured.

The analyser consists basically of a frequency changer followed by a highly stable i.f. amplifier operating at supersonic frequency and incorporating a series of calibrated step attenuators which, by virtue of the low i.f. are free from frequency errors. A meter, protected against overload, is provided for interpolation between steps, and external h.f. attenuators may be plugged into the input socket, giving a total attenuation of 100dB. Further advantages of the low i.f. include the elimination of a second channel, absence of gaps in the frequency coverage due to the i.f. falling within the operating range, and the excellent shape of the response curve, with steep sides and a sharp null point facilitating accurate frequency location.

Amplitude or frequency modulation on the signal may be detected, and the audio signal is available at a telephone jack. An i.f. output is also provided for external monitoring.

The signal frequency is indicated on a direct-reading film scale, driven by 90:1 gearing and having a calibration accuracy of 2 per cent. The effects of noise are reduced to a minimum by the use of three alternative i.f. bandwidths, and the instrument is inherently stable against supply voltage variations.

EE 20 753 *for further details*

PORTABLE BETA GAUGE

(Illustrated below)

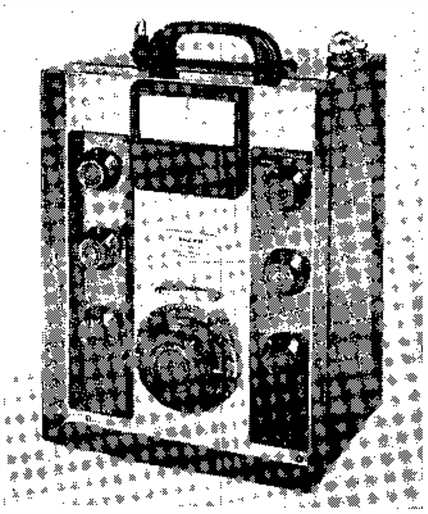
Baldwin Industrial Controls, Lowfield Street, Dartford, Kent

Baldwin Industrial Controls have recently introduced into their range of nucleonic thickness gauges a small portable mains operated unit.

It has been designed to meet the requirements of those sections of industry where the expense and refinements of the standard beta gauge are not justified.

The gauge is calibrated in arbitrary units proportional to weight per unit area and the user can easily correlate a factor for his own particular product.

For simplicity, the gauge embodies



only one radioactive source and detector. The output from the detector is compared with a calibrated reference potential and the difference in signal is amplified by an extremely stable low gain d.c. amplifier.

The output signal is sufficiently powerful to operate both the built-in moving-coil meter and, if required, an additional external recorder.

Although the instrument has been simplified to meet a specific demand, the quality and accuracy of its measurements are excellent within the limitations imposed by economics.

Various measuring head mountings are available according to the application or trade requirements, and alternative radioactive sources are offered depending on the range and thickness or weight per unit area of material to be measured.

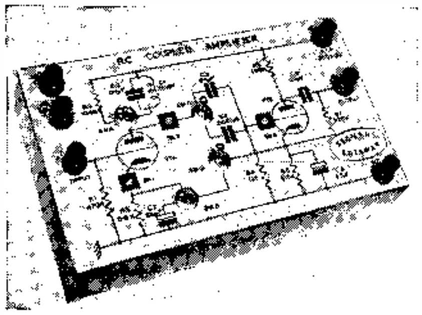
EE 20 754 *for further details*

EDUCATIONAL AIDS

(Illustrated above right)

Associated Electrical Industries (Woolwich) Ltd, 155 Charing Cross Road, London W.C.2

These units have been designed in collaboration with Northampton Polytechnic to form the basis of an extensive range of laboratory experiments to be



carried out in conjunction with courses in radio and electronic theory.

Eleven chassis are available, to teach the basic connexions of International Octal, B7G and B9A valves, and the operation of an *RC*-coupled amplifier, *RC*-coupled and tuned-anode oscillators, low-pass and high-pass networks, and a power output stage.

The units have been designed to operate in the audio frequency range, so that associated equipment—power supplies, audio oscillator, oscilloscope—is easily obtainable. The manufacturers are able to supply suitable power packs and an audio oscillator.

The circuit diagram of each major unit is printed on the top of the chassis, and where a terminal or switch occurs in the theoretical circuit, there the actual component is mounted. To clarify the lesson, the student can view the wiring and components beneath the chassis through a Perspex base, and compare the layout with the circuit diagram.

Each unit is supplied with instruction sheets covering a number of experiments and the underlying theory. These experiments are designed to cater for the needs of students at both elementary and more advanced levels.

The illustration shows the *RC*-coupled amplifier type R.2330 which is a versatile unit to which switches are fitted for altering coupling capacitances, introducing feedback or bringing a tuned anode load into circuit.

EE 20 755 *for further details*

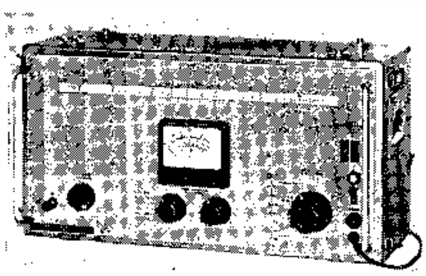
WIDE RANGE L.F. SIGNAL GENERATOR

(Illustrated below)

Distributed in U.K. by R. H. Cole (Overseas) Ltd, 2 Caxton Street, London S.W.1

This signal generator, which is manufactured by Siemens and Halske, covers the wide frequency range of 10c/s to 1Mc/s in seven switched bands. This frequency range is achieved by combining an *RC* with an *LC* generator.

High amplitude and frequency stability



together with low harmonic content are maintained over the whole range and the sinusoidal output voltage can be varied continuously and in steps from 20µV to 20V.

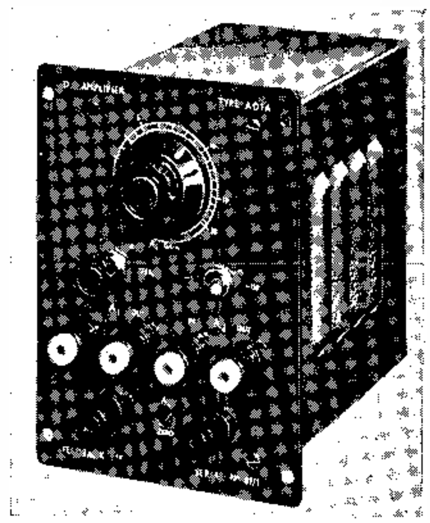
EE 20 756 *for further details*

VARIABLE GAIN D.C. AMPLIFIER

(Illustrated below)

Feedback Ltd, 1 Broadway, Crowborough, Sussex

The type AD1A amplifier consists of two amplifiers in a single unit. One amplifier has an operating range of d.c. to 100kc/s (±3dB) with balanced input stage to reduce drift effects and gain continuously variable from +1 to +20 by feedback adjustment. The other amplifier can be switched to give a gain of either −1 (d.c. to 100kc/s) for phase reversal, or −150 (d.c. to 40kc/s) for addition, integration, or function genera-



tion using external networks. Both amplifiers have high input impedance and low output impedance with a maximum output of ±50V, ±2mA. The two amplifiers may be used independently or connected in series to give maximum gain of −3 000 for more precise function generation.

The AD1A amplifier is intended for experimental work on aspects of system frequency and transient response, and as a unit for analogue computer arrangements; many such applications requiring only moderate accuracy and performance and the main need being for a unit which is flexible and adaptable and which contains some built-in control facilities.

The amplifier is one of a range comprising transfer function units, a square-wave unit, servo test waveform generator, and stabilized d.c. power supplies.

EE 20 757 *for further details*

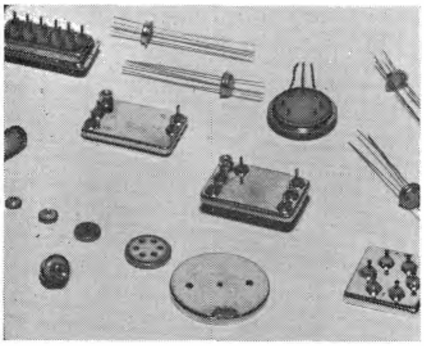
GLASS TO METAL SEALS

(Illustrated overleaf)

Wesley Coe (Cambridge) Ltd, Scotland Road, Cambridge

A range of glass to metal seals, either as standard types or to customers requirements, is available from Wesley Coe Ltd.

The compression seal can be made



as a single outlet, suitable for diode housings and similar uses, these are very rugged seals and are suitable for mass production projection welding. They can also be made as multi lead or tubular headers for relay, transistors and many other applications. The matching seal using Nilo K and a suitable glass can be used for all types of headers where very good hermetic sealing is required and these can be made into single and multi pin headers of all types, particularly where saving of weight is important in airborne equipment.

Various finishes can be given to customers requirements, such as hot tin dip, electro tin, copper, silver and gold.

A.I.D. Release can be given on all seals.

EE 20 758 *for further details*

WIDEBAND AMPLIFIERS

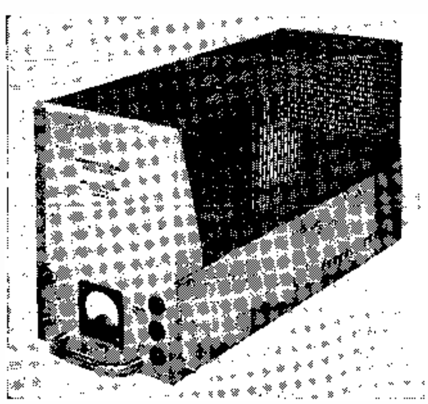
(Illustrated below)

R. H. Minns Electronics, The Lower Mill, Kingston Road, Ewell, Surrey

A periodic amplification over the range of frequencies 1Mc/s to 30Mc/s is provided by the wideband amplifier type VM30.

A number of these amplifiers has been supplied to the British Broadcasting Corporation to provide drives for h.f. broadcasting transmitters operating in that frequency range. An output of 15W into an impedance of 100Ω is obtained for frequencies up to 26Mc/s with 1V applied across the 100Ω input circuit. The output falls to 10W at 30Mc/s for the same input. Models are available for various input and output impedances, coaxial or balanced. Amplification is carried out in three stages, the last two using Mullard QQV06 40A valves.

A wideband distribution amplifier type VM40 is also available, enabling up to three of the wideband amplifiers type VM30 to be fed from a single source. There is a voltage gain of 3 : 1 between the input and each output of the distribution amplifier.



EE 20 759 *for further details*

MINIATURE CRADLECLIP

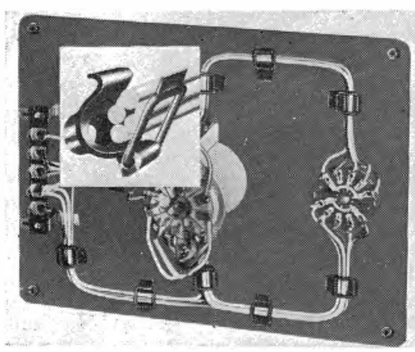
(Illustrated below)

Insuloid Manufacturing Co. Ltd, Sharston Works, Leestone Road, Wythenshawe, Manchester

The Insuloid Cradleclip wiring system has been in use for a number of years; the miniature Cradleclip has now been introduced for use in electronic equipment etc and will accommodate wiring runs up to ⅜in diameter.

Miniature Cradleclip components consist of cradles and binders moulded from nylon, with captive securing clips of tough but flexible moulded neoprene.

The cradles have a recessed hole in the boss to allow mounting by bolt or screw to panel or base in any planned position in a wiring scheme. The wiring runs are then secured by bringing round the neoprene clip and clipping in position in the groove of the cradle.



Where it is not required to anchor the wiring runs to panel or base, the binder is used.

A further refinement is the type TCO cradle which features a threaded metal insert in the boss of the cradle and makes possible the pre-fabrication of wiring schemes. Thus allowing the miniature cradleclip to be fastened from the back or exterior of the panel or cubicle at a later stage.

The system has a temperature range of approximately −60°C to +100°C and gives complete electrical and mechanical security under all extremes of climatic conditions.

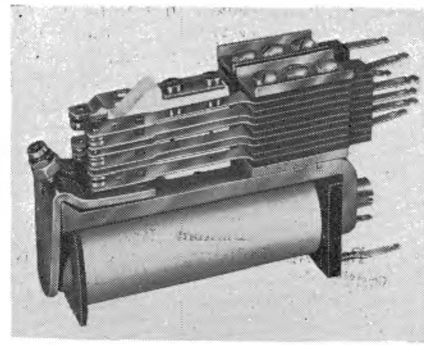
EE 20 760 *for further details*

REMANENT RELAY

(Illustrated above right)

L. E. Simonds Ltd, 5 Byron Road, Harrow, Middlesex

This relay is based entirely on the well-known Post Office Type 3000 design, but is fitted with a remanent core. After energizing, the relay will remain latched indefinitely, and can only be returned to the non-operate position by a pulse in the opposite direction, i.e. either by a reverse polarity where a single coil is employed, or by passing current through a second winding on the same coil.



The relay will hold in even under considerable shock and vibration conditions.

The contact arrangements are very comprehensive, and can be supplied to customers' specification.

EE 20 761 *for further details*

A.F. POWER METER

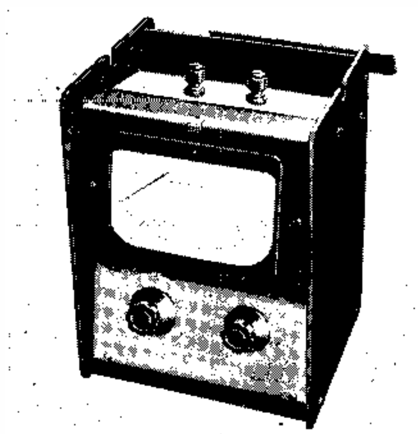
(Illustrated below)

Marconi Instruments Ltd, St. Albans, Hertfordshire

This new a.f. power meter, type TF1347, is a sensitive, accurate instrument for audio measurements over a wide range of impedance (2·5Ω to 20kΩ in 11 steps). Its frequency characteristic is substantially flat from 50c/s to 20kc/s and its 10 power ranges, covering 10μW to 6W, and 5½in meter provide excellent discrimination.

Functionally, the TF1347 may be considered as a variable attenuator whose input impedance is adjustable in steps and whose output is applied to a voltmeter calibrated in terms of power input to the attenuator.

The circuit of the instrument consists of three parts providing impedance selection, attenuation, and indication respectively. Impedance is selected by means of a tapped matching transformer which is connected to the input terminals of the instrument via an 11-position switch; the design of this transformer largely contributes to the level response of the power meter over so wide a frequency range. A 10-step resistive attenuator forms the power range multiplier and presents a constant impedance to the secondary winding of the matching transformer. The attenuator is terminated by the indicator circuit which consists of a



moving-coil meter and bridge-connected rectifier.

By means of the multiplier switch and the meter calibration, the output from an audio frequency system may be read directly in units of power, and also in decibels relative to 1mW. The accuracy (at 1kc/s and 20°C) for powers of 200μW and greater is ± 5 per cent of the reading, from full-scale to half-scale; $\pm 2\frac{1}{2}$ per cent of f.s.d. from half-scale to 1/10th full-scale. For powers of less than 200μW the accuracy is $\pm 10\mu$W.

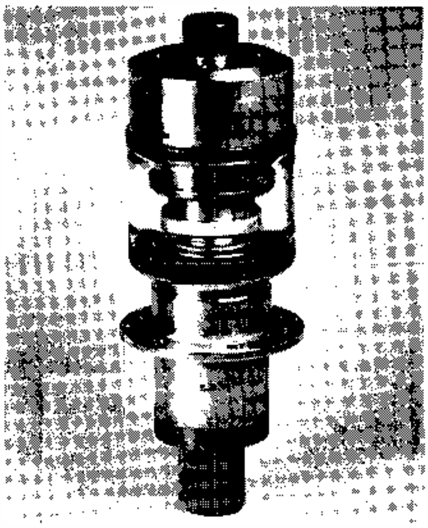
EE 20 762 *for further details*

HIGH VACUUM VARIABLE CAPACITORS

(Illustrated below)

English Electric Valve Co. Ltd, Waterhouse Lane, Chelmsford, Essex

The English Electric Valve Company Limited have developed a series of High



Vacuum Variable Capacitors for use in transmitter tank circuits and similar applications.

The types in this entirely new range of products are characterized by small physical size, high voltage and current ratings, and the ability to stand breakdowns caused by temporary overloads.

Over the specified range, the capacitance of each type varies linearly with the rotation of the capacitor shaft and the precision of the construction ensures a high degree of accuracy in the capacitance change produced by each turn. The capacitance range extends slightly above and below the limits specified below but the capacitance variation in the region below the bottom limit is not linear.

A summary of the major characteristics for a number of the types now in development is given below:

EE 20 763 *for further details*

EEV TYPE	LINEAR CAPACITANCE RANGE (pF)	SHAFT TURNS IN LINEAR CAPACITANCE RANGE	MAXIMUM PEAK R.F. VOLTAGE (kV)	MAXIMUM R.F. CURRENT (A.r.m.s.)
U30/15	5 to 30	10.4	15	10 (up to 30Mc/s)
U50/15	8 to 50	10.4	15	15 (up to 30Mc/s)
U80/15	10 to 80	10.4	15	20 (up to 30Mc/s)
U200/10	15 to 200	16	10	20 (up to 20Mc/s)

MAGNETIC AMPLIFIERS

De Havilland Propellers Ltd, Hatfield, Hertfordshire

A magnetic 'thermocouple' amplifier capable of accepting very low level input signals, with high gain and low drift has been developed recently. It has the following characteristics:

Input level	10-0-10mV d.c.
Output level	10-0-10mA d.c.
Voltage gain	1 000
Input impedance	100Ω/winding
Output impedance	1kΩ
Number of inputs	1-16
Frequency response	Flat to 20c/s
Linearity	2 per cent of maximum output
Drift with time (Temperature and supply voltage constant)	10^{-13}W at input
Drift with temperature (supply constant)	gain $-0{\cdot}15\%/°C$ Zero 2×10^{-14}W/°C
Drift with change of supply voltage (10 per cent)	$1{\cdot}27 \times 10^{-11}$W
Battery voltage	10V
Operating frequency	2-4kc/s
Size, approximately	8in × 6in × $3\frac{1}{2}$in
Weight, approximately	4lb

The amplifier, wound in high performance toroidal cores throughout, consists of two push-pull stages in cascade with overall negative feedback. Magnetic instead of electric feedback is used to provide insulation between input and output and thus 1 to 16 isolated inputs can be used. Magnetic rate feedback is also used to ensure good transient response. A new winding technique, together with selection of components, has enabled the high gain and low drift to be obtained without resorting to second harmonic modulator circuits. The battery-driven transistor invertor works at a frequency high enough to ensure adequate frequency response.

EE 20 764 *for further details*

MOULDED TRACK POTENTIOMETER

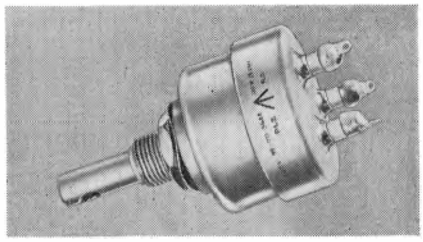
(Illustrated above right)

The Plessey Co. Ltd, Ilford, Essex

The range of moulded track potentiometers produced by The Plessey Company has been augmented by a new hermetically sealed type, the XP5.

Designed to operate within the temperature range $-40°C$ to $+70°C$, the new potentiometer conforms with the Inter-Service Standards for radio components. It will withstand severe bumping, vibration and tropical exposure.

The XP5 is entirely moisture-proof.



The spindle is sealed with neoprene rings which, seated in channels and lubricated with anti-freeze grease, give a smooth action free of bearing slop. The silver-plated terminals are integrally connected to the resistance element and protrude through sealed ceramic insulators. A neoprene washer fitted to the mounting face provides the panel seal essential for use in sealed equipment.

The resistance range extends from 500Ω to 2·5MΩ (linear or logarithmic). Maximum working voltage across the resistance is 500V d.c. subject to power rating limitations.

EE 20 765 *for further details*

PUSH-BUTTON SWITCH

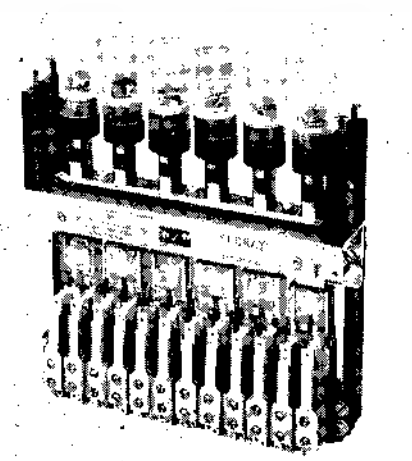
(Illustrated below)

Seton Creaghe Engineering Ltd, G.W. Trading Estate, Park Royal Road, London N.W.10

This switch is suitable for high frequency applications and is designed to reduce maintenance and the likelihood of mechanical breakdown.

The switch can be made with any number of buttons up to 16, and the buttons can be illuminated if required. Each button has a standard contact arrangement of two double-pole changeovers. Only one push-button can be operated at one time and the contacts of the operated button are cleared before a new contact is made. This arrangement can be altered to give 'make or break', and electro-mechanical interlocking facilities are available. The switch can be banked horizontally or vertically.

Tests have been made for a contact rating of 50V, 250mA d.c. non-inductive, and contact resistance is less than 20mΩ throughout life. Capacitance between contacts is less than 2pF between springs and less than 1pF between sets.



EE 20 766 *for further details*

Meetings this Month

THE BRITISH COMPUTER SOCIETY

Date: 28 April. Time: 2.30 p.m.
Held at: Northampton College of Advanced Technology St. John Street, London E.C.1.
Lecture: Automation and Uncertainty.
By: S. Beer.

Belfast Branch

Date: 28 April. Time: 7.15 p.m.
Held at: 29 University Square, Belfast.
Annual General Meeting.

Hull Branch

Date: 6 April. Time: 7.30 p.m.
Held at: Hull Chamber of Commerce, Samman House, Bowlalley Lane, Hull.
Lecture: Application of Computers to the Design and Operation of Chemical Plant.
By: E. Poulson.

Manchester Branch

Date: 5 April. Time: 7 p.m.
Held at: College of Science and Technology, Sackville Street, Manchester.
Lecture: Choosing and Installing a Computer A Case Study.
By: A. J. Platt.

Middlesbrough Branch

Date: 7 April. Time: 7.30 p.m.
Held at: Constantine Technical College, Borough Road, Middlesbrough.
Lecture: The Simulation of Melting Shop Operations on a Computer.
By: R. Neate.

Newcastle Branch

Date: 5 April. Time: 7 p.m.
Held at: University Computing Laboratory, 1 Kensington Terrace, Newcastle.
Lecture: Feasibility Studies and Fundamental Thinking on Integrated Systems.
By: R. M. Jacobs.

South Wales and Monmouthshire Branch

Date: 28 April. Time: 6.30 p.m.
Held at: Small Shandon Lecture Theatre, University College, Cardiff.
Symposium on Auto Codes.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1, commencing at 6.30 p.m.
Date: 7 April.
Lecture: The Work of the B.S.I. in Relation to the Radio and Electronics Industry.
By: H. A. R. Binney.
Date: 27 April.
Lecture: Electronics in Oceanography.
By: M. J. Tucker.

Computer Group

Date: 13 April.
Lecture: Guided Weapon Control.
By: F. R. J. Spearman.

Medical Electronics Group

Date: 21 April.
Lecture: Nerve Impulses from Stretch Receptors in Muscle.
By: Dr. J. G. Nicholls.

North Eastern Section

Date: 13 April. Time: 6 p.m.
Held at: Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Annual General Meeting, followed at 6.30 p.m. by Chairman's Address—The Development of Electronics in the North East.
By: J. Bilbrough.

South Midlands Section

Date: 29 April. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: The Application of Semi-Conductor Devices in Power Supplies.
By: D. D. Jones.
Followed by Annual General Meeting.

BRITISH SOUND RECORDING ASSOCIATION

Date: 22 April. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: Sound Radiation from Loudspeaker Cabinets.
By: J. Moir.

THE INSTITUTE OF PHYSICS

Liverpool and North Wales Section

Date: 7 April. Time: 7 p.m.
Held at: The University, Liverpool.
Lecture: New Techniques in Electron and X-ray Microscopy.
By: C. W. Oatley.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London Meetings will be held at Savoy Place commencing at 5.30 p.m.
Date: 28 April.
Lecture: Radar Observations of Birds and 'Angels'.
By: E. Eastwood.

Electronics and Communications Section

Date: 4 April.
Lecture: V.H.F. Sound Broadcasting: Subjective Appraisal of Distortion due to Multi-Path Propogation in F.M. Reception.
By: R. V. Harvey.
Date: 27 April.
Lecture: Henri de France Colour Television System.
By: R. Chaste and P. Cassagne.

North-Eastern Measurement and Electronics Group

Date: 4 April. Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle-upon-Tyne.
Annual General Meeting followed by Informal Evening on Rocket Instrumentation.

North Midland Utilization Group

Date: 26 April. Time: 6.30 p.m.
Held at: Leeds and County Conservative Club, South Parade, Leeds 1.
Lecture: Silicone Electrical Insulation.
By: J. H. Davis.

North Scotland Sub-Centre

Date: 7 April. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Lecture: Engineering Education in the Technical Universities in Western Germany.
By: D. B. Welbourn, Professor D. B. Spalding and G. L. Ashdown.
Date: 8 April. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
Lecture: As above.

North Staffordshire Sub-Centre

Date: 11 April. Time: 7 p.m.
Held at: English Electric Co., Kidsgrove.
Lecture: Silicone Transistors—Their Manufacture and Fields of Application.
By: J. T. Kendall.
Date: 22 April. Time: 7 p.m.
Held at: Duncan Hall, Stone.
Joint Meeting with the I.P.O.E.E.
Lecture: The Application of Transistors to Line Communication Equipment.
By: H. T. Prior, D. J. R. Chapman and A. A. M. Whitehead.

South-East Scotland Sub-Centre

Date: 13 April. Time: 7 p.m.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Lecture: High-Current-Density Thermionic Emitters: A Survey
By: A. H. W. Beck.

Southern Centre

Date: 7 April. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Sound Reproduction.
By: D. E. L. Shorter.
Date: 20 April. Time: 6.15 p.m.
Held at: Farnborough Technical College, Boundary Road, Farnborough, Hants.
Lecture: Machine Translation of Languages.
By: A. D. Booth.

South Midland Electronics and Measurement Group

Date: 6 April. Time: 6.30 p.m.
Held at: Coventry Technical College.
Informal Lecture: High-Capacity S.H.F. Radio Transmission Systems.
By: H. D. Hyamson.

South-Western Sub-Centre

Date: 7 April. Time: 3 p.m.
Held at: Plymouth 'B' Generating Station, Prince Rock.
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R. Stuart and P. G. Kalman.

South-West Scotland Sub-Centre

Date: 6 April. Time: 6 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow C2.
Lecture: Silicone Electrical Insulation.
By: J. H. Davies.
Date: 12 April. Time: 6 p.m.
Held at: Royal College of Science and Technology, Glasgow.
Lecture: High-Current-Density Thermionic Emitters: A Survey
By: A. H. W. Beck.

Tees-Side Sub-Centre

Date: 6 April. Time: 6.30 p.m.
Held at: Cleveland Scientific and Technical Institution, Middlesbrough.
Annual General Meeting.
Lecture: The Characteristics and Protection of Semiconductor Rectifiers.
By: D. B. Corbyn and N. L. Potter.

Western Supply Group

Date: 4 April. Time: 6 p.m.
Held at: Large Lecture Theatre, Engineering Department, Bristol University, University Walk, Bristol 8.
Lecture: Some Principles Developed on the Application of Digital Computers to the Design of Electrical Plant.
By: B. J. Chalmers.

West Wales (Swansea) Sub-Centre

Date: 14 April. Time: 6 p.m.
Held at: Conference Room, S.W.E.B. Showrooms, The Kingsway, Swansea.
Lecture: Electronic Applications in a Modern Steelworks.
By: J. K. Edwards.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Data Processing Section

Date: 7 April. Time: 7 p.m.
Held at: Manson House, 26 Portland Place, London W.1.
Lecture: The Electronic Computer as a Unit in an Automatic Electronic Data Processing System for Missile Trials.
By: W. C. J. White and D. L. Overheu.

TELEVISION SOCIETY

All meetings will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London W.C.2, commencing at 7 p.m.
Date: 8 April.
Lecture: Video Tape in Action.
By: R. H. Hammans.
Date: 28 April.
Annual General Meeting

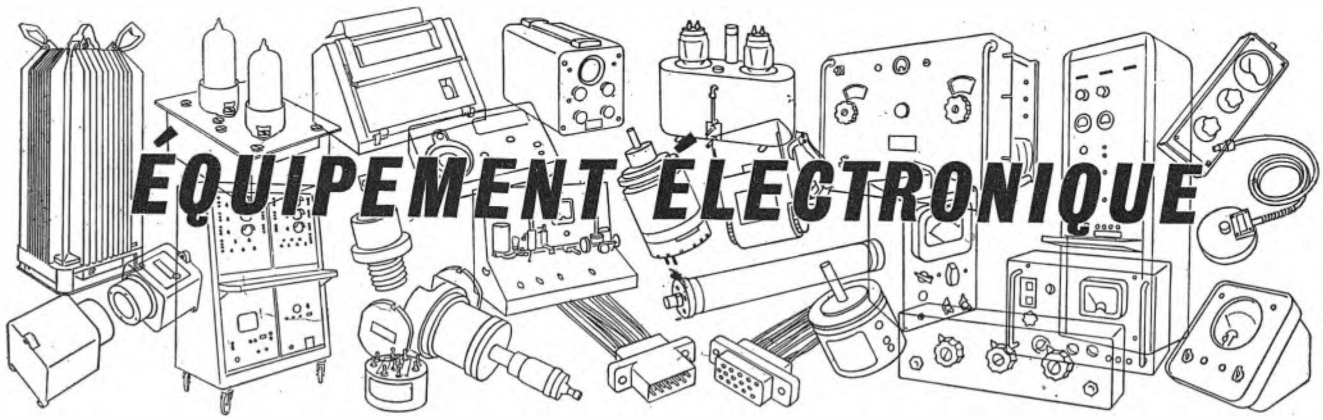
PUBLICATIONS RECEIVED

BASIC ELECTRONICS is the title of an illustrated training course in six parts, published by the Technical Press Ltd., 1 Justice Walk, London S.W.3, at a cost of 12s. 6d. per part or 66s. per set. Part 1 of this series deals with power supplies and rectifiers, Parts 2 and 3 with amplifiers and oscillators, Parts 4 and 5 with transmitters and receivers and Part 6 with frequency modulation-transistors. This course was developed for the United States Navy by the New York firm of management consultants and graphiological engineers Van Valkenburgh, Nooger and Neville, Inc and was adapted to British and Commonwealth usage by a Special Electronics Training Investigation Team of the Royal Electrical & Mechanical Engineers.

ZENITH 'VARIAC TRANSFORMERS, VAR-5 is a new catalogue which has been issued by The Zenith Electric Co. Ltd, Zenith Works, Villiers Road, Willesden Green, London N.W.2. While the layout of the previous catalogues has been retained, four pages have been added to include new models, the total number now exceeds 300.

SPEEDIVAC VACUUM COATING UNITS is a new 28-page catalogue issued by Edwards High Vacuum Ltd., Manor Royal, Crawley, Sussex, which is entirely devoted to plant suitable for such technical applications as lens blooming, front surface mirrors, interference filters, cathode-ray tubes, quartz crystals, selenium rectifiers and roll coating of tissue for paper capacitors.

EUROPEAN TECHNICAL DIGESTS is published monthly by the European Productivity Agency, 2 Rue André-Pascal, Paris. Every month over 1 000 technical journals appearing in thirteen different countries and in eleven different languages are searched by qualified engineers for information on new processes, methods, apparatus and materials which could be readily employed in industry without large capital investment. The purpose of the Digests is two-fold. Firstly, to spread from one country to another, irrespective of language barriers, technical information and knowledge likely to increase productivity and secondly to crossfeed ideas from one industry to another. Although these Digests also cover such sectors as glass, chemicals, leather, textiles and timber, they deal principally with all kinds of engineering and equipment, metallurgy and agricultural machinery. A number of countries publish translated versions of this publication so that, apart from the English edition, it now appears in German, Italian, Norwegian, Spanish, Turkish, Croat and Japanese and, on a selective basis, in Hebrew and Icelandic. Greece and the Netherlands produce translated title lists. Yearly subscription is £3 10s. 0d. ($12.00); six-monthly subscription is £1 15s. 0d. ($6.00).



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 252 à 255

BLOCS D'ALIMENTATION TRANSISTORISÉS

(Illustration à la page 252)

Metrix Instruments Ltd, 54 Victoria Road, Surbiton, Surrey

Cette société vient d'ajouter à sa production de blocs d'alimentation transistorisés de haute stabilité, une série étendue de modèles à coffret.

Six modèles couvrent la gamme 0-10V à 0,5A et 0-50V à 1A, 2A, 6A ou 10A. La tension de sortie, qui est "entièrement flottante", est ajustée exactement par une commande de réglage approximatif et précis. Deux appareils de lecture indiquent la tension de charge et le courant. Le taux de stabilité par rapport au secteur est supérieur à 2 000:1 et l'impédance de source est inférieure à 5mΩ. L'ondulation résiduelle de crête à crête est inférieure à 2mV pour le modèle 10A et inférieure à 0,5mV pour les modèles à caractéristiques de courant plus basses.

Les modèles 2A, 6A et 10A sont dotés de deux bornes de sorties supplémentaires de "perte de ligne". Lorsqu'on relie ces bornes directement à la charge même, la tension de charge effective, non pas la tension aux bornes de sortie, est indiquée sur l'instrument de lecture et stabilisée par l'amplificateur de commande. L'effet de perte dans les fils de connexion est donc complètement compensé.

Tous les modèles sont pleinement protégés contre les maxima et les pannes de composantes. Un déclencheur à microseconde polarise l'alimentation jusqu'à 0V en cas de surtension, et un relais augmente cette founction et débranche l'alimentation. Le coupe-circuit de surtension ne peut être réenclanché que lorsque le défaut est écarté. De plus, l'entrée de secteur et le circuit redresseur sont munis de fusibles.

EE 20 751 *pour plus amples renseignements*

OSCILLATEUR ULTRA-BASSES FRÉQUENCES

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

L'oscillateur ultra-basses fréquences Type 445 est à variation continue dans la gamme de fréquences 0,01 à 1kHz par accord de résistance-capacité. Les avantages avérés de cette forme d'accord s'allient ici à un nouveau circuit d'amplificateur sélectif à réaction positive commandée, pour maintenir l'oscillation continue.

L'amplificateur se compose d'un inverseur de phase, suivi de deux étages intégrateurs, et il est rendu sélectif par la réaction négative totale.

Le circuit de réaction positive comprend un limiteur à diode symétrique. La forme d'onde rectangulaire à amplitude constante émise par ce limiteur est appliquée à l'entrée de cet amplificateur. La haute sensibilité de l'amplificateur réduit le pourcentage de distorsion du signal de sortie à des proportions négligeables.

L'avantage du limiteur à diode, par rapport aux circuits de contrôle thermique communément employés, est qu'il assure une commande d'amplitude rapide jusqu'à la limite extrême de la fréquence. Après un signal transitoire soudain, tel qu'il s'en produit, par exemple, lorsqu'on change de gamme, l'amplitude d'oscillation est stabilisée à environ trois cycles près, à toutes les fréquences.

A la fréquence d'oscillation, chacun des deux intégrateurs produit un déphasage de près de 90°. Il en résulte donc une sortie en quadrature après le premier étage intégrateur, ce qui est particulièrement utile pour l'essai de servomécanismes. L'impédance de sortie de chaque étage intégrateur est très basse, de sorte que l'impédance de charge spécifiée peut être branchée directement sans affecter le rendement de l'oscillateur.

A la sortie de 0°, le niveau de sortie est à variation continue jusqu'à un maximum de 15V dans une charge minima de 10kΩ, cette sortie étant constante à ±4% près, dans toute la gamme de fréquences. Le taux de distorsion est inféricuer à 0,5% au total, dans la gamme mesurée de 20Hz à 1kHz. Le bourdonnement et le bruit sont inférieurs à 0,1%.

La sortie de 90° a une précision de phase de 90° ± 0,5° (décalé) par rapport à la sortie de °0 mesurée à 400Hz. Le niveau de sortie maxima à 90° est à ±3% prés de la sortie maxima 0°. Le taux de distorsion est inférieur à 1,5% au total, dans la gamme mesurée de 20Hz à 1kHz. Le bourdonnement et le bruit sont, une fois encore, inférieurs à 0,1%.

Cet appareil dérive d'un dessin de circuit breveté, créé par le Royal Radar Establishement.

EE 20 752 *pour plus amples renseignements*

FOIRE INTERNATIONALE DE HANOVRE

" Electronic Engineering" occupera le Stand 2, dans la Salle 13, à la Foire Internationale de Hanovre, du 24 avril au 3 mai 1960, et les visiteurs y seront les bienvenus.

ANALYSEUR D'ONDES POUR HYPERFRÉQUENCES

(Illustration à la page 252)

Airmec Ltd, High Wycombe, Buckinghamshire

L'analyseur d'ondes type 248 est un instrument de mesure sélectif destiné à une variété étendue de mesures quantitatives dans la gamme de fréquences de 5 à 300MHz. Il est tout particulièrement indiqué pour l'évaluation des harmoniques et composantes de formes d'ondes composées, mais il peut également être utilisé avec fruit pour les mesures d'intensité de champ et d'interférence, l'indication de gain et de perte par insertion, ainsi que comme détecteur en pont ou ondemètre hétérodyne.

La lecture directe de l'amplitude du signal peut être obtenue en se référant aux réglages de deux atténuateurs à plots et à un enregistreur de panneau. On peut mesurer des harmoniques secondaires jusqu'à 55dB et des harmoniques plus élevés jusqu'à 70dB sous la fondamentale.

L'analyseur se compose essentiellement d'un changeur de fréquence, suivi d'un amplificateur de fréquence intermédiaire à haute stabilité, fonctionnant à une fréquence ultra-sonore et incorporant une série d'atténuateurs à plots étalonnés. Ces derniers sont exempts d'erreurs de fréquence, grâce à la basse fréquence intermédiaire. Un compteur, protégé contre les maxima, est fourni aux fins d'interpolation entre les étages, cependant que des atténuateurs h.f. extérieurs peuvent être mis dans la prise d'entrée, donnant une atténuation totale de 100dB. Les avantages additionnels de la basse fréquence intermédiaire sont l'élimination d'un second canal, l'absence d'intervalles dans la "couverture" de fréquences, en raison du fait que la fréquence intermédiaire se trouve comprise dans la gamme de fonctionnement, et l'excellente forme de la courbe de réponse dont les côtés sont raides et le point d'equilibre aigu, facilitant ainsi une localisation de fréquence précise.

On peut détecter la modulation de fréquence ou d'amplitude sur le signal, et le signal acoustique s'obtient sur un jack téléphonique. Une sortie de fréquence moyenne est également prévue pour le contrôle extérieur.

La fréquence du signal est indiquée sur une échelle à pellicule à lecture directe, actionnée par engrenage dans le rapport de 90 à 1 et ayant une précision d'étalonnage de 2%. Les effets du bruit sont réduits au minimum par l'emploi de trois largeurs de bande alternatives de fréquence moyenne. Enfin, l'instrument est stable par inhérence, n'étant donc pas affecté par les variations de tension d'alimentation.

EE 20 753 *pour plus amples renseignements*

APPAREIL PORTATIF DE MESURE DES ÉPAISSEURS À RAYONS β

(*Illustration à la page 253*)

Baldwin Industrial Controls, Lowfield Street, Dartford, Kent

La société Baldwin Industrial Controls vient d'ajouter à sa serie de jauges d'épaisseurs nucléaires un petit appareil portatif à alimentation secteur.

Cet appareil a été conçu afin de répondre aux besoins de certaines branches de l'industrie pour lesquelles le coût et les raffinements de l'appareil de mesure standard à rayons β ne conviendraient pas.

Il est étalonné en unités arbitraires, proportionnelles au poids par surface d'unité, et l'utilisateur peut facilement mettre un facteur en corrélation avec son propre produit.

Pour des raisons de simplicité, l'appareil ne renferme qu'une seule source radioactive et un détecteur. La sortie du détecteur est comparée à un potentiel de référence calibré et la différence de signal est amplifiée par un amplificateur c.c. de gain réduit, extrêmement stable.

Le signal de sortie est suffisamment puissant pour actionner tant l'instrument de mesure à cadre mobile incorporé qu'au besoin, un enregistreur extérieur supplémentaire.

Quoique l'appareil ait été simplifié pour répondre à une demande particulière, la qualité et la précision de ses mesures demeurent de premier ordre dans les limites imposées par l'économie.

Divers montages à têtes de mesure peuvent être livrés, selon l'application qu'on leur destine ou les exigences commerciales. De même, des variantes de sources radioactives sont offertes, dépendant de la gamme et de l'épaisseur ou du poids par surface d'unité du matériau devant être mesuré.

EE 20 754 *pour plus amples renseignements*

APPAREILS D'ENSEIGNEMENT

(*Illustration à la page 253*)

Associated Electrical Industries (Woolwich) Ltd, 155 Charing Cross Road, London W.C.2

Ces appareils, conçus par la susdite société en collaboration avec le Northampton Polytechnic, seront à la base d'une série étendue d'expériences de laboratoire devant être effectuées conjointement avec des cours de théorie de la radio et de l'électronique.

Il existe onze appareils actuellement pour servir à l'enseignement des connexions de base des lampes octales internationales et des lampes B7G et B9A, ainsi qu'au fonctionnement de l'amplificateur à couplage résistance-capacité, des oscillateurs à couplage résistance-capacité et à anode accordée, des réseaux passe-bas et passe-haut et de l'étage de sortie de puissance.

Ils sont prévus pour le fonctionnement dans la gamme basses fréquences, de sorte que les accessoires correspondants —alimentations, oscillateurs b.f. oscilloscope—seront obtenus sans difficulté. Les fabricants sont, en outre, en mesure de fournir des blocs d'alimentation appropriés, ainsi qu'un oscillateur b.f.

Le schéma du circuit de chacun des principaux appareils est imprimé sur le haut du châssis et lorsqu'une borne ou un commutateur est indiqué en un point quelconque du circuit théorique, c'est à ce point précisément qu'est montée la pièce elle-même. Afin de rendre l'enseignement aussi clair que possible, l'étudiant peut voir le câblage et les diverses composantes sous le châssis, à travers une base de Perspex et en comparer la disposition avec le schéma du circuit.

Chaque appareil est fourni avec des feuilles d'instructions, s'appliquant à un certain nombre d'expériences et à leurs principes fondamentaux. Ces expériences répondent aux besoins d'étudiants tant au niveau d'études élémentaires qu'au niveau supérieur.

Notre gravure montre l'amplificateur à couplage résistance-capacité type R.2330. C'est un instrument des plus adaptables, muni de commutateurs pour modifier les capacités de couplage, introduire la rétroaction ou mettre en circuit une charge anodique accordée.

EE 20 755 *pour plus amples renseignements*

GÉNÉRATEUR DE SIGNAUX B.F. À GAMME ÉTENDUE

(*Illustration à la page 253*)

Distributeurs pour le Royaume-Uni: R. H. Cole (Overseas) Ltd, 2 Caxton Street, London S.W.1

Ce générateur de signaux fabriqué par la société Siemens & Halske couvre la gamme étendue de fréquences allant de 10Hz à 1MHz en sept bandes commutées. Cette étendue de fréquences est réalisée en alliant un générateur inductif-capacitif à un générateur résistif-capacitif.

Une haute stabilité d'amplitude et de fréquence, ainsi qu'un pourcentage réduit de distorsion, sont maintenus dans toute la gamme et on peut faire varier la tension de sortie sinusoïdale, continûment et par étapes de 20μV à 20V.

EE 20 756 *pour plus amples renseignements*

AMPLIFICATEUR DE COURANT CONTINU À GAIN VARIABLE

(*Illustration à la page 253*)

Feedback Ltd, 1 Broadway, Crowborough, Sussex

L'amplificateur type AD1A se compose de deux amplificateurs en un seul ensemble. L'un des amplificateurs a une gamme de fonctionnement allant du courant continu à 100kHz ($\pm$3dB), avec un étage d'entrée équilibrée pour réduire les effets de dérive, et un gain à variation continue de +1 à +20 par réglage de réaction. L'autre amplificateur peut être branché soit pour donner un gain de −1 (du courant continu à 100kHz) pour l'inversion de phase, soit pour un gain de −150 (du courant continu à 40kHz), pour ajouter, intégrer ou produire une fonction en employant les réseaux externes. Les deux amplificateurs ont une impédance d'entrée élevée et une basse impédance de sortie, avec une sortie maximum de $\pm$50V, $\pm$2mA. Ils peuvent, tous deux, être utilisés indépendamment l'un de l'autre ou être reliés en série afin de donner un gain maximum de −3 000, pour produire des fonctions plus précises.

EE 20 757 *pour plus amples renseignements*

SCELLEMENT VERRE-MÉTAL

(*Illustration à la page 254*)

Wesley Coe (Cambridge) Ltd, Scotland Road, Cambridge

La société Wesley Coe Ltd. offre une gamme de scellements métalliques, soit en modèles courants, soit fabriqués sur commande.

Le scellement de compression peut être réalisé sous forme de simple ouverture pour boîtiers de diodes ou pour des usages analogues. C'est un scellement fort robuste et convenant au soudage par projection pour la production en grandes séries. Des scellements peuvent également être fournis sous forme de collecteurs à plusieurs fils ou tubulaires, pour relais, transistors et de nombreuses autres applications. Le scellement d'adaptation en Nilo K et verre approprié, peut être employé pour tous les genres de collecteurs où une fermeture hermétique est exigée. Ces derniers peuvent être faits sous forme de collecteurs à broche unique ou à plusieurs broches, particulièrement pour l'équipement de bord où l'économie de poids est une considération d'importance.

Les scellements sont livrés en divers finis, conformément à la demande des clients, soit, par exemple, sous revêtement étamé par immersion à chaud, sous revêtement d'étain électrolytique, de cuivre, d'argent et d'or.

Tous les scellements peuvent être munis d'un certificat d'homologation des services de contrôle du Ministère de l'Air.

EE 20 758 *pour plus amples renseignements*

AMPLIFICATEUR À LARGE BANDE

(Illustration à la page 254)

R. H. Minns Electronics, The Lower Mill, Kingston Road, Ewell, Surrey

L'amplificateur à large bande type VM30 assure une amplification périodique dans la gamme de fréquences de 1MHz à 30MHz.

Un certain nombre d'amplificateurs de ce genre a déjà été fourni à la B.B.C. pour ses émetteurs de radiodiffusion h.f. fonctionnant dans cette gamme de fréquences. Ils permettent, en effet, d'obtenir une sortie de 15W dans une impédance de 100Ω, pour des fréquences allant jusqu'à 26MHz, avec application de 1V à travers le circuit d'entrée de 100Ω. La sortie tombe à 10W, à 30MHz, pour la même entrée. Il existe des modèles pour différentes impédances d'entrée et de sortie, coaxiales ou équilibrées. L'amplification s'effectue en trois étapes, les deux dernières utilisant des lampes Mullard QQV06-40A.

Il existe également un amplificateur de distribution à large bande type VM40 qui permet d'alimenter jusqu'à trois amplificateurs à large bande type VM30 d'une seule source. Entre l'entrée et chacune des sorties de l'amplificateur de distribution, il y a un gain dans le rapport de 3 à 1.

EE 20 759 *pour plus amples renseignements*

GRIFFES À BERCEAU MINIATURE

(Illustration à la page 254)

Insuloid Manufacturing Co. Ltd, Sharston Works, Leestone Road, Wythenshawe, Manchester

Le câblage à griffes à berceau Insuloid est utilisé depuis de nombreuses années. L'emploi de la griffe miniature a maintenant été étendu au matériel électronique etc, car elle peut recevoir des épaisseurs de fils pouvant atteindre 0,95cm de diamètre.

Les composantes de la griffe à berceau miniature consistent de berceaux et d'attaches moulés en nylon, avec des griffes de fixation captives en néoprène moulé, souple mais solide.

Les berceaux ont une ouverture évidée dans la matrice pour permettre le montage à boulon ou à vis sur panneau ou base en n'importe quelle position déterminée d'une installation de câblage. On assujettit les tours de câblage au moyen de la griffe en néoprène et on les agrafe ensuite à la position requise dans la gorge du berceau.

Lorsqu'il n'est pas nécessaire de pincer les tours de fils sur un panneau ou une base, on emploiera l'attache.

Un perfectionnement supplémentaire de la griffe est le berceau type TCO qui comporte un raccord métallique fileté dans la matrice du berceau et autorise la préfabrication d'installations de câblage, permettant ainsi de fixer par la suite la griffe à berceau miniature par l'arrière ou l'extérieur du panneau ou de la cabine.

Le système a une gamme de températures allant à peu près de −60°C à 100°C et offre une parfaite sécurité du point de vue électrique et mécanique dans les conditions climatériques les plus extrêmes.

EE 20 760 *pour plus amples renseignements*

RELAIS RÉMANENTS

(Illustration à la page 254)

L. E. Simmonds Ltd, 5 Byron Road, Harrow, Middlesex

Le principe de ce relais est entièrement basé sur celui, fort connu, du type 300, des services postaux britanniques. Après avoir été amorcé, le relais demeurera verrouillé indéfiniment, et il ne pourra être remis à la position de non-fonctionnement que par une impulsion dans le sens opposé, c'est à dire, soit par une polarité inversée, en cas d'emploi d'un seul bobinage, soit en faisant passer un courant à travers un second enroulement sur le même bobinage.

La relais gardera son efficacité même dans des conditions extrêmes de chocs et de vibrations.

Des dispositifs de contact des plus variés sont prévus et ils peuvent être fournis selon les spécifications particulières des clients.

EE 20 761 *pour plus amples renseignements*

APPAREIL DE MESURE DE PUISSANCE B.F.

(Illustration à la page 254)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Ce nouvel enregistreur de puissance b.f., type TF1347, est un instrument précis et sensible pour les mesures de fréquence acoustique dans une gamme étendue d'impédances (2,5Ω à 20kΩ en onze plots). Sa caractéristique de fréquence est substantiellement plate de 50Hz à 20kHz et ses dix gammes de puissance, s'étendant de 10μW à 6W, de même que son enregistreur de 139,7mm, assurent une excellente discrimination.

Fonctionnellement, cet appareil peut être considéré comme un atténuateur variable dont l'impédance d'entrée est réglable par degrés et dont la sortie est appliquée à un voltmètre étalonné en fonction de la puissance d'entrée à l'atténuateur.

Le circuit de l'appareil consiste de deux parties assurant de choix de l'impédance, l'atténuation et l'indication respectivement. L'impédance est choisie au moyen d'un transformateur d'adaptation à prises, relié aux bornes d'entrée de l'instrument par un commutateur à onze directions; le modèle de ce transformateur contribue largement à la réponse soutenue de l'appareil type TF1347 dans une gamme de fréquences aussi étendue. Un atténuateur résistif à dix plots constitue le multiplicateur de gamme de puissance et présente une impédance constante à l'enroulement secondaire du transformateur d'adaptation. L'atténuateur est terminé par le circuit indicateur qui se compose d'un instrument de mesure à cadre mobile et d'un redresseur relié en pont.

Au moyen du commutateur multiplicateur et de l'étalonnage de l'instrument de mesure, la sortie d'un système à fréquence acoustique peut être lue directement en unités de puissance, ainsi qu'en décibels relatifs à 1mW. La précision (à 1kHz et 20°C) pour des puissances de 200μW et au dessus, est de ±5% de la lecture, de l'échelle totale à la moitié de l'échelle. Elle est de ±2,5% de la déviation totale, d'une moitié d'échelle à 1/10ème de l'échelle totale. Pour des puissances inférieures à 200μW, la précision est de ±10μW.

EE 20 762 *pour plus amples renseignements*

CONDENSATEURS VARIABLES À VIDE POUSSÉ

(Illustration à la page 255)

English Electric Valve Co. Ltd, Waterhouse Lane, Chelmsford, Essex

La English Electric Valve Co. Ltd a réalisé une série de condensateurs variables à vide poussé destinés à l'emploi dans des circuits de sortie d'émetteurs et à des applications similaires.

Les différents types de cette gamme entièrement nouvelle de produits se distinguent par leurs dimensions réduites, leurs caractéristiques élevées de tension et de courant, et leur pouvoir de résistance à des pannes de surtensions temporaires sans subir d'endommagement.

Dans la gamme spécifiée, la capacité de chaque type varie linéairement avec la rotation de l'axe du condensateur, et la précision de la construction garantit un degré d'exactitude élevé dans le changement de capacité produit par chaque tour. La gamme de capacités s'étend légèrement au dessus et au desous des limites spécifiées ci-après, mais la variation de capacité dans la zône au dessous de la limite inférieure n'est pas linéaire.

Un résumé des principales caractéristiques d'un nombre de types en cours de réalisation est donné ci-dessous:

EEV TYPE	GAMME DE CAPACITES LINEAIRES (pF)	TOURS D'AXE DANS LA GAMME DE CAPACITES LINEAIRES	TENSION HAUTE FREQUENCE DE CRETE MAXIMA (kV)	COURANT HAUTE FREQUENCE MAXIMUM (Amp. valeur efficace)
U30/15	5 à 30	10,4	15	10 (jusqu' à 30 MHz)
U50/15	8 à 50	10,4	15	15 („ 30 „)
U80/15	10 à 80	10,4	15	15 („ 30 „)
U200/10	15 à 200	16,0	15	20 („ 20 „)

EE 20 763 *pour plus amples renseignements*

AMPLIFICATEURS MAGNÉTIQUES

De Havilland Propellers Ltd, Hatfield, Hertfordshire

Un amplificateur à "thermocouple" magnétique pouvant recevoir des signaux d'entrée de très bas niveau, à gain élevé et dérive réduite, vient d'être mis au point par la susdite société. Ses caractéristiques sont les suivantes:

Niveau d'entrée:	10-0-10mV c.c.
Niveau de sortie:	10-0-10mA c.c.
Gain de tension:	1 000
Impédance d'entrée:	100Ω enroulement
Impédance de sortie:	1kΩ
Nombre d'entrées:	1-16
Réponse de fréquence:	Plate jusqu'à 20Hz.
Linéarité:	2% de la sortie maxima.
Dérive en fonction du temps (Température et tension d'alimentation constantes):	10^{-12}W à l'entrée.
Dérive en fonction de la température: (tension d'alimentation constante)	gain −0,15%/°C. Zéro 2×10^{-14}W/°C
Dérive en fonction du changement de tension d'alimentation (10%):	$1{,}27\times10^{-11}$W
Tension de batterie:	10V
Fréquence de service:	2,4kHz.
Dimensions approximatives:	203,2mm × 152,4 mm × 88,9mm
Poids approximatif:	814 grammes.

L'amplificateur, entièrement enroulé dans des noyaux toroïdaux de haute performance, se compose de deux étages push-pull en cascade avec une réaction négative totale. La réaction est magnétique plutôt qu'électrique, afin de maintenir l'isolement entre l'entrée et la sortie, de sorte que 1 à 16 entrées isolées peuvent être utilisées. Une réaction à taux magnétique est également employée de façon à garantir une bonne réponse transitoire. Une nouvelle technique d'enroulement, avec choix de composantes, a permis d'obtenir un gain élevé et une dérive réduite sans devoir recourir à des circuits modulateurs d'harmoniques secondaires. L'inverseur de transistor, actionné par batterie, fonctionne à une fréquence suffisamment élevée pour assurer une réponse de fréquence adéquate.

EE 20 764 *pour plus amples renseignements*

POTENTIOMÈTRE À PISTE MOULÉE

(Illustration à la page 255)

The Plessey Co. Ltd, Ilford, Essex

La gamme des potentiomètres à pistes moulées que produit la société Plessey s'est enrichie d'un nouveau modèle hermétiquement scellé, le XP5.

Prévu pour l'emploi dans une gamme de températures de −40°C à 70°C, le nouveau potentiomètre se conforme aux normes des Forces Armées Britanniques pour les pièces de radio. Il résiste aux chocs les plus violents, aux vibrations et aux conditions tropicales. En outre, il est entièrement à l'épreuve de l'humidité.

L'axe est scellé par des anneaux de néoprène, logés dans des canaux et lubrifiés à la graisse anti-gel, qui assurent une action aisée, exempte de crasse de roulement. Les bornes argentées sont reliées intégralement à l'élément de résistance et sortent d'isolateurs en céramique scellés. Une rondelle de néoprène, fixée sur le côté du montage, fournit le scellement de panneau essentiel à l'emploi dans un équipement scellé.

La gamme de résistances va de 500Ω à 2,5MΩ (linéaires ou logarithmiques). La tension de service maximum à travers la résistance est de 500V c.c., dans les limites des caractéristiques de puissance.

EE 20 765 *pour plus amples renseignements*

COMMUTATEUR À BOUTON-POUSSOIR

(Illustration à la page 255)

Seton Creaghe Engineering Ltd, G.W. Trading Estate, Park Royal Road, London N.W.10

Ce commutateur convient aux applications hautes fréquences et il a été étudié en vue de réduire au minimum l'entretien et les risques de panne.

Il peut être fabriqué avec n'importe quel nombre de boutons jusqu'à un maximum de 16, et les boutons peuvent être éclairés si nécessaire. Chaque bouton est pourvu d'un dispositif standard de contact comportant deux inverseurs bipolaires. On ne peut actionner qu'un seul bouton-poussoir à la fois et les contacts du bouton actionné doivent être déconnectés avant d'effectuer tout nouveau contact. Ce dispositif peut être modifié pour "mettre ou retirer" le courant et on peut aussi effectuer l'interconnexion électromécanique. Enfin, le commutateur peut être monté horizontalement ou verticalement.

Des essais ont été faits à un régime de contact de 50V 250mA c.c. non-inductif, et la résistance de contact est inférieure à 20mΩ pendant toute la durée du commutateur. La capacité entre les contacts est inférieure à 2pF entre ressorts et à 1 pF entre ensembles.

EE 20 766 *pour plus amples renseignements*

Résumés des Principaux Articles

Un dispositif de contrôle acoustique à fréquence balayée de haute intensité pour la gamme de 1 à 60 kHz par D. J. Birchall

Résumé de l'article aux pages 202 à 208

Cet article décrit le dispositif de contrôle pour une sirène à fréquence balayée de haute intensité, ainsi que l'instrumentation prévue pour mesurer les niveaux de pression sonore, et comprenant des préamplificateurs de faible bruit, des amplificateurs passe-bande sélectifs et des amplificateurs logarithmiques.

La sirène est logée dans une chambre sourde et l'ensemble de l'équipement est utilisé pour effectuer divers essais acoustiques.

Une des caractéristiques les plus intéressantes est la méthode d'utilisation des thermistors comme éléments de contrôle de fréquence, par laquelle la sirène et les amplificateurs sélectifs sont ajustés à une référence de fréquence. Les amplificateurs sélectifs sont ainsi accordés à la fréquence fondamentale du champ sonore et s'ajustent l'un à l'autre lorsque la référence de fréquence est balayée, fournissant de la sorte un système de contrôle à fréquence balayée automatique dans les canaux à bandes étroites.

La gamme de fréquences de l'équipement est de 1 à 60 kHz, parcourue en un seul balayage sans commutation de gamme, à une cadence maxima de 2 octaves/minute.

La sensibilité de seuil pour les mesures de son est d'environ 70 dB et la gamme dynamique des amplificateurs logarithmiques est de 60 dB.

Choix de pistes à action rapide pour tambour enregistreur magnétique par A. D. Booth

Résumé de l'article aux pages 209 à 211

L'auteur donne un bref aperçu des méthodes adoptées pour le choix de pistes de tambour enregistreur magnétique. Il discute ensuite de deux commutateurs de tête de transistors et décrit, pour terminer, un nouveau commutateur de tête utilisant un seul transistor symétrique et ayant l'avantage de ne pas "injecter" un signal transitoire important dans la tête et l'amplificateur au moment de la commutation.

Les effets de la teneur en argon sur les caractéristiques des tubes à décharge luminescente par F. A. Benson et P. M. Chalmers

Résumé de l'article aux pages 218 à 223

Des mesures ont été effectuées pour déterminer les effets des variations de teneur en argon dans des tubes stabilisateurs à décharge sur les tensions d'amorçage et de marche, les courbes voltage de marche/température, les dérives initiales et les caractéristiques d'impédance/fréquence et de bruit. Les premiers essais furent réalisés avec des tubes au néon et au hélium, ayant des cathodes au cérium avec teneur en argon variant de 0 à 3,5%, par étapes de 0,5%. L'étude des effets en question se limita par la suite aux tubes au néon dont la teneur en argon variait dans la gamme relativement étendue de 0,001 à 10%, ayant des anodes au molybdène et des cathodes et utilisant des techniques de fabrication à tube étalon de haute stabilité. Les résultats de ces travaux sont présentés et discutés dans le présent article.

La mesure du niveau de bruit de transistors par F. J. Hyde

Résumé de l'article aux pages 224 à 226

L'auteur examine la technique de la diode de bruit courante. Il présente des formules générales qui tiennent compte, en premier lieu, des propriétés transmissives du circuit de couplage entre le générateur à diode de bruit et le transistor et, en second lieu, de l'effet du bruit d'amplification subséquent. Ces formules contiennent des admittances et des rapports de tension aisément mesurables.

L'effet d'ondes porteuses secondaires sur la caractéristique dynamique des diodes paramétriques par I. Hefni

Résumé de l'article aux pages 226 à 227

Au cours de travaux sur des dispositifs paramétriques à diodes de capacité variable, une caractéristique V-I anormale a été observée dans le cas de la jonction pn. Des oscillogrammes montrent les effets de cette anomalie qui serait due, pense-t-on, à l'existence d'ondes porteuses secondaires.

Circuit pour la protection de l'alimentation stabilisée de transistors par H. Kemhadjian et A. F. Newell

Résumé de l'article aux pages 228 à 230

Les transistors à grande puissance sont utilisés de plus en plus fréquemment dans les circuits. Il est souvent nécessaire de les faire fonctionner à un degré proche de leurs indices admissibles maxima, de sorte qu'ils risquent d'être détruits par une surcharge subite du circuit. Il devient donc indispensable d'avoir un circuit de protection qui puisse couper l'alimentation stabilisée des transistors, plus rapidement que les relais les plus rapides existant actuellement.

Le circuit de protection de surcharge décrit dans cet article peut couper l'alimentation en moins de 50 µsec. De plus, le courant auquel le circuit de protection opère peut être réglé à des niveaux de courant beaucoup plus bas que la sortie totale de l'alimentation. Ainsi les circuits expérimentaux actionnés par l'alimentation stabilisée sont également protégés contre tout endommagement dû à des défauts à l'intérieur de ces circuits.

Analyse de l'étage à charge fractionée (Transistorisé) par H. Pfyffer

Résumé de l'article aux pages 231 à 235

La première partie de cet article traite de l'étage à charge fractionnée en tant qu'étage unique avec réaction négative locale. Les expressions de gain de courant et de tension sont également étudiées.

Dans la seconde partie, un amplificateur réactif à trois étages, utilisant un étage à charge fractionnée à la sortie, est analysé par rapport au gain total, au gain de cadre ouvert et à l'impédance de sortie. Des valeurs numériques classiques ont été introduites dans deux spécimens afin de donner une idée du comportement de l'étage à charge fractionnée.

Données sur les matériaux en ferrite utilisés comme noyaux à tige dans les inducteurs et les transformateurs solénoïdaux par A. J. Baden Fuller

Résumé de l'article aux pages 236 à 237

Des tableaux présentés par l'auteur permettent de comparer la perméabilité et les pertes de différentes ferrites dans la gamme de fréquences de 2 MHz à 80 MHz, lorsqu'elles sont utilisées comme noyaux à tiges pour inducteurs et transformateurs solénoïdaux.

Divers spécimens de tiges en ferrite ont été employés comme noyaux d'une bobine dont l'impédance a été mesurée. Les résultats sont présentés en fonction du niveau réduit de surtension et du facteur d'insertion du matériau.

Un indicateur de foyer de télévision par J. B. Potter

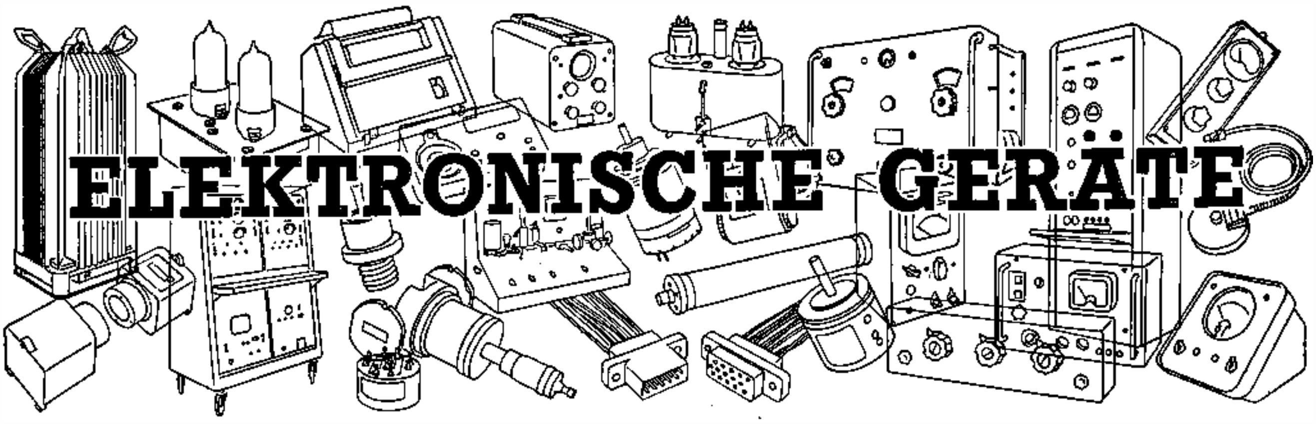
Résumé de l'article aux pages 240 à 241

Cet article décrit un indicateur transistorisé de foyer de télévision permettant d'appliquer un critère objectif à la focalisation électrique et optique de n'importe quelle source de signal d'image de télévision. Cet appareil peut être shunté dans un câble coaxial de 75 ohms et son fonctionnement dépend du fait que, lorsque l'équipement de conversion d'images est correctement focalisé du point de vue électrique et optique, le contenu haute fréquence du signal d'image est un maximum relatif.

Le circuit à charge cathodique plaque par C. H. Vincent et D. Kaine

Résumé de l'article aux pages 242 à 244

Un nouveau circuit simple, appelé circuit à charge cathodique plaque est décrit dans cet article. Inversant les impulsions négatives, il permet de transmettre les impulsions positives qui en résultent à une charge à faible impédance, avec un gain très proche de l'unité. L'élément renfermant ce circuit est également décrit. Il a été prévu pour la transmission d'impulsions allant jusqu'à 50V d'amplitude dans une charge de 100 ohms et il comporte un dispositif permettant de choisir par un interrupteur soit des entrées positives ou négatives, soit des sorties positives ou négatives.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 252 bis 255

TRANSISTORISIERTE NETZGERÄTE

(*Abbildung Seite 252*)

Metrix Instruments Ltd. 54 Victoria Road, Surbiton, Surrey

Das Fertigungsprogramm der Firma für hochstabile Transistor-Netzgeräte wurde durch eine umfangreiche Serie von Modellen in Gehäusen erweitert.

Sechs Modelle umfassen Geräte von 0...10V für 0,5A und 0...50V für 1 A, 2 A, 6 A oder 10 A. Die „völlig schwebende" Ausgangsspannung wird mittels Grob- und Feinregelung genau eingestellt, und zwei Messgeräte zeigen Spannung und Strom unter Belastung an. Das Stabilisierungsverhältnis ist besser als 2 000 : 1 und die Quellenimpedanz geringer als 5 m. Der doppelte Spitzenwert der Restwelligkeit ist unter 2 V für das 10 A-Modell und unter 0,5 mV für alle Modelle mit geringerem Strom-Nennwert.

Die Modelle für 2 A, 6 A und 10 A sind mit zwei zusätzlichen „Leitungsverlust" - Ausgangsklemmen versehen. Bei Verbindung der tatsächlichen Belastung zu diesen Klemmen wird die wahre Belastungsspannung und nicht die Spannung an den Ausgangsklemmen angezeigt und vom Steuerverstärker stabilisiert. Der Verlust in den Verbindungsleitungen wird daher völlig ausgeglichen.

Alle Modelle sind gegen Überlastung und Ausfall von Bauelementen geschützt. Im Falle eines Überlastungsstosses gibt eine Mikrosekunden-Triggervorrichtung eine Differenzspannung, die die abgegebene Spannung auf Null herunterbringt, und diese Aktion wird durch ein Relais verstärkt, das die Versorgung abschaltet. Die Schütze kann nicht rückgestellt werden, bis der Fehler behoben ist. Ausserdem sind Sicherungen im Netzeingang und Gleichrichterkreis vorgesehen.

EE 20 751 *für weitere Einzelheiten*

OSZILLATOR FÜR SEHR NIEDRIGE FREQUENZEN

Dawe Instruments Ltd. 99 Uxbridge Road, London W.5

Der Oszillator für sehr niedrige Frequenzen ist durch RC-Abstimmung über einen Bereich von 0,01 Hz ... 1 kHz kontinuierlich variabel. Die bereits feststehenden Vorteile dieser Abstimmart wirken zusammen mit einer neuartigen, selektiven Verstärkerschaltung mit gesteuerter positiver Rückkopplung, um ungedämpfte Schwingungen aufrecht zu erhalten.

Der Verstärker besteht aus einem Phasenumkehrer mit zwei folgenden integrierenden Stufen und wird durch eine alle Stufen umfassende Gegenkopplung selektiv gemacht.

In der positiven Rückkopplungsschaltung befindet sich ein symmetrischer Diodenbegrenzer. Die von diesem Begrenzer abgegebene Rechteckwellenform wird dem Verstärkereingang zugeführt. Durch die grosse Verstärkerempfindlichkeit wird der Anteil der im Ausgangssignal auftretenden Harmonischen auf ein unbeträchtliches Verhältnis herabgesetzt.

Gegenüber den herkömmlicherweise benutzten thermischen Reglerschaltungen hat der Diodenbegrenzer den Vorteil schnellster Amplitudenregelung bis zur untersten Frequenzgrenze herunter. Beim Auftreten plötzlicher Einschwingvorgänge, z.B. beim Bereichschalten, wird die Schwingungsamplitude für alle Frequenzen ungefähr innerhalb drei Perioden stabilisiert.

Für die Schwingungsfrequenz gibt jeder der beiden Integratoren eine Phasenverschiebung von nahezu 90°. Dadurch steht hinter der ersten Integratorstufe ein Vierungsausgang zur Verfügung, der besonders für Tests von Servosystemen nützlich ist. Die Ausgangsimpedanz jeder integrierenden Stufe ist sehr niedrig, so dass der vorgeschriebene Abschlusswiderstand direkt angeschlossen werden kann, ohne die Leistung des Oszillators zu beeinflussen.

Für den 0°-Ausgang ist der Ausgangspegel max. 15 V in eine Mindestbelastung von 10 kΩ stufenlos regelbar, wobei diese Ausgangsspannung über den ganzen Frequenzbereich innerhalb ±4% konstant ist. Der Klirrfaktor ist über den gemessenen Bereich von 20 Hz ... 1 kHz geringer als 0,5%. Brummen und Rauschen sind unter 0,1%.

Der 90°-Ausgang hat eine Phasengenauigkeit von 90° ±0,5° (nacheilend) in Bezug auf den bei 400 Hz gemessenen 0°-Ausgang. Für 90° ist der Ausgangshöchstpegel innerhalb ±3% des max. 0°-Ausgangs. Der Gesamtklirrfaktor für 90° ist für den gemessenen Bereich von 20 Hz ... 1 kHz geringer als 1,5%. Brummen und Rauschen sind unter 0,1%.

Das Gerät ist eine Weiterentwicklung einer vom Royal Radar Establishment stammenden patentierten Schaltung.

EE 20 752 *für weitere Einzelheiten*

UKW-ANALYSATOR

(*Abbildung Seite 252*)

Airmec Ltd. High Wycombe, Buckinghamshire

Der Analysator Type 248 ist ein selektives Messgerät, das für mannigfache quantitative Messungen im Bereich von 5 ... 300 MHz entwickelt wurde. Es ist besonders zur Bestimmung von Oberwellengehalten und Komponenten komplexer Wellenformen geeignet und kann auch zur Messung von Feldstärken und Störungen, Anzeige von Einfügungsgewinn und -verlust, sowie als Brückendetektor oder Überlagerungswellenmesser benutzt werden.

Direktablesung der Signalamplitude erfolgt durch Bezugnahme auf die Einstellung zweier Stufendämpfungsregler und ein Frontplatten-Anzeigegerät. Zweite Harmonische können bis zu 50 dB und höhere Harmonische bis zu 70 dB unter der Grundfrequenz gemessen werden.

Der Analysator besteht grundsätzlich aus einem Frequenzwandler mit nachfolgendem hochstabilen ZF-Verstärker, der auf Überschallfrequenz abgestimmt ist und eine Reihe geeichter Stufendämpfungsregler enthält, die auf Grund der niedrigen ZF frequenzfehlerfrei sind. Ein gegen Überlastung geschütztes Anzeigegerät interpoliert zwischen den Stufen; weitere HF-Dämpfungsglieder können in den Eingangssockel gesteckt werden, so dass eine Gesamtdämpfung von 100 dB erzielt werden kann. Die niedrige ZF bringt folgende weitere Vorteile mit sich: Wegfall der Spiegelfrequenz, keine Lücken im Frequenzbereich durch in den Arbeitsbereich fallende ZF, und die ausgezeichnete Kurvenform mit steilen Flanken und scharfem Nullpunkt erleichtern genaue Frequenzbestimmung.

AM- und FM-Signale können demoduliert und das NF-Signal über einen Telephonsockel abgenommen werden. Ein ZF-Ausgang ist für Aussenüberwachung vorgesehen.

Die Signalfrequenz wird direkt von einer Filmskala abgelesen, die über eine Übersetzung von 90 : 1 angetrieben wird und eine Eichgenauigkeit von 2% hat. Durch Verwendung von drei verschiedenen Bandweiten wird der Einfluss des Rauschens auf ein Minimum reduziert. Das Gerät selbst ist gegen Netzschwankungen unempfindlich.

EE 20 753 *für weitere Einzelheiten*

TRAGBARER BETADICKENMESSER

(*Abbildung Seite 253*)

Baldwin Industrial Controls, Lowfield Street, Dartford, Kent

Baldwin Industrial Controls hat vor kurzem ein kleines tragbares, netzgespeistes Gerät in das Fertigungsprogramm der Strahlungsdickenmesser aufgenommen.

Es wurde besonders für Industriegruppen entwickelt, für deren Bedarf die Kosten und Verfeinerungen des Standarddickenmessers nicht gerechtfertigt erscheinen.

Das Gerät ist in willkürlichen Einheiten proportional zum Gewicht der Flächeneinheit geeicht, und der Benutzer kann sehr einfach einen für sein besonderes Produkt gültigen Bezugsfaktor ermitteln.

Der Einfachheit halber hat das Instrument nur eine Strahlenquelle und Detektor. Der Detektorausgang wird mit einem geeichten Bezugspegel verglichen und die Signaldifferenz einem äusserst stabilen Gleichspannungsverstärker für geringe Verstärkung zugeführt.

Das Ausgangssignal ist stark genug, um sowohl das eingebaute Drehspulinstrument, als auch ein zusätzliches Aussenschreibgerät zu betreiben.

Trotzdem das Gerät für diesen besonderen Bedarf vereinfacht wurde, sind Qualität und Messgenauigkeit innerhalb der durch die Preislage gesetzten Grenzen ausgezeichnet.

Verschiedene Messkopffassungen entsprechend Einsatz und Fachanforderungen sind lieferbar, und andere Strahlenquellen können dem gewünschten Messbereich und Dicke oder Gewicht per Flächeneinheit gemäss angeboten werden.

EE 20 754 *für weitere Einzelheiten*

HILFSMITTEL FÜR DEN UNTERRICHT

(*Abbildung Seite 253*)

Associated Electrical Industries (Woolwich) Ltd, 155 Charing Cross Road, London W.C.2

Diese Geräte wurden in Zusammenarbeit mit dem Northampton Polytechnic entwickelt und bilden die Grundlage für eine umfangreiche Laborversuchsserie im Zusammenhang mit Kursen in Radio-und Elektroniktheorie.

Auf den elf lieferbaren Chassis können die grundsätzlichen Verbindungen für Röhren mit internationalen Oktal-, B7G- und B9A-Sockeln hergestellt und die Arbeitsweise eines RC-gekoppelten Verstärkers, eines Oszillators mit RC-Kopplung und Anodenabstimmung, Tief- und Hochpassnetzwerken, sowie einer Hochleistungsendstufe vorgeführt werden.

Die Geräte wurden für den NF-Bereich ausgeführt, so dass die notwendigen Zusatzgeräte wie z.B. Netzgeräte, NF-Oszillatoren und Oszillografen leicht beschafft werden können. Die Hersteller können auch geeignete Netzgeräte und einen NF-Oszillografen liefern.

Die Schaltung jedes Hauptbausteins ist auf der Chassisoberfläche aufgedruckt. Wo Klemmen oder Schalter in der theoretischen Schaltung vorgesehen sind, sind die Bauelement als solche montiert. Um den Unterricht klarer zu gestalten, können die Hörer Verdrahtung und Bauelemente unterhalb des Chassis durch eine Kunststoffscheibe beobachten und die Anordnung mit der Schaltung vergleichen.

Jeder Baustein wird mit Bedienungsvorschriften geliefert, die eine Versuchsreihe mit der zugrundeliegenden Theorie beschreiben. Die Versuche sind sowohl für Hörer im Elementarunterricht als auch in höheren Unterrichtsstufen bestimmt.

Die Abbildung zeigt einen RC-gekoppelten Verstärker Type R.2330, ein sehr vielseitiges Gerät, dessen eingebaute Schalter die Kopplungskapazität ändern, Gegenkopplung einführen oder einen abgestimmten Anodenkreis als Belastung einschalten können.

EE 20 755 *für weitere Einzelheiten*

BREITBEREICH-NF-MESSENDER

(*Abbildung Seite 253*)

Vertrieb in England durch R. C. Cole (Overseas) Ltd, 2 Caxton Street, London S.W.1

Dieser von Siemens und Halske hergestellte Messender bestreicht den breiten Frequenzbereich von 10 Hz...1 MHz in sieben Stufen. Der Frequenzbereich wird durch Verbindung eines RC- und eines LC-Generators erzielt.

Hohe Amplituden- und Frequenzstabilität sowie geringer Oberwellengehalt werden über den ganzen Bereich aufrechterhalten, und die sinusförmige Ausgangsspannung kann kontinuierlich und stufenweise von 20μV...20 V geregelt werden.

EE 20 756 *für weitere Einzelheiten*

GLEICHSTROM-REGELVERSTÄRKER

(*Abbildung Seite 253*)

Feedback Ltd, 1 Broadway, Crowborough, Sussex

Der Verstärker Type AD1A besteht aus zwei Verstärkern in einem Gehäuse. Der eine hat einen Arbeitsbereich von 0 . . . 100 kHz (±3 dB) mit ausgeglichener Eingangsstufe zur Herabsetzung des Drifteffekts und kontinuierlich variabler Verstärkung von +1 . . . +20 mittels Gegenkopplungsregelung. Der andere Verstärker kann entweder auf Verstärkung von −1 (0 . . . 100 kHz) für Phasenumkehr oder −150 (0 . . . 40 kHz) für Addition, Integration oder Funktionsgebung unter Verwendung von Aussennetzwerken geschaltet werden. Beide Verstärker haben hohe Eingangs- und niedrige Ausgangsimpedanz mit einer Höchstleistung von ±50 V, ±2 mA. Die beiden Verstärker können entweder unabhängig von einander oder seriengeschaltet für eine Höchstverstärkung von −3000 für genauere Funktionsgebung benutzt werden.

Der AD1A ist für versuchsmässige Untersuchungen von Systemfrequenzen und Einschwingvorgangskurven bestimmt, sowie als Baustein in Analogrechneranordnungen. Für viele dieser Verwendungszwecke wird nur mittelmässige Genauigkeit und Leistung verlangt, und der Hauptbedarf besteht für ein Gerät, das anpassungsfähig ist und eingebaute Regelvorrichtungen enthält.

Der Verstärker bildet einen Teil eines Fertigungsprogramms, das Übergangsfunktionsgeräte, einen Rechteckwellengeber, Servo-Test wellenformgeber und stabilisierte Netzgeräte umfasst.

EE 20 757 *für weitere Einzelheiten*

GLAS-METALL-VERSCHMELZUNG

(*Abbildung Seite 254*)

Wesley Coe (Cambridge) Ltd, Scotland Road, Cambridge

Wesley Coe Ltd liefert eine Reihe von Glas-Metall-Verschmelzungen in serienmässiger oder Sonderausführung.

Die Druckverschmelzung kann mit einer einzelnen Durchführung hergestellt werden und eignet sich dann besonders für Diodengehäuse und ähnliche Verwendung; es handelt sich um sehr stabile Ausführungen, die sich besonders für Massenherstellung mittels Warzenschweissen eignen. Sie können auch als Mehrfach-Durchführungen oder röhrenförmige Gehäuse für Relais, Transistoren und ähnliche Zwecke gefertigt werden. Für die angepasste Verschmelzung wird Nilo K und ein entsprechendes Glas benutzt, die besonders dort geeignet sind, wo sehr guter luftdichter Abschluss verlangt wird. Sie können als Gehäuse mit ein oder mehreren Durchführungen hergestellt werden und sparen Gewicht, was besonders für Flugzeug-Bordausrüstungen wichtig ist.

Kundenwünschen entsprechend können verschiedene Überzüge geliefert werden, z.B. feuerverzinnt, elektroverzinnt, verkupfert, versilbert oder vergoldet.

Auf Wunsch können Aufträge entsprechend den Prüfvorschriften für die Britische Luftwaffe (A.I.D.) geliefert werden.

EE 20 758 *für weitere Einzelheiten*

BREITBANDVERSTÄRKER

(Abbildung Seite 254)

R. H. Minns Electronics, The Lower Mill, Kingston Road, Ewell, Surrey

Der Breitbandverstärker Type VM30 gibt aperiodische Verstärkung im Bereich 1 MHz...30 MHz.

Eine Anzahl dieser Verstärker wurde an die British Broadcasting Corporation als Treiber für KW-Sender, die in diesem Bereich arbeiten geliefert. Für 1 V Spannung am 100Ω-Eingangskreis wird für Frequenzen bis 26 MHz eine Leistung von 15 W in eine 100Ω-Impedanz abgegeben. Ausführungen mit verschiedenen Eingangs- und Ausgangsimpedanzen, koaxial und ausgeglichen, sind lieferbar. Die beiden letzten der drei Verstärkerstufen sind mit Mullard-Röhren Type QQV06-40A bestückt.

Aus dem Fertigungsprogramm ist auch ein Breitband-Verteilungsverstärker Type VM40 lieferbar, der die Speisung von bis zu drei Breitbandverstärkern VM30 von einer Quelle ermöglicht. Die Spannungsverstärkung zwischen Eingang und jedem der Ausgänge des Verteilungsverstärkers ist 3 : 1.

EE 20 759 *für weitere Einzelheiten*

MINIATUR-VERDRAHTUNGSSCHELLE

(Abbildung Seite 254)

Insuloid Manufacturing Co. Ltd, Sharston Works, Leestone Road, Wythenshawe, Manchester

Das Insuloid "Cradleclip" Leitungs-Verlegungssystem ist seit einigen Jahren in Gebrauch; der Miniatur-Cradleclip wird jetzt für den elektronischen Gerätebau usw. eingeführt und kann Leitungen bis zu 9,5 mm ϕ aufnehmen.

Miniatur-Cradleclip-Teile bestehen aus Polyamid-Satteln und -Bindern mit anmontierten Befestigungs-Clips aus zähem, jedoch geschmeidig gepresstem Neopren.

Die Sattel haben im Ansatz ein vertieftes Loch, mittels dessen sie durch Schrauben in jeder gewünschten Position im Verdrahtungsplan auf Frontplatte oder Chassis montiert werden können. Der Neopren-Clip wird dann über die Verdrahtung geschwenkt und in die Sattelkehle eingeklemmt.

Wenn die Verdrahtung nicht auf Platte oder Chassis verankert werden soll, werden die Binder benutzt.

Eine weitere Verfeinerung ist die Metalleinlage mit Gewinde im Sattelansatz der Type TCo, die es ermöglicht, vorgefertigte Verdrahtungen herzustellen. Die Miniatur-Cradleclips können dann von der Rück- oder Aussenseite der Frontplatten oder Zellen später festgeschraubt werden.

Das System hat einen Temperaturbereich von −60°C...+100°C und gibt vollständige elektrische und mechanische Sicherheit selbst unter den ungünstigsten klimatischen Bedingungen.

EE 20 760 *für weitere Einzelheiten*

REMANENZRELAIS

(Abbildung Seite 254)

L. E. Simmonds Ltd, 5 Byron Road, Harrow, Middlesex

Dieses Relais stützt sich völlig auf das bekannte Fernsprechrelais 3000, ist jedoch mit einem Kern hoher Remanenz ausgerüstet. Nach Erregung wird das Relais für unbestimmte Zeit verriegelt und kann nur durch einen Impuls in entgegengesetzter Richtung in die Ruhestellung zurückgebracht werden, d.h. entweder durch umgekehrte Polarität für Relais mit einer Spule oder durch einen Strom durch eine zweite Wicklung derselben Spule.

Das Relais hält selbst bei erheblichen Stössen und Erschütterungen fest.

Vorhandene Kontaktanordnungen sind sehr umfangreich und können entsprechend Kundenbestellung geliefert werden.

EE 20 761 *für weitere Einzelheiten*

NF-LEISTUNGSMESSER

(Abbildung Seite 254)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Der neue NF-Leistungsmesser Type TF 1347 ist ein empfindliches und genaues Messgerät für NF-Messungen in einem breiten Impedanzbereich (2,5Ω... 20 kΩ in elf Stufen). Der Frequenzgang ist zwischen 50 Hz...20 kHz weitgehend flach; die 10 Leistungsbereiche von 10 µW...6 W und die 140 mm Skala ergeben ausgezeichnete Ableseunterscheidung.

Der TF 1347 kann in seiner Wirkungsweise als variables Dämpfungsglied betrachtet werden, dessen Eingang stufenweise regelbar ist und dessen Ausgang an einem Voltmeter liegt, das für den Leistungseingang zum Dämpfungsglied geeicht ist.

Die Schaltung des Gerätes besteht aus drei Teilen: Impedanzwahl, Dämpfung und Anzeige. Die Impedanzeinstellung erfolgt über einen Anpassungstransformer mit Abgriffen, der über einen elfstufigen Schalter mit den Eingangsklemmen des Gerätes verbunden ist. Die Konstruktion dieses Transformers trägt weitgehend zur flachen Kurve des Instruments über einen so breiten Frequenzbereich bei.

Ein zehnstufiges Widerstands-Dämpfungsglied bildet den Multiplikator für die Leistungsbereiche und stellt gegenüber der Sekundärwicklung des Anpassungstransformers eine konstante Impedanz dar. Das Dämpfungsglied wird durch den Anzeigekreis abgeschlossen, der aus einem Drehspulinstrument und Gleichrichter in Brückenschaltung besteht.

Mittels des Multiplikatorschalters und der Messgeräteichung kann der Ausgang eines NF-Systems direkt in Leistungseinheiten oder auch in dB auf 1 mW bezogen abgelesen werden. Für 1 kHz und 20°C ist die Genauigkeit für 200 µW und derüber ±5% der Ablesung zwischen Vollausschlag und halbem Ausschlag und ±2½% des Vollausschlages zwischen halbem Ausschlag und 1/10 Ausschlag. Für Leistungen unter 200 µW ist die Genauigkeit ±10 µW.

EE 20 762 *für weitere Einzelheiten*

VARIABLER HOCHVAKUUM-KONDENSATOR

(Abbildung Seite 255)

English Electric Valve Co. Ltd, Waterhouse Lane, Chelmsford, Essex

Die English Electric Valve Company Limited hat eine Reihe von variablen Hochvakuum-Kondensatoren für Senderschwingkreise und ähnliche Anwendungszwecke entwickelt.

Die Typen dieses vollständig neuen Programms zeichnen sich besonders durch kleine Abmessungen, hohe Spannungs- und Stromnennwerte und die Fähigkeit aus, Durchschlag durch vorübergehende Überlastung ohne Schaden auszuhalten.

Die Kapazität jeder Type ändert sich innerhalb des angegebenen Bereiches geradlinig mit jeder Umdrehung des Kondensatorschaftes. Die Präzisionsausführung garantiert, dass die Kapazitätsänderungen per Umdrehung mit grösster Genauigkeit erfolgen. Der Kapazitätsbereich geht etwas über die angegebene untere und obere Grenze heraus, Kapazitätsänderungen unterhalb der unteren Grenze erfolgen aber nicht linear.

Untenstehend folgt eine Zusammenstellung der Hauptkennzeichen einiger Typen, die entwickelt werden:—

EEV TYPE	LINEARER C-BEREICH (pF)	SCHAFTUMDREHUNGSZAHL IM C-BEREICH	MAX. SCHEITELSPANNUNG (kV)	MAX. HF-STROM (A_{eff})
U30/15	5.. 30	10,4	15	10 (bis zu 30 MHz)
U50/15	8.. 50	10,4	15	15 (bis zu 30 MHz)
U80/15	10.. 80	10,4	15	20 (bis zu 30 MHz)
U200/10	15..200	16	10	20 (bis zu 20 MHz)

EE 20 763 *für weitere Einzelheiten*

MAGNETISCHE VERSTÄRKER

De Havilland Propellers Ltd, Hatfield, Hertfordshire

Vor kurzem wurde ein magnetischer "Thermoelement"-Verstärker mit hoher Verstärkung und geringer Drift entwickelt, der sehr schwache Eingangssignale akzeptieren kann. Er hat folgende Merkmale:

Eingangspegel	10-0-10 mV Gleichspannung
Ausgangspegel	10-0-10 mA Gleichstrom
Spannungsverstärkung	1000
Eingangsimpedanz	100Ω/Wicklung
Ausgangsimpedanz	1 kΩ
Anzahl der Eingänge	1..16
Frequenzgang	flach bis zu 20 Hz
Geradlinigkeit	20% der Höchstleistung
Zeitdrift (Temperatur und Netzspannung konstant)	10^{-12}W der Eingangsleistung

Temperaturdrift (Netzspannung konstant)	Verstärkung — 0,15 %/°C Eingangspegel Null 2×10^{-14} W/°C
Drift mit Netzschwankungen (10%)	$1{,}27 \times 10^{-11}$ W
Batteriespannung	10 V
Abmessungen ungefähr	203 × 152 × 89 mm
Gewicht ungefähr	1,8 kg

Der durchgehend auf Hochleistungs-Ringkernen gewickelte Verstärker besteht aus zwei Gegentaktstufen in Kaskadenschaltung mit Gegenkopplung über alle Stufen. Um Isolation zwischen Ein- und Ausgang zu erreichen, wird magnetische statt elektrischer Gegenkopplung benutzt, und dadurch können 1...16 isolierte Eingänge vorgesehen werden. Gegenkopplung für magnetische Änderungen wird auch zur Erzielung einer guten Frequenzdurchlasskurve benutzt. Eine neue Wickeltechnik und Bauelementauslese führte zu hoher Verstärkung und niedriger Drift ohne Rückgriff auf Modulatorschaltungen für zweite Harmonische. Die Frequenz des batteriegespeisten Transistor-Zerhackers liegt hoch genug, um einen ausreichenden Frequenzgang zu garantieren.

EE 20 764 *für weitere Einzelheiten*

POTENTIOMETER MIT GEPRESSTER REGELBAHN

(*Abbildung Seite 255*)

The Plessey Co. Ltd, Ilford, Essex

Die Reihe der von der Plessey Company hergestellten Potentiometer mit gepresster Regelbahn wurde durch eine neue luftdicht verschlossene Type XP5 erweitert.

Das neue Potentiometer ist für Verwendung im Temperaturbereich −40° C ... +70° C bestimmt und entspricht den Waffenamtsnormen. Es kann schwere Prellung und Vibration sowie tropisches Klima vertragen.

Das XP5 ist völlig feuchtigkeitssicher. Der Schaft ist mit Neopren-Ringen abgedichtet, die in Nuten liegen und mit Frostschutzfett geschmiert sind, so dass sich gleichförmiges Betriebsverhalten ohne Lagerspiel ergibt. Die versilberten Lötfahnen sind mit dem Widerstandselement zusammengebaut und ragen durch abgedichtete keramische Isolatoren heraus. Die Befestigungsfläche ist mit einer Neopren-Scheibe ausgerüstet, die für dichte Geräte die notwendige Abdichtung an der Frontplatte herstellt.

Der Widerstandsbereich geht von 500 ....2,5 M (linear und logarithmisch).

EE 20 765 *für weitere Einzelheiten*

DRUCKKNOPFSCHALTER

(*Abbildung Seite 255*)

Seton Creaghe Engineering Ltd, G.W. Trading Estate, Park Royal Road, London N.W.10

Dieser Schalter ist für Hochfrequenzeinsatz geeignet und so konstruiert, dass Wartung und die Möglichkeit mechanischen Ausfalls verringert werden.

Der Schalter kann für jede Druckknopfzahl bis zu 16 hergestellt und auf Bestellung mit leuchtenden Knöpfen geliefert werden. In Normalausführung betätigt jeder Knopf zwei zweipolige Umkehrschalter. Nur ein Druckknopf kann jeweils bedient werden, und alle getätigten Kontakte werden freigegeben, bevor ein neuer Kontakt hergestellt wird. Diese Anordnung kann geändert werden, um für „Schliessen oder Unterbrechen" zu schalten. Elektromechanische Verriegelung ist auch lieferbar. Die Schalter können waagerecht oder senkrecht gruppiert werden.

Versuche wurden mit 50 V 250 mA induktionsfreier Gleichstrombelastung durchgeführt und ergaben während der ganzen Lebensdauer einen Kontaktwiderstand unter 20 mΩ. Die Kapazität der Kontakte ist unter 2 pF zwischen Federn und unter 1 pF zwischen Gruppen.

EE 20 766 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Eine Wobbelfrequenz-Schalltestanlage für den Bereich von 1 . . . 60 kHz

von D. J. Birchall

Zusammenfassung des Beitrages auf Seite 202-208

Der Beitrag beschreibt die Steuergeräte einer Wobbelfrequenzsirene für grosse Lautstärke und die zur Messung von Schalldruckpegeln entwickelten Messeinrichtungen, die aus rauscharmen Vorverstärkern, selektiven Bandfilterverstärkern und logarithmischen Verstärkern bestehen.

Die Sirene ist in einem schalltoten Raum untergebracht, und die ganze Anlage wird zur Ausführung der verschiedensten akustischen Untersuchungen benutzt.

Das interessanteste Merkmal dieser Methode ist die Verwendung von Thermistoren als Frequenzsteuerungselemente, durch die die Sirene und die selektiven Verstärker mit einer Bezugsfrequenz spuren. Dabei werden die selektiven Verstärker auf die Grundfrequenz des Schallfeldes abgestimmt und spuren mit der Bezugsfrequenz während des Durchlaufs, so dass dadurch ein System für automatisch gewobbelte Überwachung von Engbandkanälen entsteht.

Der Frequenzbereich des Gerätes von 1 . . . 60 kHz wird in einem Durchlauf ohne Schaltung mit einer Höchstgeschwindigkeit von zwei Oktaven/Minute bestrichen.

Die Schwellensensitivität für Schallmessungen liegt bei 70 dB und der dynamische Bereich des logarithmischen Verstärkers ist 60 dB.

Schnellzugriff für Magnettrommelspeicherspur

von A. D. Booth

Zusammenfassung des Beitrages auf Seite 209-210

Der Beitrag gibt eine kurze Übersicht der Methoden, die für den Spurenzugriff bei Magnettrommelspeichern angewandt werden. Danach werden zwei übliche Transistor-Kopfschalter besprochen und schliesslich ein neuer Kopfschalter beschrieben, der einen einzigen symmetrischen Transistor benutzt. Diese Anordnung hat den Vorteil, dass im Schaltmoment keine grosse Einschwingung in den Kopf und Verstärker gespeist wird.

Der Einfluss von Argon auf Glimmröhren-Kennlinien von F. A. Benson und P. M. Chalmers

Zusammenfassung des Beitrages auf Seite 218-283

Messungen zur Bestimmung des Einflusses verschiedener Argonquantitäten in Stabilisator-Glimmröhren auf Zünd- und Brennspannungen, Brennspannung-Temperaturkurven, anfängliche Drift und Impedanz-Frequenz sowie Rauschkennlinien wurden ausgeführt. Ursprünglich wurden neon- und heliumgefüllte Röhren benutzt, deren Zerkathoden 0 . . . 3,5% Argon in 0,5% Stufen enthielten. Spätere Untersuchungen wurden auf neongefüllte Röhren mit einem Argongehalt über den verhältnismässig weiten Bereich von 0,001 . . . 10%, Molybdänanoden und -kathoden beschränkt, deren Herstellung nach der Technik für hochstabile Bezugsröhren erfolgt. Untersuchungsresultate werden beschrieben und diskutiert.

Messen der Transistor-Rauschzahl von F. J. Hyde

Zusammenfassung des Beitrages auf Seite 224-225

Die Standard-Rauschdiodentechnik wird betrachtet. Allgemeine Formeln werden angegeben, die (1) die Übertragungseigenschaften der Koppelkreise zwischen der Rauschdiode und dem Transistor und (2) die Wirkung des folgenden Verstärkerrauschens in Betracht ziehen. Diese Formeln enthalten leicht messbare Leitwerte und Spannungsverhältnisse.

Der Einfluss von Nebenträgern auf die dynamischen Kennlinien parametrischer Dioden von I. Hefni

Zusammenfassung des Beitrages auf Seite 226-227

Im Laufe von Versuchsarbeiten an parametrischen Vorrichtungen unter Benutzung von Dioden mit veränderlicher Kapazität wurden unregelmässige U–I Kennlinien für pn Flächen beobachtet. Der Einfluss dieser Beobachtung wird mittels Oszillografspuren gezeigt, und es wird angedeutet, dass er auf das Vorhandensein von Nebenträgern zurückzuführen ist.

Schutzschaltung für einen stabilisierten Transistor-Netzteil von H. Kemhadjian und A. J. Newell

Zusammenfassung des Beitrages auf Seite 228-230

Leistungstransistoren werden immer häufiger in Schaltungen verwandt. Sie müssen oft nahe ihren Betriebshöchstwerten arbeiten und können daher schnell durch plötzliche Überlastung zerstreut werden. Daher besteht Bedarf für eine Schutzschaltung, die den stabilisierten Transistor-Netzteil schneller abschaltet als das schnellste bekannte Relais.

Die im Beitrag beschriebene Schutzschaltung kann den Netzteil in unter 50 μs abschalten. Ausserdem kann der den Schutzkreis betätigende Strom viel niedriger sein als der gesamte vom Netzteil abgegebene. Daher können von stabilisierten Netzteilen gespeiste Versuchsschaltungen gegen Beschädigung durch in diesen Schaltungen auftretende Fehler geschützt werden.

Analyse der transistorisierten Stufe mit geteilter Belastung von H. Pfyffer

Zusammenfassung des Beitrages auf Seite 231-235

Im ersten Teil dieses Beitrags wird die Stufe mit geteilter Belastung als einzelne Stufe mit eigener Gegenkopplung untersucht. Ausdrücke für Strom- und Spannungsverstärkung werden auch in Betracht gezogen.

Im zweiten Teil wird ein dreistufiger Verstärker mit Gegenkopplung und Endstufe mit geteilter Belastung in Bezug auf Gesamtverstärkung, Verstärkung der offenen Schleife und Ausgangsimpedanz analysiert. In zwei Beispielen werden typische Zahlenwerte eingesetzt, um einen Begriff für das Verhalten der Stufe mit geteilter Belastung zu geben.

Daten für Ferrit-Werkstoffe, als Stabkerne in Magnetspulen und Ubertragern verwandt von A. J. Baden Fuller

Zusammenfassung des Beitrages auf Seite 236-237

Aufgeführte Tabellen ermöglichen Permeabilitäts- und Verlustvergleiche für verschiedene Ferrite bei Verwendung als Stabkerne in Magnetspulen und Übertragern im Bereich von 2 MHz . . . 80 MHz. Verschiedene Ferrit-Stabprüflinge wurden als Kern einer Spule benutzt und deren Impedanz gemessen. Die Resultate werden als „Ruhestrom-Q" und „Einfügefaktor" des Werkstoffes gegeben.

Ein Fernsehfokus-Anzeigegerät von J. B. Potter

Zusammenfassung des Beitrages auf Seite 240-241

Der Beitrag beschreibt einen transistorisierten Fernsehfokusanzeiger, der eine objektive Beurteilung der elektrischen und optischen Fokussierung jeder Fernsehbild-Signalquelle ermöglicht. Das Gerät kann in 75 Ω koaxiales Kabel eingeschaltet werden. Die Arbeitsweise stützt sich auf die Tatsache, dass der Hochfrequenzgehalt des Bildsignals bei korrektem elektrischen und optischen Fokus der Bildwandleranordnung ein relatives Maximum hat.

Der Anoden-Kathodenverstärker von C. H. Vincent und D. Kaine

Zusammenfassung des Beitrages auf Seite 241-244

Eine neue, einfache, als Anoden-Kathodenverstärker bezeichnete Schaltung wird beschrieben, in der negative Impulse umgekehrt werden. Die entstehenden positiven Impulse werden in einen Widerstand geringer Impedanz mit einem Verstärkungsfaktor gespeist, der sehr nahe bei 1 liegt. Ein Gerät nach dieser Schaltung wird beschrieben. Es wurde entwickelt, um Impulse mit Amplituden bis zu 50 V in einen 100 Ω Abschlusswiderstand zu übertragen und ermöglicht durch einen eingebauten Schalter die Wahl positiver oder negativer Eingänge und positiver oder negativer Ausgänge.