ELECTRONIC ENGINEERING

VOL. 39 No. 468

FEBRUARY 1967

Commentary

THE Council of Engineering Institutions (C.E.I.) came in for considerable criticism from Professor Emrys Williams when he read his Presidential Address to the Institution of Electronic and Radio Engineers last December. To be more precise it was one aspect of C.E.I.'s work which came under fire: its endeavours to promote the 'complete engineer' and to set itself up as an all-embracing examining body.

Briefly, and to quote Professor Williams, "The majority opinion within C.E.I. is that the essential distinguishing feature of a professional engineer is that he is a man with such education, training and experience as will enable him to see a major engineering project through to completion. He is the whole engineer, the complete engineer, the man who can see the whole picture. It follows that those who hold this opinion believe that there can indeed be such a person and that there is a definable body of engineering science and knowledge which is a requirement common to all branches of the engineering profession." Further, it seems that the C.E.I. is intent on setting itself up as an examining body with an examination syllabus firmly based on the concept of the 'complete engineer' and when this is accomplished the individual institutions will be expected to discontinue their own Graduateship examinations.

Since these are majority decisions it must be presumed that they are acceptable to most of the thirteen constituent institutions; they are, however, profoundly unacceptable to the minority, and this, of course, includes the Institution of Electronic and Radio Engineers.

To bring forth the idea of the 'complete engineer' at this time is rather strange and represents a complete reversal of modern thinking on the education of engineers: indeed, it seems to be an attempt to put the clock back thirty years or more. Undoubtedly, there was a time when a man could be designated simply as an engineer and when he could easily turn his mind to any engineering project which was at that time possible. In fact, it was not so long ago when one man could reasonably hope to embrace the whole gamut of 'engineering knowledge' as it then existed. Those days have gone and, to quote Professor Williams, "The era of the complete engineer able to turn his mind to any aspect of engineering saw its close thirty years ago. The last strains of the lament for the general practitioner were drowned in the tumult of the 1939-45 war. Now it is difficult to be a general practitioner even in the limited field of Electronic Engineering, so wide has that field become." There are certainly many electronic engineers who will agree most heartily with that last statement!

The pattern of engineering education throughout the last fifty years or so has been one of increasing specialization. In the universities, Departments of Engineering have been split into Departments of Civil, Mechanical and Electrical Engineering and, more latterly, Electronic Engineering. In some universities these departments are already being split into even more specialized units. The explosive growth in the knowledge of electronic engineering and the necessity for increasing emphasis on pure science and mathematics for students of electrical and electronic engineering has inevitably led to a decrease of emphasis on other subjects, such as the mechanical aspects of design while, in varying degree, subjects such as fluid-mechanics, stress-analysis and applied thermodynamics are ceasing to be thought essential parts of electrical engineering—and even less of electronic engineering. The amount of time available for learning and a man's capacity for learning are both strictly limited and nothing can change that. As mankind's fund of knowledge increases, the fraction of the whole which any one man can assimilate must necessarily decrease. Put another way, a man can acquire a superficial knowledge of a wide range of subjects or an intimate knowledge of a narrow range and, until the C.E.I. made its proposals, it would have been thought that it is a universally accepted truth that 'this is an age of specialization'.

So far as examinations go, it is rather difficult to see why the C.E.I. should find it necessary to set itself up as an examining body at all. Several of the constituent institutions, including the I.E.E. and the I.E.R.E. have conducted their own examinations for a number of years and it would seem that the C.E.I. would have been better employed as a certifying body with the task of ensuring that the various institutions' examinations did not fall below a required minimum standard. After all, the institutions have evolved their present examination syllabuses after much careful thought and patient investigation into the most suitable types and scope of education for engineers in their respective fields. Indeed, it was largely the inadequacy of the type of syllabus which the C.E.I. is now proposing which led the specialist institutions to institute their own Graduateship examinations.

There are, of course, many other aspects of the work of the C.E.I., such as encouraging the cross-fertilization of knowledge and enabling the engineering profession as a whole to speak in unison on matters of common interest both at home and overseas and, in these matters, it has the enthusiastic support of all the constituent institutions. It would, though, seem that if the affairs on the C.E.I. are to prosper, then some very clear rethinking and frank discussion must take place on the very vital subjects of education, examination syllabuses and qualifications and surely our professional institutions are mature enough to resolve these difficulties.

V.H.F. Solid-State Power Amplifier Design

By T. Spencer*, M.Sc.(Eng.)

A design technique for class-C power amplifiers based on standard data published by device manufacturers is given. The analysis for transistor class-C operation is identical to that used for valves and normalized design curves are given. Using a step-by-step procedure design calculations are shown to compare favourably with practical results.

Several power amplifier and multiplier circuits are illustrated, together with those of free-running and crystal controlled oscillator drivers. Factors influencing efficiency and output power are discussed.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

OUTPUT powers comparable with those obtainable from low-power klystrons are now available at X-band frequencies, using solid-state drivers and multipliers together if necessary with multipliers using coaxial line and waveguide techniques. The light weight and compactness of the solid-state source as a portable unit operating from dry batteries at a competitive price is an important feature.

Basic Design Considerations

Transistor class-B or class-C power amplifiers are restricted by V_T, the collector turn-over voltage. At v.h.f. the common-base configuration has the larger power-gain, but $V_T = V_P$, the punch-through voltage is a criterion.

The reverse bias, together with the peak input drive voltage, is limited to V_{ebo}. With self-bias, the removal of the input signal would result in a condition of high collector current, and an instantaneous collector voltage exceeding V_{cc}, if the base-emitter diode is even slightly forward biased, thereby destroying the transistor.

Earthing the output tuning capacitor increases the insertion loss through increased strays, with decreased output power. $Q_o = 1/R\sqrt{(L/C)}$ decreases while $Q_w = R_L\,\omega C$ increases with strays, increasing the insertion loss. Because the harmonic content of the output varies inversely with Q_w, a compromise is necessary.

Since P_{cm}, the maximum collector dissipation occurs at zero frequency, the criterion for thermal stability is satisfied if V_{cm}, the maximum collector voltage at room temperature is[4]:

$$V_{cm} = \frac{1 - \lambda\,|V_{cm}/V_{cbo}|^n}{\theta\,\lambda\,\phi_c I_{co}}$$

when both avalanche-multiplication and thermal effects are considered. $\phi_c = 0{\cdot}07$, $n = 3$ for germanium; θ, I_{co} and V_{cbo} are obtained from manufacturers' data. The stability factor λ is:

$$\lambda = \frac{(1 + h_{FE})}{1 + (h_{FE}r_e/r_b)} = dI_c/dI_{co} \geqslant 1$$

The common-base configuration permits large peak output voltages, provided $r_e/r_b \geqslant 1$, where r_e and r_b include external resistance. A minimum temperature stability factor close to unity is also assured.

The instantaneous collector current $f(\phi)$ may be expressed as a Fourier expansion:

$$f(\phi) = a_o + \sum_{n=1}^{\infty} (a_n \cos n\,\phi + b_n \sin n\,\phi)$$

The fundamental has the same period as $f(\phi)$ where, for class-B operation:

$$f(\phi) = I_c = I_{cm}/\pi\,(1 + (\pi/2)\sin\phi - (2/3)\cos 2\phi - (2/15)\cos 4\phi + \ldots)$$

With the tank tuned to the fundamental:

$$(I_{c2}/I_{c1})^2\,\frac{R(Z_2)}{Z_{\omega 1}} = (I_{cn}/I_{c1})^2\,\frac{1}{n^2 + Q_w^2(n^2-1)^2} = 16/9\pi^2 \cdot \frac{1}{4 + 9Q_w^2}$$

where Z_2 is the tuned circuit impedance to the second harmonic, $Z_{\omega 1}$, is the dynamic impedance at the fundamental, and I_{cn} the harmonic component of collector current. For a low Q_w the harmonic rejection is still considerable.

For the second harmonic, with the driver frequency equal to the fundamental, the relative output power is:

$$(I_{c2}/I_{c1})^2\,\frac{Z_{\omega 2}}{R(Z_1)} = (I_{c2}/I_{c1})^2 = 16/9\pi^2 = -7{\cdot}5\text{dB}$$

if the second harmonic dynamic impedance $Z_{\omega 2}$, after retuning, equalled the impedance of the tank to the fundamental Z_1. The power amplifier is therefore a multiplier.

The device input characteristic is:

$$I_b = I_s \sum_{n=0}^{\infty} \gamma^n/n!\ \ ;\ \ \gamma = qV/KT \simeq 40$$

where I_b is the base current due to an external voltage V and I_s the reverse leakage current. Both even and odd harmonic multipliers are therefore possible for class-C operation. Other factors contributing to distortion are hole storage and phase-shift due to base spreading.

Design Parameters

In addition to the parameters required to ensure thermal stability and prevent punch-through, the gain-bandwidth product f_T is given in the data sheets. This parameter, also designated f_1, is the gain-bandwidth product when h_{fe} is unity. As power gain is possible at frequencies approaching f_T, it is assured with the base common since $f_\alpha = Kf_T$, K varying between 1·2 and 2. K depends on the amount of drift field; for alloy transistors $K = 1{\cdot}2$ while for diffused-base mesa transistors, $K = 1{\cdot}9$.

For small signal operation the maximum power gain is:

$$\text{M.P.G.} \simeq 1/4\ \frac{\omega_1}{\omega^2 C_{b'c} r_{bb'}}$$

$$\text{Since } f_m = \sqrt{\left[\frac{f_1}{8\pi C_{b'c} r_{bb'}}\right]} :- \quad \omega_1 = 4\omega_m^2 C_{b'c} r_{bb'}$$

$$\text{M.P.G.} = (f_m/f)^2 = 20\log(f_m/f)$$

If $C_{b'c}$ and $r_{bb'}$ are known at f, f_m is known, therefore the power gain.

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The power gain at any frequency can be approximated, if the slope with frequency for a given device, is known. Since a slope less than 3dB/octave is unlikely, the minimum power gain will be:

$$G = a\left(1 - (b/a)(f/f_T)\right) \text{ where } (a-b) = 1$$

For a 3dB/octave slope: $a = 3$, $b = 2$

For a 6dB/octave slope: $a = 7$, $b = 6$

For a 10dB/octave slope: $a = 19$, $b = 18$

A third method of device selection is based on an analysis of class-C amplification using transistors. The method of design given in the Appendix is best illustrated by an example.

Class-C Amplifier Design

For the Fairchild 2N2368 transistor:

At 25°C: $P_{cm} = 0{\cdot}36\text{W}$ (in air)

$V_{cbo} = 40\text{V}$

$V_{ceo} = 15\text{V}$

$I_{cm} = 100\text{mA}$

$f_{T(min)} = 400\text{MHz}$

$V_{ebo} = 4{\cdot}5\text{V}$

$T_{jm} = 200°\text{C}$

Requirements: $P_o = 0{\cdot}5\text{W}$ into 50Ω

$f = 250\text{MHz}$

Using a 'Redpoint' type 18DC heatsink:

Thermal resistance $\simeq 0{\cdot}2°\text{C/mW}$

$$P_{cm} = 0{\cdot}875\text{W}$$

$$V_{cc} = \sqrt{(2P_oR_L)} = \sqrt{50} = 7{\cdot}06\text{V}$$

$$I_{c1m} = 2P_o/V_{cc} = \frac{1\,000}{7{\cdot}06} = 141{\cdot}6\text{mA}.$$

Assume a collector efficiency $\eta = 84$ per cent.

From Fig. 3: $\theta = 150°$

From Fig. 1: $\gamma = 0{\cdot}455 \rightarrow I_{cm} = \dfrac{141{\cdot}6}{0{\cdot}455} = 311\text{mA}$

$$I_o = 1/2\,(I_{c1m}/\eta) = \frac{70{\cdot}8}{0{\cdot}84} = 84{\cdot}3\text{mA}$$

Check: From Fig. 2: $\beta = 0{\cdot}269$

$$I_o = \beta I_{cm} = 0{\cdot}269 \times 311 = 83{\cdot}7\text{mA}.$$

$$P_{cm}' = P_o\left((1/\eta) - 1\right) = 0{\cdot}19\,P_o = 0{\cdot}095\text{W}$$

$$P_{cm} = P_o + P_{cm}' = 0{\cdot}595\text{W}$$

$$V_{cc} = P_{cm}/I_o = \frac{0{\cdot}595 \times 10^3}{84{\cdot}3} = 7{\cdot}06\text{V}.$$

From the $h_{FE}-I_c$ characteristic of the 2N2368 transistor:

$h_{FE} = 21$ ($I_{cm} = 311\text{mA}$).

$$\alpha_{FB} = \frac{h_{FE}}{(1 + h_{FE})} = 21/22$$

$$f_{\alpha(min)} = f_T/\alpha_{FB} = 22/21 \times 400 = 419\text{MHz}$$

$$\frac{f}{f_{\alpha(min)}} = 250/419 = 0{\cdot}596$$

$$|\alpha/\alpha_{FB}| = 1/\sqrt{[1 + (f/f_\alpha)^2]} = 0{\cdot}855$$

$$|\alpha| = \frac{0{\cdot}855 \times 21}{22} = 0{\cdot}82$$

Peak input current $I_{em} = I_{cm}/|\alpha| = 311/0{\cdot}82 = 379\text{mA}.$

Assume $R_{IN} = 5\Omega$: $V_m = \dfrac{I_{em}R_{IN}}{(1 - \cos(\theta/2))} = \dfrac{5 \times 379 \times 10^{-3}}{0{\cdot}741}$

$$V_m = 2{\cdot}56\text{V}.$$

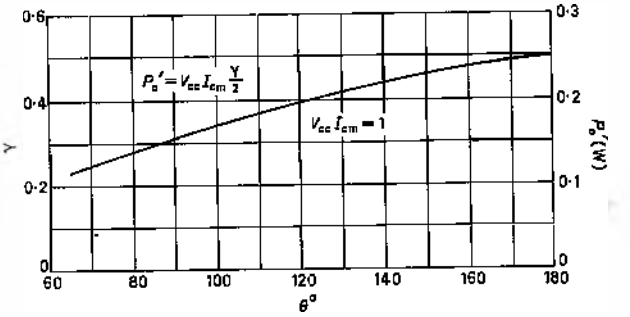


Fig. 1. Rates of fundamental/maximum collector current and normalized output power as a function of angle of conduction

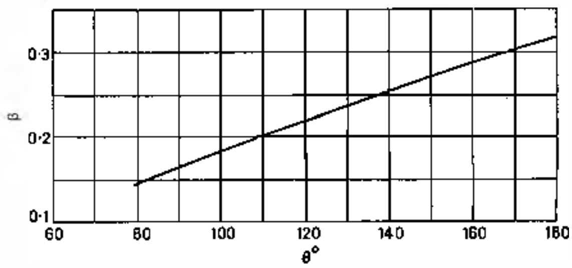


Fig. 2. Rates of average/maximum collector current as a function of angle conduction

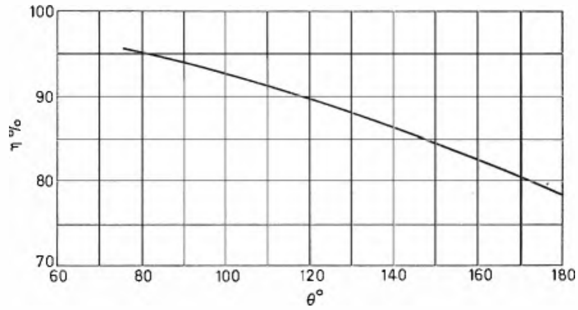


Fig. 3. Collector efficiency as a function of angle of conduction

$$I_{oe} = V_m/\pi R_{IN}\left(\sin(\theta/2) - (\theta/2)\cos(\theta/2)\right) = \frac{2{\cdot}56 \times 0{\cdot}627}{5\pi}$$

$$= 102{\cdot}2\text{mA}$$

$$V_B = V_m \cos(\theta/2) = 2{\cdot}56 \times 0{\cdot}259 = 0{\cdot}66\text{V}$$

$$I_{oe}V_B = 102{\cdot}2 \times 0{\cdot}66 = 67{\cdot}5\text{mW}$$

$$R_B = V_B/I_{oe} = \frac{0{\cdot}66}{0{\cdot}1022} = 6{\cdot}5\Omega$$

Power Gain $G = \dfrac{\gamma V_{cc}\alpha\left(1 - \cos(\theta/2)\right)}{\rho V_m} = \dfrac{0{\cdot}445}{0{\cdot}337}$

$$\frac{7{\cdot}06 \times 0{\cdot}82 \times 0{\cdot}741}{2{\cdot}56} = 2{\cdot}26$$

Check: $P_{IN(pk)} = \dfrac{V_m^2}{2\pi R_{IN}}\psi = 2{\cdot}56^2/10\pi \times 0{\cdot}735 = 153\text{mW}$

where

$$\psi = \{\theta - (3/2)\sin\theta + (\theta/2)\cos\theta\}$$

$$\therefore\ P_{IN} = P_{IN(pk)} + I_{oe}V_B = 220{\cdot}5\text{mW}$$

Check: $R_{eq} = R_{IN}/\rho = 5/0{\cdot}337 = 14{\cdot}83\Omega$

$$\therefore\ P_{IN} = V_m^2/2R_{eq} = 2{\cdot}56^2/29{\cdot}66 = 220{\cdot}7\text{mW}.$$

$$\text{P.G.} = P_o/P_{IN} = 500/220{\cdot}7 = 2{\cdot}26 = 3{\cdot}54\text{dB}.$$

This power gain, based on the minimum f_T is satisfactory and will probably be greater.

Output Rejector Circuit

For reasonable harmonic rejection let $Q_w = 2{\cdot}5$ and $R_o = R_L = 50\Omega$.

$$C = Q_w/R_L'\omega = \frac{2{\cdot}5}{25 \times 5\pi \times 10^8} = 63{\cdot}6\text{pF}$$

$$L = 1/\omega^2 C = 0{\cdot}00637\mu\text{H}$$

Because this value is unrealistic, L may be increased by decreasing C, having a lower limit of 24pF.

$$L_2 = C_1/C_2\,L_1 = \frac{63{\cdot}6}{24} \times 6{\cdot}37 \times 10^{-3}\mu\text{H} = 0{\cdot}0169\mu\text{H}$$

In air with $A = \pi D^2/4$, $L = \dfrac{KD^2N^2}{l}\ \pi^2 \times 10^{-9}\text{H}$

where K is a form factor obtained from tables with (D/l) as the parameter. Using 'Cambion' ceramic coil forms type P.L.S.T.:

$$D = 0{\cdot}205\text{in}$$

$$l = 0{\cdot}3123\text{in}$$

With $(D/l) = 0{\cdot}656$, $K = 0{\cdot}7745$

$$L = 2{\cdot}612N^2 \times 10^{-3}\mu\text{H}$$

For $L_2 = 1{\cdot}69 \times 10^{-2}\mu\text{H}$, $N_2 = 2{\cdot}54$

„ $L_1 = 6{\cdot}37 \times 10^{-3}\mu\text{H}$, $N_1 = 1{\cdot}56$

Turns ratio $m/n = 2{\cdot}54/1{\cdot}56 = 1{\cdot}628$

Check: $m/n = \sqrt{(L_2/L_1)} = \sqrt{(16{\cdot}9/6{\cdot}37} =$

$$\sqrt{(C_1/C_2)} = \sqrt{(63{\cdot}6/24)} = 1{\cdot}628$$

$Q_w = 2{\cdot}5$ remains unchanged since looking into the coil at the tapping point, approximately $1\frac{1}{2}$ turns, the effective inductance is still $6{\cdot}37 \times 10^{-3}\mu$H.

It is convenient to have a small trimmer for tuning, therefore C_{min} may be raised slightly or L_2 reduced.

Let $N_2 = 2$: At the centre tap: $N_1 = 1$

$$L = 2{\cdot}61 \times 10^{-3}\mu\text{H}$$

$$C = 155{\cdot}2\text{pF}$$

$$Q_w = R_L'/\omega L = \frac{25}{5\pi \times 10^8 \times 2{\cdot}612 \times 10^{-9}} = 6{\cdot}1$$

With $N_2 = 2$, $L = 4 \times 2{\cdot}612 \times 10^{-3}\mu\text{H} = 10{\cdot}45 \times 10^{-3}\mu\text{H}$

$$\therefore\ C = 38{\cdot}8\text{pF}$$

Assuming $R = 0{\cdot}1\Omega$, $Q_o = 10\sqrt{(L/C)} = 10\sqrt{\left(\dfrac{10{\cdot}45 \times 10^{-9}}{38{\cdot}3 \times 10^{-12}}\right)}$

$= 164$

In the absence of a trimmer $C_{min} = 24$pF; this is equivalent to 6pF across the 2-turn coil requiring 38·8pF. A 33pF trimmer will therefore be sufficient for tuning.

The complete output circuit of Fig. 4(a) has the original of Fig. 4(b) as an alternative. Fig. 4(a) has better harmonic rejection and lower insertion loss, but a greater decrement; based on insertion loss, it should provide greater output power.

Input Matching Circuit

With reference to Fig. 5, with $R_{IN} = 5\Omega$, $C_{IN} = 10$pF and $f = 250$MHz:

$$\text{Equivalent series resistance } R_s = \frac{R_{IN}}{1 + \omega^2 R_{IN}^2 C_{IN}^2} \simeq 5\Omega$$

$$\text{Equivalent series capacitance } C_s = C_{IN} + \frac{1}{\omega^2 R_{IN}^2 C_{IN}}$$

$$\therefore\ C_s = 10 + 1620 = 1630\text{pF}$$

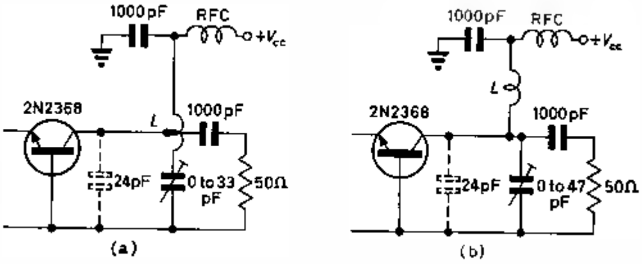


Fig. 4. Output rejector circuit

(a) $N = 2$, $L = 0{\cdot}01045\mu H$, $Q_o = 164$, $Q_w = 6{\cdot}1$

(b) $N = 1{\cdot}56$, $L = 0{\cdot}00637\mu H$, $Q_o = 100$, $Q_w = 2{\cdot}5$

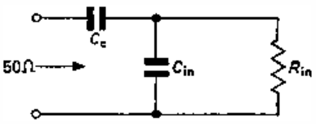


Fig. 5. Capacitor input matching

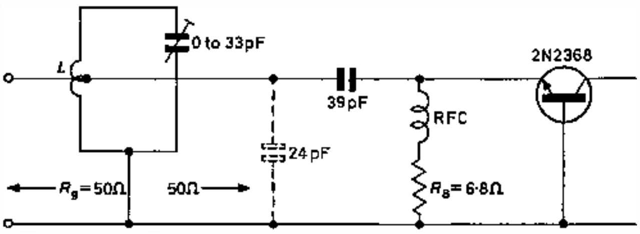


Fig. 6. Input rejector circuit

$L = 0{\cdot}01045\mu H$, $N = 2$, $Q_o = 164$, $Q_w = 6{\cdot}1$

$$\text{Equivalent parallel resistance } R_p = R_s + \frac{1}{\omega^2 R_s C^2} = R_g = 50\Omega$$

$$\therefore\ C^2 = \frac{1}{\omega^2 R_s (R_g - R_s)} = 42{\cdot}4\text{pF}$$

$$C_c = \frac{CC_s}{C_s - C} = 43{\cdot}6\text{pF}$$

A coupling capacitor C_c of 39pF should be satisfactory.

Input Rejector Circuit

With an effective load of 25Ω and $C_{min} = 24$pF, as in Fig. 4(a), the design will be unchanged, Fig. 6. An alternative will have the source and load directly across the coil with $Q_w = 1{\cdot}525$, Q_o remaining unchanged. The insertion loss will be negligible with the trimmer smaller in value and size. For good selectivity, the circuit of Fig. 6 is preferable.

A Practical Circuit

The performance figures for the circuit of Fig. 7, at

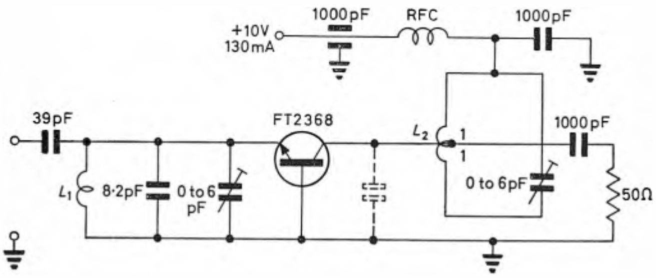


Fig. 7. 265MHz, 0·5W power amplifier
L_1—N = 2, PLST Cambion : L_2—N = 1:1, PLST Cambion

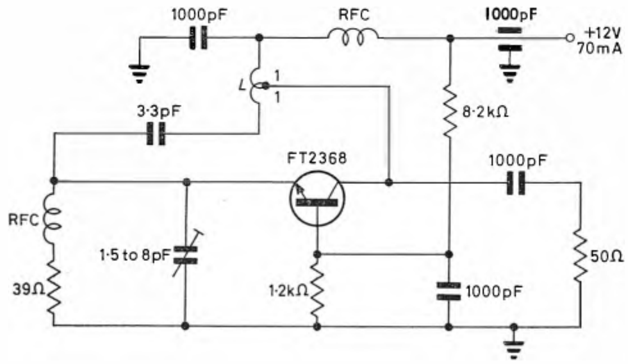


Fig. 8. 230MHz free-running oscillator

265MHz were:

P_o = 0·5W into 50Ω
P_{IN} = 50mW into 50Ω
I_o = 130mA.

With $R_B = 0$; $\theta = 180°$

$V_{cc} = 7{\cdot}06V$
$I_{cm} = I_{c1m}/\gamma = 141{\cdot}6/0{\cdot}5 = 283{\cdot}2mA$
$I_o = \beta I_{cm} = 0{\cdot}318 \times 283{\cdot}2 = 90{\cdot}1mA$

As previously, a sinusoidal current waveform with matching has been assumed. The excess current measured will be due to mismatch at the output and input, together with current waveform distortion; the stage has to be driven harder to obtain the required output. The high operating frequency, together with the use of the FT2368 transistor will contribute to I_o exceeding the maximum rated value of 100mA.

If the output admittance is known, a coupling capacitor may be used to match the load to it. A supply voltage of 7V is apparently sufficient, but the output power decreases for values less than 10V. The excess voltage allows for the collector bottoming voltage, the locus of which is the unity power gain curve. This curve permits the output conductance to be calculated[5].

Driver Circuits

The series fed oscillator of Fig. 8 is the 265MHz free-running driver for the power amplifier of Fig. 7 and is capable of 0·4W into 50Ω.

With two transistors in parallel located in a Redpoint type 18DC heatsink, the power amplifier of Fig. 7 with no bias resistor, has an output of 1W into 50Ω, when driven by the circuit of Fig. 8 at 230MHz; the collector efficiency is 50 per cent and overall efficiency 30 per cent.

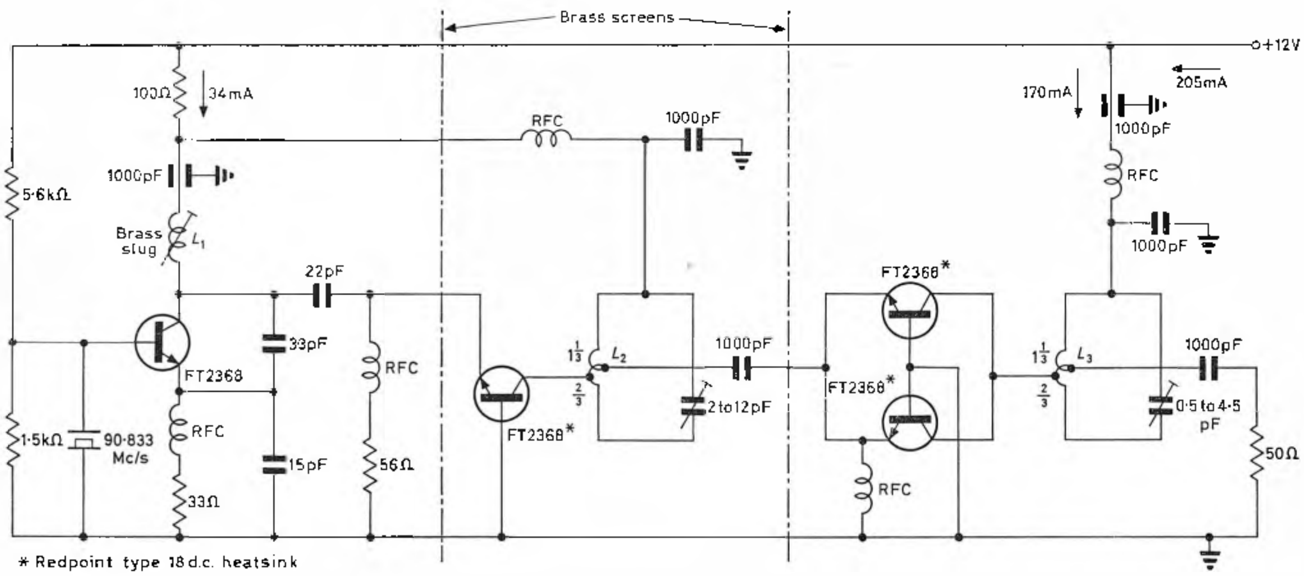
Because of their compactness, circuits such as Fig. 9 consisting of a power amplifier, tripler and crystal-controlled series fed oscillator, are housed in an outer screen of waveguide *X*, 1·5in in height. This circuit will give at least 0·6W into 50Ω at 272·5MHz, with a total supply drain of 2·4W. The oscillator works equally well with almost equal output power, using the Fairchild C111 transistor.

The common-base oscillator circuits given are basically Colpitts circuits with the anti-phase voltage across the tuned circuit divided between the collector-emitter feed-back capacitance and emitter-base input capacitance.

With the crystal in the base operating in the series resonant condition, its dissipation will be a minimum. It is desirable that the crystal oscillator operates as near class-A as possible, with its output reasonably independent of supply voltage variations.

A particularly serious fault with free-running oscillators is their tendency to change suddenly from an efficient three-terminal device to a two-terminal or negative resist-

Fig. 9. 272·5MHz crystal-controlled 0·6W source



ance device, which is less efficient with greater output power, while tuning. This discontinuity in their mode of operation can have serious consequences, with the destruction of the power amplifier transistors.

Conclusions

An average collector current of 130mA was required to obtain 0·5W across 50Ω; with $I_o = 100$mA, the decreased P_o would have to be acceptable. Other factors influencing the output power and efficiency of class-C amplifiers are:

(a) Current waveform distortion due to α decreasing with I_o. A constant $\alpha - I_c$ characteristic is required.

(b) Because of base spreading due to the frequency being comparable with the diffusion time of the holes in the base region, θ increases, increasing I_o, decreasing efficiency and output.

(c) The output current waveform is distorted due to the non-linearity of both the input and transfer characteristics, together with change over effect as the input resistance suddenly changes from a high to a low value during conduction.

(d) Hole-storage accelerates cut-off, since current flows in the reverse direction prior to cut-off. Because the hole-storage current increases I_o, P_{cm}, β and P_{cm}' increase, while P_o decreases.

(e) Since I_o increases with frequency, so must I_{cm}, therefore γ decreases i.e., I_{c1}.

(f) Because of positive feedback through C_{ce}, the common-base configuration has greater power gain at v.h.f.; the common-emitter circuit has lower gain because of Miller effect through C_{cb}.

(g) Because the transistor is not bottomed during conduction, its transfer efficiency is low.

But for the frequency, the collector current waveform could be observed on an oscilloscope by monitoring the voltage drop across a small resistor in the supply and the angle of flow θ altered to the required value by changing R_B. When the collector bottoms, the collector-base diode is forward biased causing the current to reverse, indicated by a notch in the peak of the waveform. While tuning, this notch will shift about the peak value, being symmetrical for correct tuning.

To provide the large input currents required by the power amplifier, a voltage source is required. Because the output resistance of the driver is reduced to low values in regions of high currents, it is capable of providing these currents.

Acknowledgment

The author would like to thank the Directors of Microwave Associates Ltd., in particular, Mr. J. J. Tither, the General Manager, for permission to publish this article. Thanks are due to Mr. J. Eagling for his valuable assistance with the experimental work.

Appendix (1)

With reference to Fig. 10:

$$V_{IN} = V_m \sin\alpha - V_B = V_m \cos(\theta/2) - V_B \quad \ldots\ (1)$$

$$(V_m - V_B) = I_{em} R_{IN} \quad \ldots\ (2)$$

where I_{em} is the peak input current with base common.

$$V_B = V_m \cos(\theta/2) \quad \ldots\ (3)$$

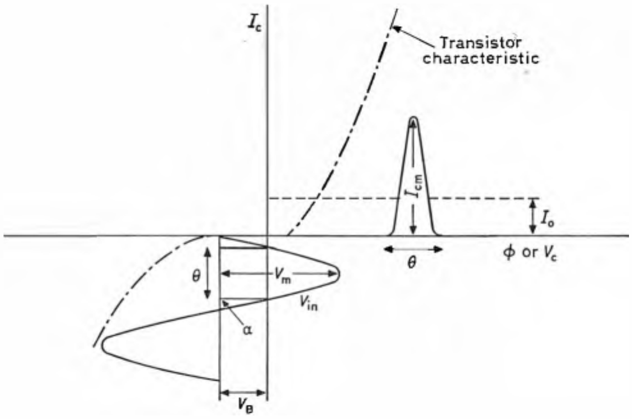


Fig. 10. Transistor input characteristic

Let:

$$i_c = f(\phi) = G_m(V_m \cos\phi - V_B) \text{ for } -(\theta/2)<\phi<(\theta/2)$$
$$i_c = f(\phi) = 0 \qquad \text{for } (\theta/2)<\phi<2\pi \quad \ldots (4)$$

For an even function: $f(\phi) = a_o + \sum_{n=1}^{\infty} a_n \cos n\phi$

$$a_o = 1/2\pi \int_{-\theta/2}^{\theta/2} f(\phi)\, d\phi = (G_m/\pi) \int_0^{\theta/2} (V_m \cos\phi - V_B)\, d\phi$$

$$a_o = I_o = (G_m V_m/\pi)\,[\sin(\theta/2) - (\theta/2)\cos(\theta/2)] \quad \ldots\ (5)$$

$$a_1 = 1/\pi \int_{-\theta/2}^{\theta/2} f(\phi)\cos\phi\, d\phi = 2G_m/\pi \int_0^{\theta/2} (V_m\cos\phi - V_B)\cos\phi\, d\phi$$

$$a_1 = I_{c1} = G_m V_m/\pi\,[(\theta/2) - (1/2)\sin\theta] \quad \ldots\ (6)$$

Also $i_c = G_m(V_m\cos\phi - V_B) = G_m V_m(\cos\phi - \cos(\theta/2))$

When:

$$\phi = 0:\ i_c = I_{cm} = G_m V_m(1 - \cos(\phi/2)) \quad \ldots\ (7)$$

$$\gamma = I_{c1}/I_{cm} = (1/\pi)\,.\,\frac{(\theta/2) - (1/2)\sin\theta}{1-\cos(\theta/2)} = 1/\pi$$

$$\frac{(\theta/2) - \sin(\theta/2)\cos(\theta/2)}{1-\cos(\theta/2)} \quad \ldots\ (8)$$

$$\beta = I_o/I_{cm} = 1/\pi\,\frac{\sin(\theta/2) - (\theta/2)\cos(\theta/2)}{1-\cos(\theta/2)} \quad \ldots\ (9)$$

$$\gamma/\beta = I_{c1}/I_o = \frac{(\theta/2) - \sin(\theta/2)\cos(\theta/2)}{\sin(\theta/2) - (\theta/2)\cos(\theta/2)} \quad \ldots\ (10)$$

Collector efficiency $\eta = 1/2\,(I_{c1}/I_{co}) =$

$$1/2\,(\gamma/\beta) = 1/2\,\frac{(\theta/2) - \sin(\theta/2)\cos(\theta/2)}{\sin(\theta/2) - (\theta/2)\cos(\theta/2)} \quad \ldots\ (11)$$

γ, β and η are plotted as functions of the angle of conduction θ, in Figs. 1, 2 and 3 respectively.

Input Power P_{IN}

$$V_{IN} = V_m\cos\phi - V_B$$

$$P_{IN(pk)} = \frac{1/2\pi \int_{-\theta/2}^{\theta/2} (V_m\cos\phi - V_B)^2 d\phi}{R_{IN}}$$

$$P_{IN(pk)} = \frac{V_m^2}{2\pi R_{in}}\{\theta-(3/2)\sin\theta+(\theta/2)\cos\theta\} \quad \ldots (12)$$

$$I_{oe} = \frac{1}{2\pi R_{IN}}\int_{-\theta/2}^{\theta/2}(V_m\cos\phi - V_B)\,d\phi \quad \ldots\ldots\ldots\ldots\ldots (13)$$

$$\therefore I_{oe} = V_m/\pi R_{IN}\,[\sin(\theta/2) - (\theta/2)\cos(\theta/2)] \quad \ldots (14)$$

$$P_{IN} = P_{IN(pk)} + I_{oe}V_B = \frac{V_m^2}{2\pi R_{IN}}\,[(\theta/2) - (1/2)\sin\theta]$$

$$= (V_m^2/2R_{IN})\,.\,\rho = V_m^2/2R_{eq} \quad \ldots\ldots (15)$$

where $R_{eq} = R_{IN}/\rho$.

Power Gain P.G.

From Fig. 1:

$$P_o = P_o' V_{cc} I_{cm} \quad \ldots\ldots\ldots\ldots\ldots\ldots (16)$$

From equations (2) and (3):

$$V_m^2 = \frac{I_{cm}^2 R_{IN}^2}{(1-\cos(\theta/2))^2}$$

$$\therefore\ V_m^2/R_{IN} = \frac{I_{em}^2 R_{IN}}{(1-\cos(\theta/2))^2} = \frac{I_{cm}^2 R_{IN}}{|A_i|^2\,(1-\cos(\theta/2))^2}$$

$$\therefore\ \frac{V_m^2/R_{IN}}{V_{cc}I_{cm}} = \frac{I_{cm}R_{IN}}{V_{cc}\,|A_i|^2\,(1-\cos(\theta/2))^2} = \frac{2P_{IN}P_o'}{\rho\,P_o} \quad \ldots\ldots\ldots\ldots (17)$$

From equation (17):

$$\text{P.G.} = P_o/P_{IN} = \frac{2P_o'V_{cc}\,|A_i|^2\,(1-\cos(\theta/2))^2}{\rho\,I_{cm}R_{IN}}$$

$$\therefore\ \text{P.G.} = \frac{\gamma\,V_{cc}\,|A_i|\,(1-\cos(\theta/2))}{\rho\,V_m} \quad \ldots\ldots\ldots\ldots\ldots (18)$$

with γ obtained from Fig 1.

Appendix (2)

Pulse Response

Let R equal the tank loss resistance in Fig. 11:

$$I = I_1 + I_2 + I_3 = C\,(dV/dt) + (V/R) + (1/L)\int V\,dt \quad \ldots\ldots\ldots\ldots (1)$$

$$f(t) = -e = L\,(dI/dt) = LC\,(d^2V/dt^2) + (L/R)\,(dV/dt) + V \quad \ldots\ldots\ldots\ldots (2)$$

Let $F(s) = \mathcal{L}\,f(t) = f(t) + (1/s)\,f'(t) + (1/s^2)f''(t) + \ldots$

$$(LC\,s^2 + (L/R)\,s + 1)\,F(s) = \mathcal{L}\,f(t) \quad \ldots\ldots\ldots (3)$$

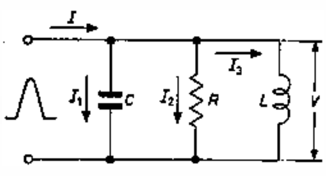


Fig. 11. Pulsed output circuit

For class-B operation:

$$I_c = I_{cm}/\pi\,(1 + (\pi/2)\sin\phi - (2/3)\cos 2\phi - (2/15)\cos 4\phi \ldots)$$

Considering the fundamental only:

$$L\,(dI/dt) = \frac{\omega L I_{cm}}{2}\cos\omega t = f(t) \quad \ldots\ldots\ldots\ldots (4)$$

$$\mathcal{L}\,f(t) = \int_0^\infty a\cos\omega t\,e^{-st}\,dt \quad \text{where } a = \frac{\omega L I_{cm}}{2}$$

$$\mathcal{L}_{T\to\infty}\,f(t) = \frac{a\,s}{(s^2+\omega^2)} \qquad \text{where } s = (\sigma + j\omega) \quad \ldots\ldots (5)$$

$$(LC\,s^2 + (L/R)\,s + 1)\,F(s) = \frac{a\,s}{(s^2+\omega^2)} \quad \ldots\ldots (6)$$

$$F(s) = \frac{a\,s}{(s^2+\omega^2)}\,\frac{1}{(LC\,s^2 + Ks + 1)}$$

$$= (a/K)\,\frac{1}{s^2+\omega^2} - (a/K)\,\frac{1}{(s+\alpha)\,(s+\beta)} \quad \ldots\ldots\ldots (7)$$

where:

$$\alpha = \frac{-\omega^2 K + \sqrt{(\omega^4K^2 - 4\omega^2)}}{2}$$

$$\beta = \frac{-\omega^2 K - \sqrt{(\omega^4K^2 - 4\omega^2)}}{2}$$

$$K = L/R$$

Now:

$$\frac{1}{(s+\alpha)}\,\frac{1}{(s+\beta)} = \frac{1}{(\beta-\alpha)}\,\frac{1}{(s+\alpha)} + \frac{1}{(\alpha-\beta)}\,\frac{1}{(s+\beta)}$$

$$F(s) = (a/K)\left[\frac{1}{(s^2+\omega^2)} - \frac{1}{(\beta-\alpha)}\,\frac{1}{(s+\alpha)} - \frac{1}{(\alpha-\beta)}\,\frac{1}{(s+\beta)}\right] \quad \ldots\ldots\ldots\ldots (8)$$

$$f(t) = (a/K)\left[(1/\omega)\sin\omega t - \frac{1}{(\beta-\alpha)}\,e^{-\alpha t} - \frac{1}{(\alpha-\beta)}\,e^{-\beta t}\right]$$

$$f(t) = \frac{\omega I_{cm}R}{2}\left\{(1/\omega)\sin\omega t + \frac{1}{(\alpha-\beta)}\,[e^{-\alpha t} - e^{-\beta t}]\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (9)$$

$(\alpha - \beta) = \omega/Q_w\ \sqrt{(1 - 4Q_w^2)}$ and $R = Q_w\omega L$.

$$\lim_{Q_w\to 0} e(t) = f(t) = \frac{I_{cm}R}{2}\sin\omega t$$

$$= I_{c1}R\sin\omega t = V_T/2\sin\omega t$$

Thus when Q_w is low, the instantaneous voltage amplitude across the tank is constant, i.e. there is no decrement.

For $(1 - 4Q_w^2) \neq 0$, β is negative and always greater than α, hence the second exponent is positive and large making the second term in $e(t)$ negative, the response decreasing with time. $\lim_{Q_w\to\infty} e(t) = 0$, hence P_o decreases inversely with Q_w.

In general:

$$e(t) = I_{c1}R\left\{\sin\omega t + \frac{Q_w}{\sqrt{(1-4Q_w^2)}}\,[e^{-\alpha t} - e^{-\beta t}]\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (10)$$

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Simulation of Cone of Confusion

By S. Dalal*, B.Sc.

In the design of radio aid navigation simulation for a flight simulator, engineers often encounter the problem of simulating the cone of confusion which exists on top of a radio beacon. The system described is a comparatively cheap, completely transistorized and an integrated system to simulate a realistic cone of confusion.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

ON top of any v.o.r. beacon, used for radio navigation, there exists a zone where little or no beacon signal is present. The shape of the zone is shown in Fig. 1. When an aircraft enters this zone, the r.m.i. indicator or radio compass which gives the bearing of this beacon, starts to hunt at 2Hz. The amplitude of this hunt progressively increases as the aircraft approaches the top of the beacon, the r.m.i. indicator then rotates 180° and the amplitude of this hunt progressively decreases as the aircraft goes away from the beacon.

Before describing the system which will simulate the cone of confusion in a flight simulator, it is necessary to describe the system used in flight simulators to obtain the bearing of a beacon.

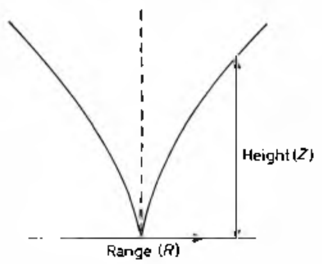

Fig. 1. Cone of confusion zone on a radio beacon

The system, shown in Fig. 2, is an a.c. system which transforms cartesian co-ordinates to polar co-ordinates.

If X co-ordinate is applied to OX and Y co-ordinate is applied to OY, the voltage will be induced to two windings RN and MN of the resolver. MN is fed to the armature of a motor via an amplifier A. The field of the motor is 90° out of phase with the output of the amplifier A. The motor M will rotate, and as it is joined with the resolver by a shaft, the resolver will also rotate. When the voltage at MN is zero the motor will stop. Then the voltage at RN is the range signal and the angle of turn of the resolver is the bearing angle. The bearing angle is transmitted to the instrument of an aircraft by a teletorque instrument. Fig. 2 is represented in Fig. 3, in conventional servo-mechanism form.

If a voltage n is injected preceding the amplifier A then the equation of the system is

$$J\,\frac{d^2\,\theta_o}{dt^2} + F\,\frac{dt}{d\theta_p} = A\,(\theta_i - \theta_o + n).$$

(This system is shown in Fig. 4.)

The transient solution of the equation is

$\theta_o\ =\ \exp\ (-Ft/2J)\ P\ \sin\ (\omega f + \alpha)$ where P and α are constants.

The steady state solution is

$$\theta_o = \frac{A\,(\theta_i + n)}{JD^2 + FD + A} = \frac{A\,(\theta_i + n)}{A} = \theta_i + n$$

Neglecting the transient state it is seen that if a voltage n is injected at the input of the amplifier A, (refer to Fig. 2), the motor M will rotate $\theta_i + n$, instead of θ_i. Now if the phase of n is made to oscillate 180° at 2Hz, then θ_o will oscillate from $\theta_i + n$ to $\theta_i - n$, or in other words the system will appear to be hunting around θ_i at 2Hz.

At any angle θ, the voltage across MN is $N \sin \theta - E \cos \theta$. Where N and E are the voltages applied to OX and

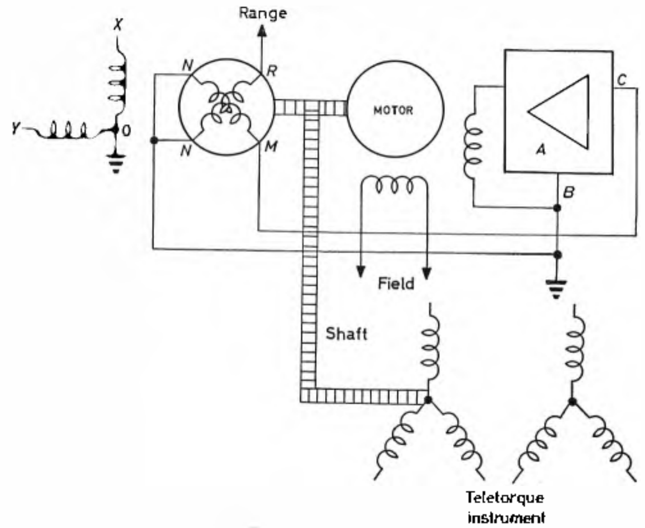

Fig. 2. System used in a simulator to obtain the bearing of a beacon

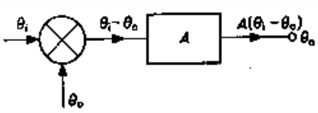

Fig. 3. Mathematical representation of existing system

OY respectively. The motor and the resolver will assume a position where $N \sin \theta = E \cos \theta$ hence the output at NM is zero. The angle θ will be the bearing angle and the voltage across RN is the range signal.

To investigate the angle of deflexion θ for any voltage v, injected at MN, it is assumed that the aircraft is approaching the beacon from the north and E is zero.

$\therefore$ voltage $v = N \sin \theta$ or $\theta = \sin^{-1} v/N$

If the aircraft approaches the beacon at any other angle, v gives the same deflexion but the calculation becomes complex.

When E is zero N is also the range of the aircraft from the beacon. If v is kept constant N is inversely proportional to $\sin \theta$. Hence as the aircraft approaches the beacon, N decreases and θ increases. So if a constant voltage is made to oscillate 180° in phase at 2Hz and if it is injected at MN, then the system will appear to be hunting at 2Hz and the amplitude of this hunting will progressively increase as the aircraft approaches the beacon.

The system shown in Fig. 5 detects and simulates the cone of confusion.

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Description of the System

Assume the scale of Range signal/height signal $= m$, and when the aircraft enters into the cone

$\therefore$ Range, R/height $Z = C$

$$\left.\begin{array}{l}\therefore R = ZC\\ \text{Or } R - ZC = O\end{array}\right\}\text{at the cone of confusion.}$$

When the aircraft is out of the cone the function $R-ZC$ is positive, when the aircraft is inside the cone, the function is negative.

Referring to Fig. 5, $R_2/R_1 = mC$

Hence the output of the amplifier A_1 is proportional to $-ZC$, $R_3 = R_4 = R_5$, so the output of the amplifier A_2 is proportional to $-R + ZC$. A_3 is acting as an amplifier and the output of A_3 is proportional to $R - ZC$.

Hence the output of A_2 will shift 180° when the aircraft enters the cone. The output of A_3 is fed to a p.s.r. (phase sensitive rectifier) and the output of the p.s.r. is smoothed by a 1μF capacitor. Therefore the voltage of the capacitor will be positive when the aircraft is out of the cone and negative when the aircraft is in the cone.

S_1 is a switching amplifier which is switching relay B on at the cone and switching it off when the aircraft is out of the cone.

M_1 in the system is a free running multivibrator, running at approximately 2Hz switching relay A on and off.

P_1 is a phase-splitter; suitable input is fed by choosing the ratio of R_6 and R_7. The output of this phase-splitter is fed to amplifier A_1 via a contact of relay A and R_8.

Assume the output of the phase splitter is $\pm n$ volt. At the point L of contact A_1 the phase of voltage n, is oscillating 180° at 2Hz.

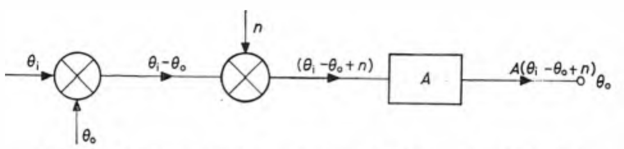
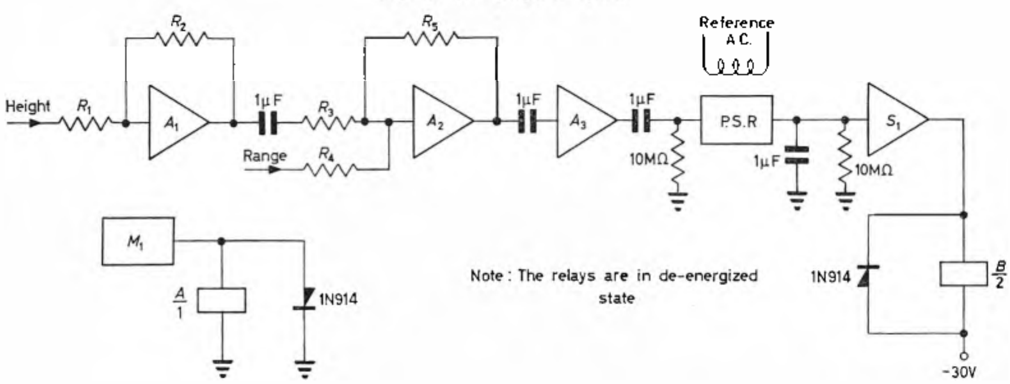
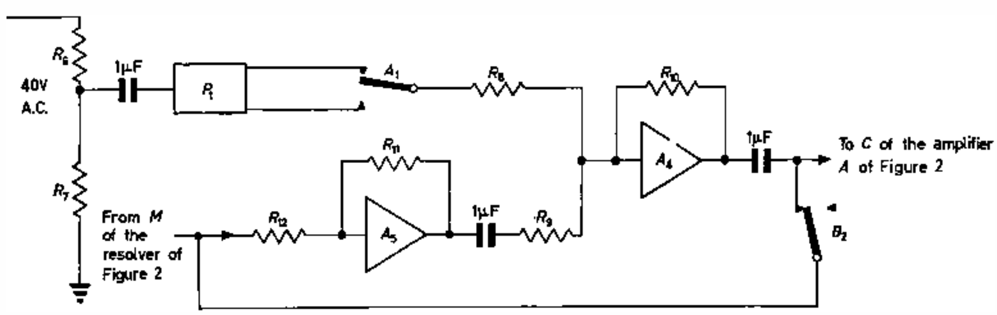


Fig. 4. Mathematical representation of proposed modification

Fig. 5. The proposed system





Assume the input of the amplifier A_5 from the resolver is θ_1 at one instant.

The output of the resolver is fed to amplifier A_4 via A_5 and R_9. $R_{11} = R_{12}$ and $R_8 = R_9 = R_{10}$. Hence the output of the amplifier A_4 is oscillating from $\theta_1 + n$ to $\theta_1 - n$ at 2Hz. When the resolver finds a bearing θ_1, the input to A_5 is zero and the output of A_4 then is $\pm n$ volt. This voltage is fed to the input of the amplifier A of Fig. 2. Hence the resolver will appear to be hunting around θ_1.

When the aircraft is out of the cone, relay B breaks, hence the power supply of R_1 breaks and amplifiers A_4 and A_5 are short-circuited. The contacts of relay B are employed also to switch off the audio tone of the v.o.r. beacon and other suitable functions.

Description of Circuits

Data given

(1) Range scale 2·8nm/V

(2) Height scale 40V at 32 000ft

(3) At 14 000ft the cone starts at 1·5nm from the beacon.

Because of the large difference between range scale and height scale, the height scale has been stepped down by R_1, R_2, R_3 and R_4.

In the complete circuit diagram in Fig. 6, S_1 consists of VT_7, VT_8, VT_9 and VT_{10}. One of the criteria of this circuit is that S_1 has a very high input impedance. At 3mA collector current α' of VT_8 is 80 and at about 40mA collector current of VT_7 is 50. Therefore the input impedance of this stage is $80 \times 50 \times 10\text{k} = 40\text{M}\Omega$. When the emitter of MR_5 goes negative, diode VT_8 does not conduct, hence VT_9 and VT_{10} conducts by its own bias and the relay is switched on. When on the other hand emitter of VT_8 goes positive, diode MR_5 conducts, and VT_9 and VT_{10} are cut off and relay B breaks.

The advantages of this circuit are:

(1) The switch action is devoid of any variation of power supply, i.e. the base of VT_7 will remain biased at zero if the power supply voltages change.

(2) As silicon transistors are used, leakage current variation with temperature will be small.

(3) The input impedance VT_9 and VT_{10} is very high as MR_5 conducts only when VT_9 and VT_{10} are cut off. Hence the voltage drop due to current depletion at the emitter of VT_8 is very small indeed.

It has been found that approximately at $-0{\cdot}5$V at the base of VT_7, the changeover of relay B occurs.

P.S.R. Circuit

This is a conventional phase sensitive rectifier circuit.

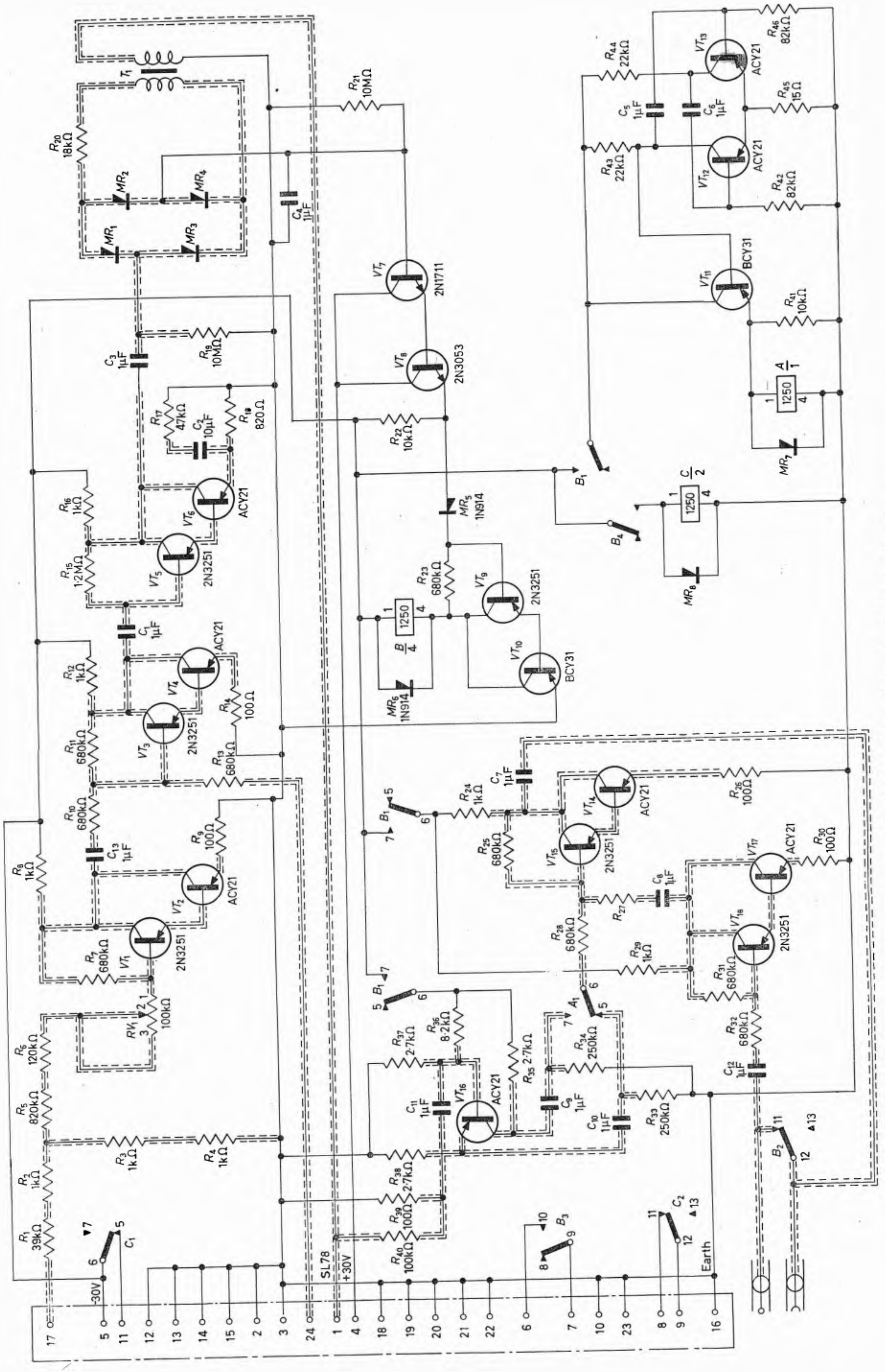


Fig. 6. Cone of confusion board circuit diagram

The reference voltage, applied at T_1, is in phase with Z and R signal. The output of this p.s.r. is rectified half waves, positive going when the function $(R - mcZ)$ is positive and negative going when the function is negative. Hence the capacitor C_4 is charged accordingly.

Amplifier A_3 Circuit

Amplifier A_3 consists of VT_5 and VT_6

α' of VT_5 is 80 at 0·1mA emitter current and α' of VT_6 is 90 at 30mA emitter current. The pair has been biased by 1·2MΩ, the d.c. collector current is 15mA and the collector d.c. voltage is −15V. An emitter resistance of 820Ω is necessary, otherwise to provide a suitable base current, a high base to collector resistance is needed. A.C.-wise, however, the emitter resistance is by-passed by a series combination of 47Ω and 10μF. Hence about 15V a.c. swing at the collector is possible. The a.c. input impedance of this stage is $90 \times 80 \times 47 = 7\,200 \times 47 = 400\text{k}\Omega$ and the a.c. voltage amplification is 10. The phase shift at the output for 1kΩ and 10μF is negligible. A_3 amplifier can be designed with only one ACY21.

A_1 and A_2, A_4 and A_5 Amplifier Circuit

The same circuit has been used in all amplifiers.

The input impedance is $7\,200 \times 100 = 720\text{k}\Omega$.

The collector d.c. voltage is −25V and a.c. voltage gain is approximately 10. It serves as a slightly inaccurate computing amplifier due to low voltage gain but this fact has been used to provide the curvature of the cone and has been discussed in the conclusion.

A_2 amplifier is used as a summing amplifier and at the input the function $R - mcZ$ is fed; at the output $R + mcZ$ appears. The function is inverted again by A_3.

It has been found that switching occurs at −0·5V. A_3 has a voltage amplification of 10. Therefore at the input of A_2, instead of providing $R - mcZ$, it is necessary to change mc, so that at the required Z and R, $R - mcZ$ is not zero but −50mV.

Cone of confusion occurs when the aircraft is at 14 000ft and the range is 1·5nm. At 14 000ft Z signal from the Z slave servo is 17·47V and the range signal is 0·536V or 536mV.

At the input of R_5, the voltage is 830mV. The amplifier A_1 should attenuate 830mV to $536 + 50 = 586$mV. Therefore, the computation resistance is $(680\text{k} \times 830)/586 = 964\text{k}\Omega$. $R_5 + R_6 = 940\text{k}\Omega$ and RV_1 is 100kΩ.

M_1, which is a conventional multivibrator, and P_1 a conventional phase splitter need not be discussed. It is sufficient to say that 40mV has been chosen as an input to the phase splitter.

The difference between the system diagram, Fig. 5, and circuit diagram, Fig. 6, is that in the actual circuit, multivibrator, phase splitter and the summing amplifier, which is summing resolver bearing angle, and error, has been switched on by relay B.

In the circuit diagram, Fig. 6, resolver bearing angle is fed in via PL_2 and bearing angle + error is fed to the amplifier, A (Fig. 2) via SK_1.

Multivibrator M consists of VT_{11} to VT_{13} and the phase splitter consists of VT_{16}.

Input of PL_1

Sockets 17, from Z slave servo; sockets 5, −30V.

Sockets 24, range; sockets 1, 40V a.c.; sockets 4, +30V.

Results and Conclusion

Table 1 shows the deflexion, θ of the r.m.i. needle with respect to range, or the angle of hunt/range.

From 0·3nm range, range signal will be affected by the resolver oscillation and the amplitude of hunting will be more than predicted, but up to 1nm range the variation of range signal due to oscillation is negligible.

A graph of expected cone of confusion and experimentally obtained cone of confusion is shown in Fig. 7.

As mentioned before, due to inaccuracy of the computing amplifier, the cone does not consist of straight

TABLE 1

RANGE *NM*	VOLTAGE EQUIVALENT TO RANGE (V)	$\sin\theta = \frac{40 \times 10^{-3}}{V}$	$\theta°$
3·5	1·25	0·032	±1·8
3	1·07	0·037	±2·2
2·3	1	0·04	±2·3
2·5	0·894	0·0447	±2·6
2·2	0·785	0·051	±3
2	0·715	0·056	±3·2
1·8	0·643	0·062	±3·6
1·5	0·536	0·074	±4·3
1·2	0·428	0·093	±5·3
1	0·357	0·112	±6·5
0·8	0·286	0·140	±8
0·5	0·1785	0·224	±13
0·3	0·1071	0·373	±21·9
0·2	0·0714	0·56	±34·1
0·1	0·0357		

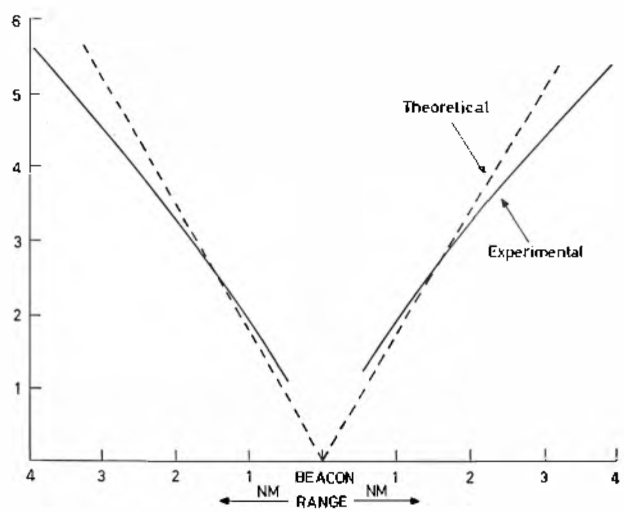


Fig. 7. Experimental and theoretical cone of confusion curves of the system

lines but curved lines. In this case this is needed because in an actual case the cone of confusion is curved, it is narrow at the bottom and wide at the top. The exact nature of the curvature can be simulated, if it is known, with amendments to this circuit. In this case the nature of curvature is not controlled but is an effect of inaccuracy. Only one datum was given to the author that the diameter of the cone is 3nm at the height of 14 000ft, and that is met.

It is, however, advisable to set up one point in the middle between 14 000ft to 20 000ft depending upon the maximum height the simulator can achieve. Then the curvature, due to inaccuracy, will render a realistic cone of confusion simulation.

Acknowledgment

The author thanks the authorities of General Precision Systems Ltd, Aylesbury, for their kind permission to publish this article. He also thanks his colleagues at G.P.S. for their help and encouragement in the preparation of this article.

A Square-Law Detector for Power Measurements

By J. F. Golding*

The detector circuit arrangements described in this article were originally devised for the purpose of producing an oscilloscope deflexion proportional to the mean power output from a receiver. The principle can, however, be applied in other ways including the use of the circuit as an aperiodic frequency doubler.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

THE transfer characteristics of most point-contact germanium diodes are very near to true square law for voltages less than about a volt. Admittedly this is no virtue when it comes to designing a sensitive linear detector; but it can be a very useful attribute when the problem is that of producing a d.c. voltage which is proportional to the mean power in a load.

Such an application may seem a somewhat unlikely one to some readers; so a brief note about the original application is not out of place.

The Original Purpose

The detector circuits described in this article are intended primarily for use in a swept-frequency system for the measurement of radio receiver sensitivity. Fig. 1 shows the basic test-gear arrangement, in which the frequency of the signal generator is swept through the pass band of the receiver in unison with the traverse of the spot across the oscilloscope screen.

With the modulation applied to the signal, a display of the frequency/response curve is obtained, the amplitude at the centre frequency being proportional to the 'in tune' output of the receiver. With the input signal unmodulated, the displayed curve shows the output noise level as a function of input carrier frequency. It is thus a simple matter to compare the displayed noise level with the signal-plus-noise at the centre frequency, to obtain an accurate assessment of signal-to-noise ratio and, hence, the noise-limited sensitivity of the receiver.

However, signal-to-noise ratio is necessarily a power ratio; so the detector must deliver to the oscilloscope an output voltage which is proportional to the power dissipated in the a.f. load. For the purposes of the measurement this load is assumed to be resistive, so that a simple square-law device can be regarded as an adequate conversion system.

The receiver measurements themselves form the subject of a separate article; and, from this juncture, attention will be confined to the design of the detector.

The Basic Detector Arrangement

The full line in Fig. 2 is the relevant portion of the forward transfer characteristic of an HG1005 diode; and the broken line is the plot of a true square law ($I = KV^2$). It can be seen that the diode characteristic is parallel to the calculated curve for voltages up to 800mV, but is displaced from it by about 75mV. In practice, the error due to this displacement is often less than the overall measurement discrimination figure. It can, however, be virtually eliminated by means of a small bias voltage applied in a manner described later in this article.

* *Marconi Instruments Ltd.*

Since the detector must accept asymetric as well as symetric waveforms its overall characteristic must be of the full involute square-law form. So the basic circuit utilizes the full-wave arrangement shown in Fig. 3(a).

The current in each of the diodes at any instant is proportional to the square of the voltage across the diode

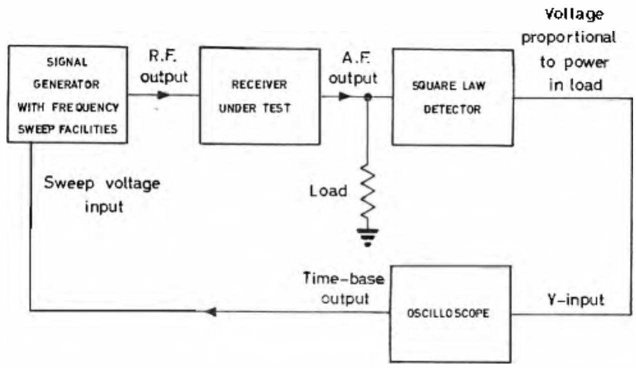
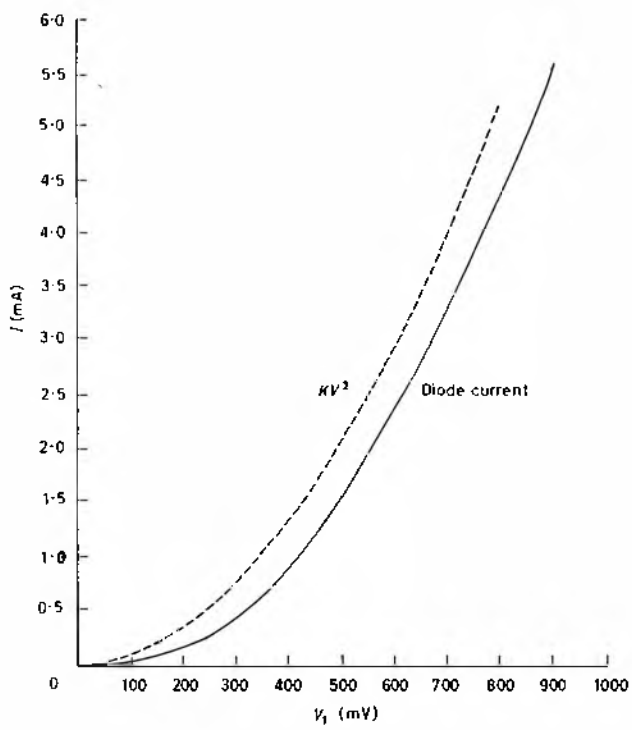


Fig. 1. Basic circuit for swept-frequency method of sensitivity measurement

Fig. 2. The diode transfer characteristic compared with a calculated square-law curve.
Note that the diode characteristic is parallel to the calculated curve but displaced from it by about 100mV



itself—not the voltage across the whole circuit. For minimum error, therefore, it is important that the total fixed resistance should be small compared with the forward resistance of the diode. Fig. 3(b) shows that the fixed resistance comprises resistor R_2 in series with the parallel combination of R_1 and R_s, where R_s is the resistance of the source. In practice this parallel combination is usually so much lower than the diode resistance that it can be neglected. Indeed, if this is not the case, the loading effect of the diode circuit must be taken into account; but this circumstance is too rare to warrant serious attention.

The output voltage, V_{out}, is, of course, developed across R_2. So, although the value of this resistor must be low in order to preserve the square-law characteristic, it must not be so small that the output voltage is unusable.

To select a suitable value it is first necessary to assess

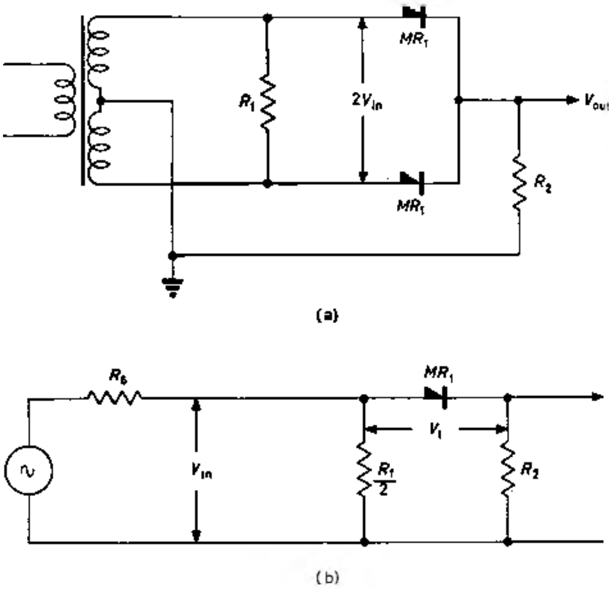


Fig. 3. Basic detector arrangement
(a) Actual circuit (b) Equivalent circuit

the magnitude of the error due to the presence of R_2; and to do so the following reasoning may be used.

The output voltage—equal to IR_2—can be expressed as:

$$V_{out} = kR_2 \cdot V_1^2$$

where k is the transfer constant of the diode.

But the input voltage is equal to $V_1 + V_{out}$; so

$$V_{out} = kR_2\ (V_{in} - V_{out})^2 = kR_2 \,.\, V_{in}^2\ (1 - V_{out}/V_{in})^2$$

Ideally, of course, V_{out} should be equal to V_{in}^2 multiplied by a constant, and therefore, the term

$$(1 - V_{out}/V_{in})^2$$

can be regarded as a factor of error. For the acceptable accuracy, this factor can be simplified to

$$1 - 2\ V_{out}/V_{in}$$

i.e., the error is equal to $-2(V_{out}/V_{in})$

Providing this is small with respect to unity, one can —for the sake of the calculation—regard V_{out} as being very nearly equal to $kR_2 \cdot V_{in}^2$, so that the error, E, is equal to $2kR_2 \cdot V_{in}$.

For an HG1005 diode, $k = 8 \times 10^{-5}$ (approximately) when V is in millivolts and I is in milliamperes. The maximum possible input voltage for square-law transfer is about 800mV; and a 10 per cent error at this level would be completely masked by other inaccuracies in the system. So, rewriting the error expression to give R_2 and then substituting the values,

$$R_2 = E/2kV_{in}$$
$$= (1/2) \times (1/10) \times \frac{1}{8 \times 10^{-5} \times 8 \times 10^2} = 7{\cdot}8\Omega$$

In the experimental detector a 10Ω resistor gave quite satisfactory results.

The detector in the form shown in Fig. 3 merely delivers a voltage which is proportional to the instantaneous power in R_1. So if the voltage across R_1 is the sine wave

$$V_{in} = V_{peak} \sin \omega t,$$

the output voltage across R_2 will be given by

$$V_{out} = KV_{peak}^2\ \tfrac{1}{2}\,(1 - \cos 2\omega t) \text{ where } K = kR_2.$$

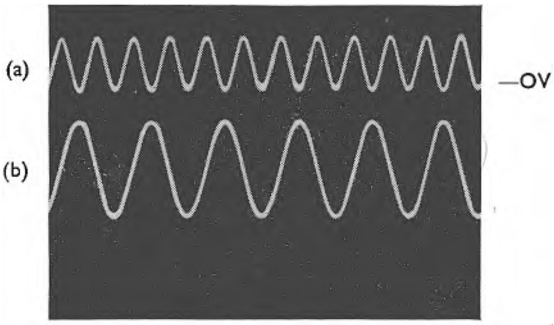


Fig. 4. Input and output waveforms.
Note that V_{out} is at twice the input frequency and that zero voltage corresponds to the negative peaks of the output waveform
(a) V_{out} (across R_2) (b) V_{in} (across R_1)

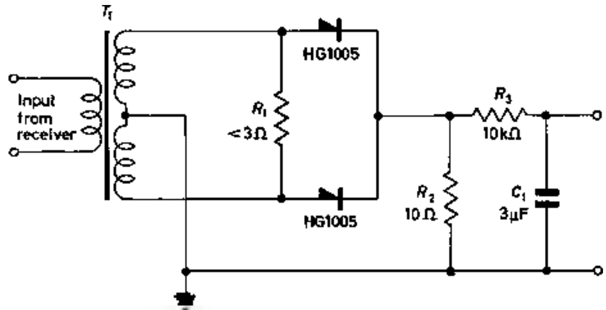


Fig. 5. A simple practical arrangement.
(R_1 can, of course, be connected across the primary winding of T_1 if more convenient)

Fig. 4 shows an oscillogram of a dual trace display of the input voltage and the voltage across R_2. The zero voltage lines have been drawn in, showing that, for sine waves, V_{out} has a mean d.c. level of half the peak-to-peak output voltage; and it is this d.c. level that is proportional to the mean power. A very simple practical square-law detector can, in fact, be constructed by simply adding a suitable smoothing filter to the basic arrangement.

Fig. 5 gives the circuit of a detector unit which can be used in conjunction with a high-sensitivity oscilloscope. A sine wave input (across the complete secondary of T_1) of 1·6V peak gives a d.c. output of just over 25mV. A.C. components above about 100Hz are effectively suppressed by the simple filter composed of R_3 and C_1. Used with an oscilloscope having a Y sensitivity of 10mV/cm, such a circuit is capable of producing a 25cm beam deflexion without appreciable error.

Increasing the Output Voltage

Although the simple basic arrangement is quite usable

—providing a high-sensitivity oscilloscope is available—it is obviously desirable to increase the output voltage so that full screen deflexion can be produced on an oscilloscope with the more common 50mV/cm sensitivity.

One solution to the problem is, of course, the inclusion of a d.c. coupled amplifier between the detector output and the Y input terminal of the oscilloscope. But such an amplifier is not too easy to construct, bearing in mind that it forms part of a measuring system and that care must be taken to ensure that it has a high standard of gain stability and freedom from d.c. drift.

A much simpler solution is available, however. Instead of amplifying the d.c. component of V_{out}, one can amplify the a.c. component only and then use a d.c. restoring diode to bring the negative peaks of the waveform back to zero voltage. A circuit arrangement for doing so is given in Fig. 6.

A Marconi Instrument TM 6951 oscilloscope pre-amplifier, used in the experimental detector arrangement, gives a voltage gain of 100, producing a maximum peak-to-peak output of 5V. At this operating level any non-linearity in the d.c. restoring diode has negligible effect; but, to avoid the possibility of error due to loading, a high impedance RC filter network was used for elimination of the a.c. from the final output to the oscilloscope.

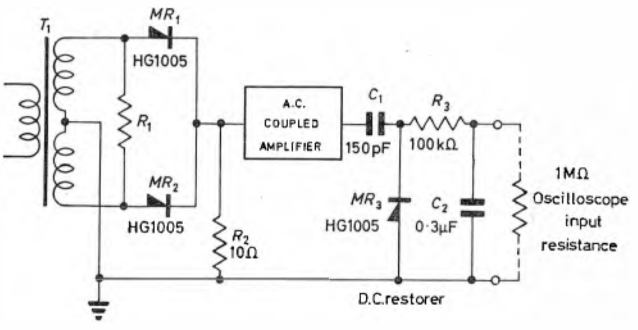


Fig. 6. Circuit arrangement for using an a.c. coupled amplifier to increase the output voltage.

(Suitable for noise or modulation frequencies up to about 5kHz)

Tests and Adjustments

The purposes for which this detector is primarily intended do not, in the main, call for an absolute transfer calibration in terms of volts out for watts in. But such a calibration is very easily made and, if plotted on linear graph paper, gives a very good indication of the fidelity of the unit's square-law characteristic.

For initial setting up, however, a quicker method is needed of checking the transfer law; and the most convenient way is by examination of the waveform across R_2—or, for greater sensitivity, the output waveform of the amplifier—on an oscilloscope. With a sine wave input voltage, the voltage across R_2 should take the form of a cosine wave at twice the input frequency; see Fig. 4.

(1) *Balance*

If the output waveform is of the shape shown in Fig. 7, with a high content of the input frequency, the circuit is unbalanced—either in transformer T_1 or the diodes. The diodes are the most probable cause, and it is advisable to select an approximately matched pair. This is a fairly simple operation; for they can be matched with sufficient accuracy by merely selecting a pair having the same forward conductance with about 1V input potential; i.e. by measuring the forward conductance with an Avo Meter.

Final balance can be obtained by means of a 100Ω potentiometer connected as shown in Fig. 8. The shunt connexion is more convenient in the low-impedance input circuit than the theoretically correct series arrangement with the balancing potentiometer between the two halves of the transformer secondary winding. In practice, there is always sufficient loss in the transformer to enable a small amount of adjustment to be made by differentially varying the loading on the two secondary halves.

The procedure for finally balancing the detector is simply that of applying a sine wave input at a level of about 1200mV peak across R_1 and adjusting the balance

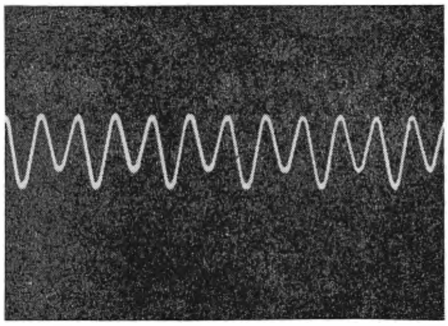


Fig. 7. Oscillogram of the waveform across R_2 with the detector unbalanced

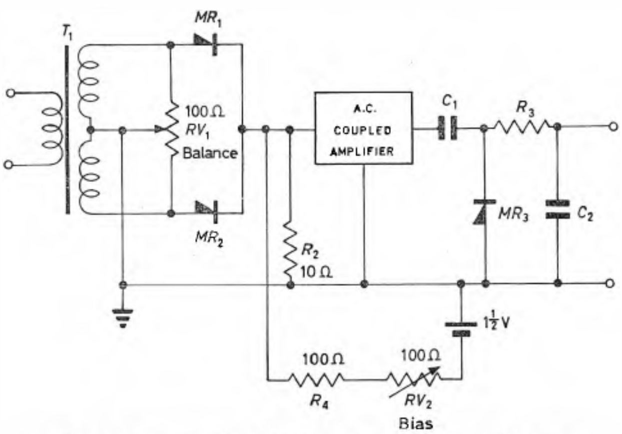


Fig. 8. Arrangement for adjustment of balance and bias

control to produce a perfectly even sine wave output as viewed on the oscilloscope.

(2) *Bias*

At fairly high input levels—above 1V peak across R_1—the error due to the displacement of the diode's square-law characteristic can be neglected. But, for signal-to-noise ratio measurements without the use of an attenuator, it is usually necessary to measure the noise power at a very much lower level. So, for maximal accuracy, a suitable biasing system is desirable.

Fig. 8 shows the most convenient method—as applied to the detector with an a.c. amplifier. The bias voltage is developed across R_2 by the use of a second variable resistor, RV_2, a fixed resistor, R_4, and a 1·5V cell. The level of bias to be developed is normally of the order of 75 to 100mV. Thus, by making the value of the series resistance some 15 times that of R_2 adjustment can be made without materially altering the effective transfer constant —kR_2.

The method of setting the bias voltage for minimum distortion is similar to that for balance adjustment in that a sine wave test waveform is used and the a.c. output voltage is displayed on an oscilloscope.

Distortion due to incorrect bias is not normally detectable for input voltages greater than about 400mV peak. This really means that the need for adjustment applies only when the detector is used in conjunction with an amplifier.

For bias adjustment the input voltage is reduced to the lowest level that will produce a readable sized display, and the bias control is then set to produce the best sine wave. More elaborate methods of measuring the distortion of the sine wave output are unnecessary and impossibly tedious. The effect of applying the bias voltage with a low-level input is shown in Fig. 9.

(a)

(b)

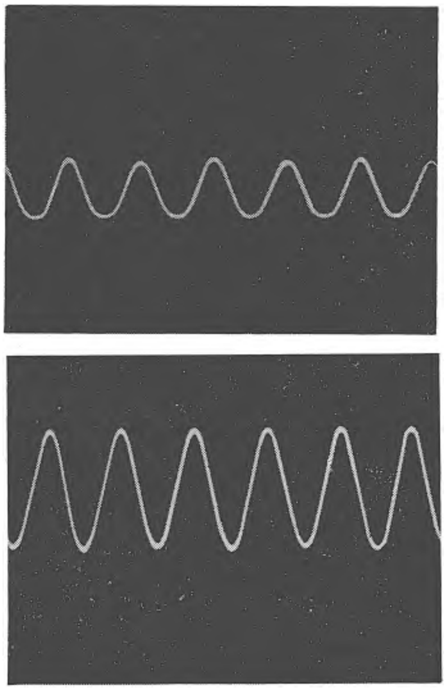


Fig. 9. Waveform of voltage across R_2 with and without bias applied
(a) Waveform without bias
(b) Waveform with bias

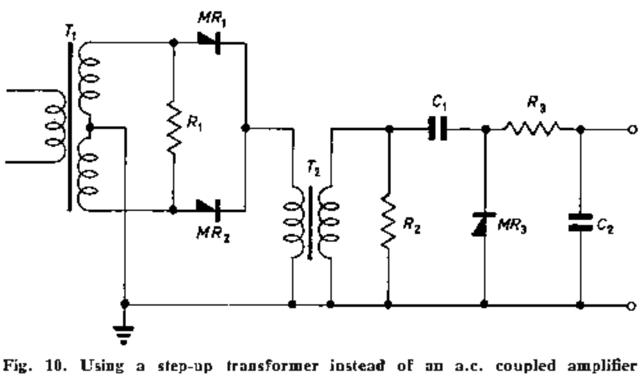


Fig. 10. Using a step-up transformer instead of an a.c. coupled amplifier

An Alternative Arrangement

Having established the principle that one can utilize the a.c. output from the square-law detector and then restore the d.c. with a third diode, one can, if it is more convenient, use a transformer to increase the output voltage instead of the a.c. amplifier.

Fig. 10 shows a basic arrangement in which transformer T_2 provides the voltage step-up. If this transformer has a turns ratio of 100:1, for example, a 10kΩ loading resistor, R_2 can be used, giving an effective diode load of 10Ω.

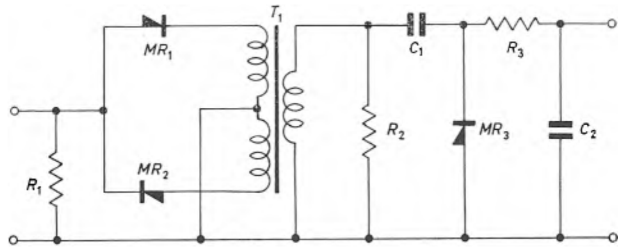


Fig. 11. Arrangement with one transformer only

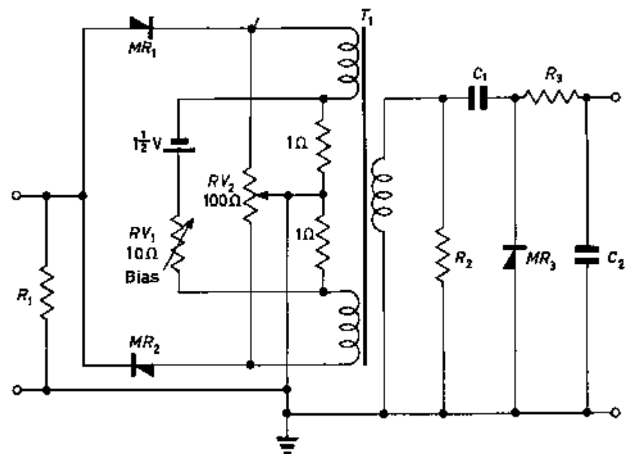
Although this arrangement is quite sound, it offers little advantage over the amplifier system—for it is often a good deal easier to build a transistor amplifier than to wind a suitable transformer. So, if a transformer is to be used in this circuit position, there is more to be gained by re-arranging the diode connexions and eliminating T_1. Many receivers have a low-impedance unbalanced a.f. outlet which connects readily to the simple two-terminal inlet of the detector shown in Fig. 11.

In this circuit, one of the diodes is reversed, so that positive half cycles are applied to one half of the transformer's primary winding and negative half cycles to the other. The voltage step-up ratio is, of course, the turns ratio between each primary half and the secondary. And it must be remembered that the potential developed across R_1 is the actual diode voltage—not twice the voltage as for the original circuit. Final balancing can be effected by means of a shunt potentiometer across the centre-tapped primary winding of the transformer. The same arguments apply as for the arrangements shown in Fig. 8.

The application of the bias voltage is not quite so straightforward, however, because of the reversal of one of the diodes. Fig. 12 shows a method of developing the d.c. bias voltage across a pair of 1Ω resistors introduced between the two halves of the primary winding of the transformer. Such low value resistors involve a rather high standing current of about 100mA; and one is tempted to reason that it would be better to develop the voltage across resistors of higher value, with shunt capacitors to reduce the fixed a.c. impedance.

A difficulty in doing so arises from the fact that the order of capacitance required necessitates the use of elec-

Fig. 12. Provision for bias and balance adjustment
RV_1 adjusts the bias and RV_2 adjusts the balance



trolytic capacitors. These have the disadvantage that their effective series loss usually rises with frequency and can amount to the order of two or three ohms at frequencies which are still too low for solid dielectric capacitors. Suitable capacitors are, no doubt, available; but, as they were not at hand when the experimental detector was built, the simpler purely resistive circuit was used and found to give satisfactory results.

Final Considerations

The overriding virtue of these detector circuits is their extreme simplicity; and the additional accuracy achieved by the use of the refinements of balancing and bias control may well be unnecessary for a large number of applications.

Signal-to-noise measurements on receivers are frequently made by conventional methods using indicators which respond to average current or voltage. Even in its simplest form, the square-law diode detector gives a more accurate measure of relative power than indicators of this type. Commercially available a.f. power meters usually offer facilities which are quite unnecessary for these measurements yet fail to indicate true power for non-sinusoidal waveforms.

It can, therefore, be argued that the detector in its simplest form can be used, with some advantage, as an input circuit to a d.c. or peak-reading a.c. electronic voltmeter. Alternatively, the circuit shown in Fig. 8 feeding a high-impedance multi-range voltmeter, such as an Avo Model 8, can provide a relative power measurement accuracy equal to that of an expensive true r.m.s. electronic voltmeter.

A New Photoconductive Television Camera

The Marconi Company has recently announced an entirely new television camera, the Mark VI, designed to provide the ultimate in performance from a photoconductive black-and-white television camera.

The Mark VI, in its basic form, is designed for telecine operation, using a Vidicon tube to provide the highest possible quality in this role.

This basic unit can then be built up into a studio or outside broadcast camera, using a number of additional units. The camera tube will be changed in these applications, for the Plumbicon. In this form, the Mark VI will be suitable for a wide range of tasks in broadcast studios, providing excellent pictures in all types of application, especially where cost and engineering simplicity are vital considerations.

The circuits associated with photoconductive tubes are considerably simpler than those required by the image orthicon tube, which still provides the highest possible picture quality yet achieved with any type of camera in studio or outside broadcast use. Increased simplicity, however, reduces both capital and running costs, and the Mark VI will provide very high quality pictures in cases where the sophistication of the Marconi Mark V, 4½in image orthicon camera are not essential.

Many standard features from the other cameras in the Marconi range have been carried through to the Mark VI, to provide the maximum compatibility. The camera uses similar types of zoom lenses to the Mark V, 4½in image orthicon camera and the Mark VII colour camera. The tilting viewfinder, pioneered by Marconi, has also been used on this camera, and again, this is identical with the viewfinder on the Mark V and Mark VII cameras. A standard camera cable has been used, and is common to the Mark V and its predecessor, the Mark IV.

A particular feature of the Mark VI is its very high sensitivity. Perfect pictures can be obtained at less than 50 foot-candles, less than half that of a normal black-and-white studio.

The camera is completely solid state throughout, with the exception of a single Nuvistor which has been used in the first stage of the head amplifier to ensure the best possible signal-to-noise ratio. In normal operation, this will be at least 46dB with a video bandwidth of 5·5MHz.

Careful design of the camera optical system and the tube yoke have ensured a very high standard of picture geometry. The distance between any two points on a test chart will be reproduced by the camera within 2 per cent of the correct distance, even at the corners of the picture.

The camera has been designed for 'hands-off' operation. The complete transistorization of the camera enables rehearsal quality pictures to be obtained one minute after switching on. After a thirty minute warm-up period, the output will be stable for a constant light input, and will remain stable with ambient temperature variations of $\pm 10°C$ and with mains variations of ± 6 per cent. This performance will be maintained throughout a full working day, and makes it possible to relegate all electronic controls to the equipment racks, where they can be treated as preset controls.

Variations in light level will normally have to be compensated for, and this can be done with a small operational control unit in the production suite, incorporating 'Iris', 'Gain', and 'Black level'.

The main structural framework of the camera is made from magnesium alloy plates, which provide very great strength and rigidity with the minimum weight penalty. The basic camera in its telecine role, is contained in a unit measuring 5in high, 15·5in long and 9in wide. This unit would normally be mounted vertically in a multiplexer, for use with a number of film and slide sources, or coupled directly to a single film or slide projector.

The studio camera is built up from this basic unit by the addition of an upper housing which contains the viewfinder power supplies and talk back circuits, and provides the mounting for the tilting viewfinder. The zoom lens package is attached to the front of the camera and can contain all the control circuits necessary for full servo control from a preselecting zoom 'shot-box'.

For outside broadcast use, provision has been made to accommodate zoom lenses with a larger relative aperture, to cater for the possibility of low light levels. These lenses are rather heavier and the mechanical design permits the addition of a longer mounting shoe, which moves the balancing point of the pan and tilt head forward to the centre of gravity of the combined camera/zoom combination.

Modular construction has been used throughout the camera and control unit circuits, to provide the maximum ease of maintenance. The camera control unit is housed in a standard 19in modular frame, suitable for rack mounting or for a mobile carrying case. All the engineering and operational controls which are needed for setting up and operating the channel, are contained in a power supply and camera control module in this camera control unit.

A feature of the camera is the gain compensated black level control, which makes it possible to alter the black level of the picture without altering the peak white level. This feature, which was first introduced on the Mark V camera, provides a useful simplification of the operational control of the camera.

A unique gamma correction circuit is used, which has been patented by The Marconi Company. It provides a continuous variation in gamma from unity to 0·4 with a single control, and without any need to adjust the video level. The circuit can accept a linear input signal with a contrast range of 30 : 1.

The same detachable tilting viewfinder is used as the Mark V and Mark VII cameras. This feature, which was pioneered by Marconi on the Mark III camera, enables the cameraman to see the viewfinder easily, whatever the attitude of the camera itself. This viewfinder can be removed and operated up to about 30ft from the camera itself if necessary.

The viewfinder position is adjusted by means of friction clamps, and the extremely high brightness of 200 foot-Lamberts is able to compete with the brightest studio lighting. Controls for the 7in tube are mounted directly below the tube face, with cue lamps on either side. An input selector enables either the camera picture or an external picture to be displayed.

A Simple Frequency-Modulation Tape Recorder System

By W. Tempest and M. E. Bryan*

A transistorized frequency modulation unit is described, which can be used with any good domestic or semi-professional tape recorder to record and replay signals over a frequency range from d.c. to 1 250Hz. The system has a signal-to-noise ratio of 42dB, and the stability is such that over a 20min period an a.c. signal reproduced by the system will not change by more than 0·1dB, and a d.c. level will typically drift about 5mV. The input signal range is ±5V and the output some 20dB lower at ±0·5V.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

When it is to be used in any application involving the precise measurement of signal levels, or the recording of slowly varying signals, the magnetic tape recorder suffers from a number of disadvantages. Chief among these are its inability to cope with changes occurring at less than about 20Hz and the fact that the magnitude of the output depends directly on the characteristics of the recording tape. In order to overcome these limitations of 'direct' recording, recorders using some form of modulation, such as frequency, pulse width, or code, are widely used in data handling and computing. However, the machines used are in a different category of cost from the semi-professional and the better domestic recorders often used in laboratories. It therefore seemed possible that there might be some interest in an instrument which in terms of cost and complexity lies between the two classes mentioned above, but which at the same time provides many of the advantages of the higher-priced instrument without including all its refinements. Such an instrument has been developed in the form of a frequency modulation unit which can be used with any good domestic type machine and which should have a cost comparable to the recorder itself.

Design Parameters

In a frequency modulation system the two basic design parameters to be chosen are the carrier frequency and the modulation swing. These two factors in turn control the design of the filter used to separate carrier from modulation in the demodulator, and in this way decide the upper frequency limit of the system. When magnetic recording is involved, the carrier frequency is limited by the highest frequency that can be recorded on the tape. At the same time, the 'wow' present in the tape transport system becomes a source of noise in the output, and its effect must be minimized by making the frequency deviation as large as possible. Taking these factors into account two alternative carrier frequencies were considered, 5kHz and 10kHz. The latter frequency, with a deviation of ±30 to 40 per cent, is the highest practical value for the recorder used (a Ferrograph 4 CFN/H at 7½in/sec). The lower alternative frequency was considered in the knowledge that the recorder output was considerably more uniform at 5kHz, and it therefore seemed possible that this speed would give a better signal-to-noise ratio. This was later confirmed experimentally and 5kHz was chosen as carrier. Calculation of the requirements of a low-pass filter to eliminate the carrier in a demodulator showed that it would require an attenuation of at least 60dB throughout the frequency range of the carrier. In order to attain this performance in a simple design a cut-off frequency of one quarter the carrier frequency was chosen.

The Modulator

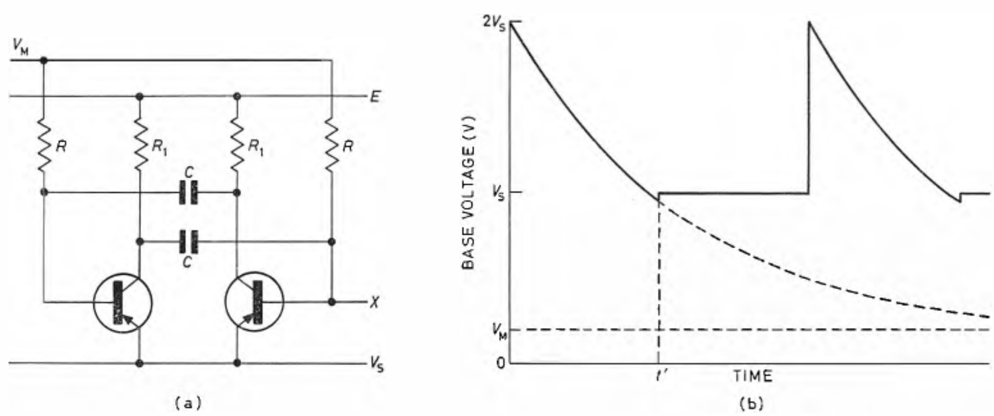
The requirements for the modulator are, a high degree of linearity, a wide frequency deviation, and simple construction, together with a range of modulation frequencies up to 25 per cent of the carrier. The only class of circuits possible seems to be relaxation oscillators and the circuit chosen is a symmetrical astable multivibrator in which the frequency is controlled by varying the voltage applied to the timing resistors. Fig. 1(a) shows the basic circuit and 1(b) the timing waveform at the base of one transistor. In this circuit the time t' (half a cycle of oscillation) depends on the value of RC and on the applied voltage V_m. The frequency dependence is not linear because the timing waveform is exponential having the form

$$v = V_m + (2V_s - V_m)e^{-t/RC}$$

where V_s is the supply voltage.

The circuit will 'flip' when $v = V_s$ at a time t'. The frequency of oscillation $f=1/2t'=1/2RC \ln (2V_s-V_m)/$

Fig. 1 (a). Basic circuit of the modulator
(b). Timing waveform at point X



* Formerly University of Liverpool, now at Royal College of Advanced Technology, Salford.

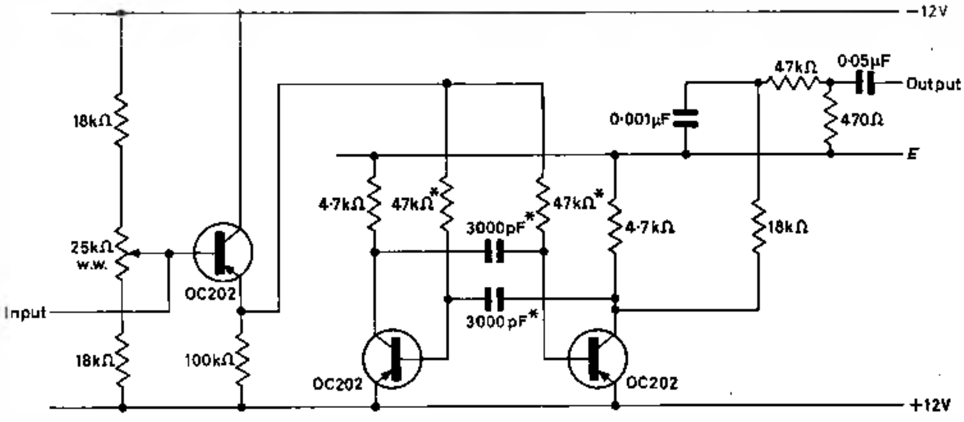


Fig. 2. Complete circuit of the modulator
*Components marked * ±2 per cent, others ±10 per cent*

$(V_s - V_m)$. This equation shows that for V_s of 12V a modulating input of ±5V will produce a ±31 per cent deviation in frequency with a non-linearity of 0·45 per cent. (Non-linearity being here defined as the maximum departure of the frequency from the best fit straight line over the modulation range, divided by the peak-to-peak frequency deviation). The complete modulator is shown in Fig. 2 and measurements on this circuit show, that for the supply and modulating voltages quoted above, the frequency deviation is 34 per cent and the non-linearity about 0·4 per cent. This suggests that the simplified analysis, which ignores the effects of base current, leakage current and transition time, on the frequency of oscillation, is an adequate approximation. In the practical circuit, base current and leakage current, and also the effect of temperature changes, are minimized by the use of high gain silicon transistors. The multivibrator circuit provides an output signal of about 11V peak-to-peak and the output circuit is designed to reduce the harmonic content of this signal and attenuate it to about 50mV level to avoid overloading the microphone input of the tape recorder.

The Demodulator

In view of the large frequency deviation, the obvious choice for a demodulator is a counter-type circuit in which each cycle of the carrier is converted to a pulse of constant amplitude and duration. It is then only necessary to pass the signal through a low-pass filter to obtain the demodulated output. The complete circuit is shown in Fig. 3. The first stage consists of a diode limiter, which is followed by an amplifier and then a Schmitt trigger circuit. This system converts the input into a square wave with a rise time of about 1μsec. The output of the Schmitt is then used to trigger a monostable which delivers an output pulse of about 10V peak amplitude and 70μsec duration to the filter. The monostable is designed to give a stable pulse in which the levels correspond to hard-on and cut-off of the transistor. An emitter-follower connects the monostable to the filter which is of the m-derived type with matching impedances of about 10kΩ. The filter included in Fig. 3 has a cut-off (−3dB) frequency of 1 250Hz and a maximum attenuation at 5kHz. The upper frequency limit (i.e. the highest instantaneous carrier frequency) of the demodulator as shown in Fig. 3 is about 8kHz, this being the highest rate at which the monostable will operate.

Performance

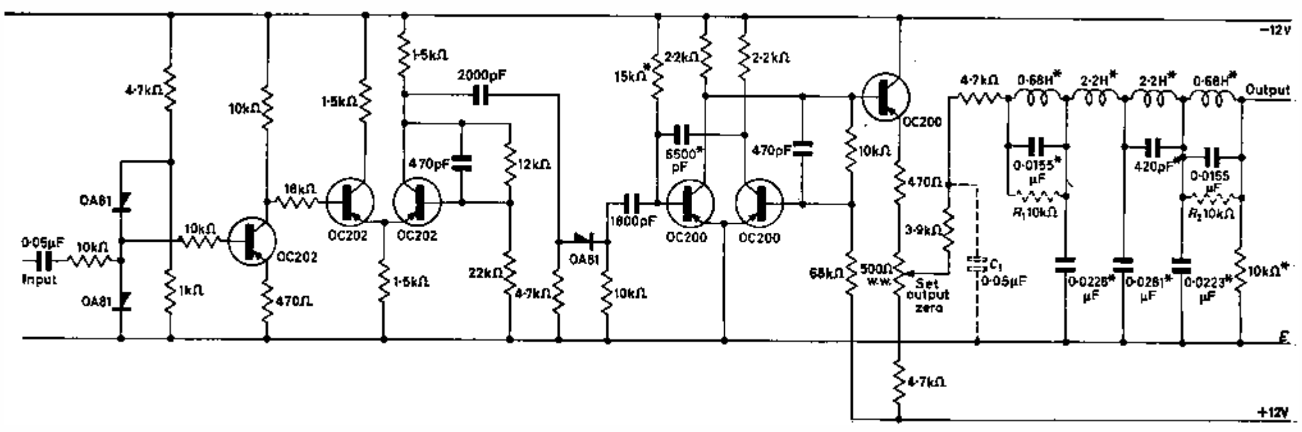
Four aspects of the overall performance of the system were measured. These were, a.c. response, d.c. response, transient response, and stability. Except where otherwise stated performance measurements were made using a 5kHz carrier and 7½in/sec tape speed.

(a) A.C. Response

The results of measurements of the frequency response of the system are shown in Fig. 4. 4(a) shows the response with the filter circuit as in Fig. 3, while 4(b) shows the modified response obtained by including the extra components R_1, R_2 and C_1 shown dotted. This modification was designed to improve the transient response and is explained in section (c).

A broad-band (2Hz to 200kHz) electronic voltmeter was used to measure the levels of maximum output and background noise, and the signal-to-noise ratio was found to be 42dB. It may be noted here that a modified demodulator and filter, designed to use a 10kHz carrier, and to

Fig. 3. Complete circuit of the demodulator
*Components marked * ±2 per cent, others ±10 per cent*



handle modulation up to 2·5kHz, was set up and tested and found to have a signal-to-noise ratio of 37dB. The difference of 5dB in noise level between the 5kHz and 10kHz systems was sufficient to cause an obvious degradation of the quality of the output on a cathode-ray oscilloscope, and the lower carrier was therefore chosen as the more satisfactory.

(b) D.C. Response

To measure this aspect of performance d.c. levels at one volt intervals from +5 to −5V were recorded and played back. The results of this test showed that the maximum departure from linearity was 0·5 per cent at the limits of the output range.

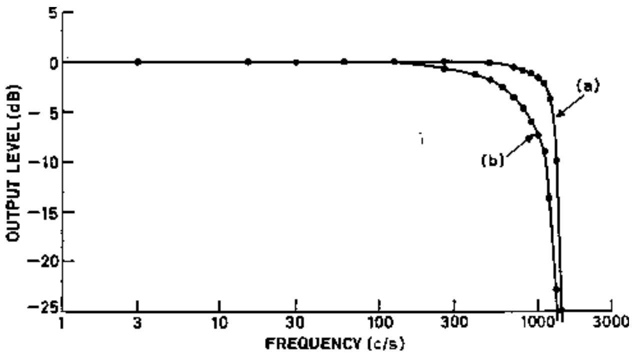


Fig. 4. Overall frequency response of the system
(a) With normal filter
(b) With filter modified to reduce overshoot

d.c. input level. In the a.c. test, which was for 30min, the output level remained steady within 0·1dB (the limit of measurement). In the d.c. tests, which were carried out over several periods of 20min, the maximum drift recorded was 7mV, the typical drift over this period being about 5mV. The total output excursion was approximately ±500mV.

(e) Other Factors Affecting Performance

Two brands of magnetic tape were used in the tests, Scotch "150" extra play, and Emitape "88" standard play. There was no detectable difference between these for this application. It was found that the use of an old and rather roughly handled sample of tape led to a substantial deterioration in performance in the form of a large increase in noise level.

The choice of level for the recording of the carrier may seem at first sight to be of little importance. It was found, however, that the best results were obtained by recording in the range 0·25 to 0·5 of the normal maximum level. If the signal was recorded at or above maximum level, a sharp increase occurred in 'noise behind the signal', due apparently, to beating between the tape bias oscillator frequency and harmonics of the carrier. This effect could of course be eliminated by removing the bias signal completely, but this gave little, if any, improvement over normal recording at a fairly low level.

Construction

The modulator, demodulator, and power supply, were built in a single instrument case 12in × 7in × 6in. There were no special features in the construction or layout, but it was found impossible to operate the modulator and demodulator simultaneously from the same supply. A two pole changeover switch was therefore used to connect the supply to which ever section was needed.

No attempt was made to estimate the total cost of construction but the cost of components for the prototype (including the case) was less than £30.

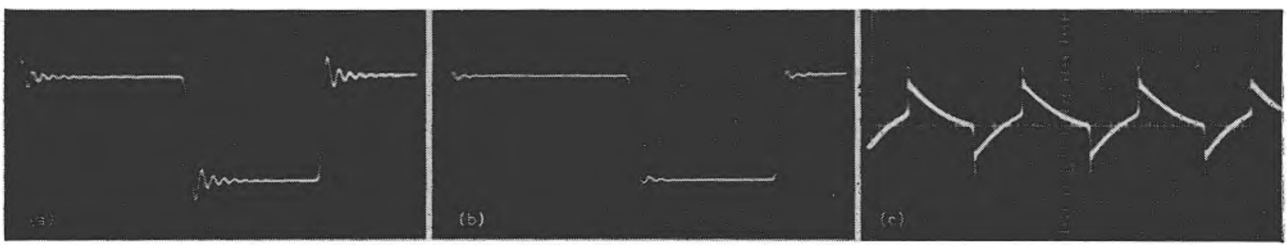


Fig. 5. Square waves reproduced by the system
(a) 40Hz with normal filter
(b) 40Hz with filter modified to reduce overshoot
(c) 40Hz square wave recorded and reproduced using the tape recorder "direct", without the f.m. system

(c) Transient Response

The transient response was measured by recording and replaying square waves of 40Hz, with the result shown in Fig. 5(a). The overall frequency response of the instrument is controlled entirely by the low-pass filter in the demodulator, and since this has a sharp cut-off the transient response is oscillatory. This is not really satisfactory for the reproduction of waveforms containing transients, and to overcome this difficulty a modification to the filter is needed to reduce the steepness of the response. The addition of the components C_1, R_1 and R_2 (shown dotted in Fig. 3) makes the required change, and improves the square wave response to that shown in Fig. 5(b). The sine wave response is modified to the curve (b) in Fig. 4. For the sake of comparison, Fig. 5(c) shows the same 40Hz square wave recorded and replayed 'direct', without the frequency modulation system.

(d) Stability

The output stability was measured from two points of view, for a constant a.c. input at 20Hz, and for a constant

Conclusions

A frequency modulation unit has been described, which can be used with any good domestic or semi-professional tape recorder to record and replay signals over a frequency range from d.c. to 1 250Hz. The system has a signal to noise ratio of 42dB and the stability is such that over a 20min period an a.c. signal will not change by more than 0·1dB and a d.c. level will typically drift about 5mV. The input signal range is ±5V and the output is some 20dB lower at ±0·5V.

Some Approaches to Photomultiplier Power Supply Design

By R. Barlow*, B.Sc., and J. Watson†, Ph.D., M.I.E.E.

The power requirements of photomultipliers are discussed and it is shown that the associated circuits and the mode of operation of these devices can sometimes permit relaxed specifications for the relevant power supplies. The design of high-voltageinvertor transformers and their associated stabilizing circuits is considered and the advantages to be gained by the use of f.e.t. comparator stages are presented.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

FOR most photomultiplier applications, power supplies delivering from 500V to 2kV at a few milliamperes are required. In counting and pulse height discrimination work, such as is usually found in nucleonic instrumentation, the voltage has been traditionally positive-live, negative ground; whereas in photometric, spectroscopic and allied fields where an average irradiation level is measured, the converse, or negative-live polarity is used. This has led to some e.h.t. units having a floating output being offered, but since the two classes of user tend to select quite different instrument types for other reasons, the following account will be restricted to the single-ended, negative-live supply, because the work performed has been carried out with this particular application in mind[1].

Photomultiplier tubes require that each dynode be held at fixed potential with respect to the adjacent dynodes, and this is normally achieved by the use of a chain of dynode resistors as shown in Fig. 1. The current passing down this chain, I_D, must be considerably greater than the anode, or signal current I_A if good linearity of I_A with respect to the irradiation level is to be attained. In general I_D is 100 to 1 000 times I_A.

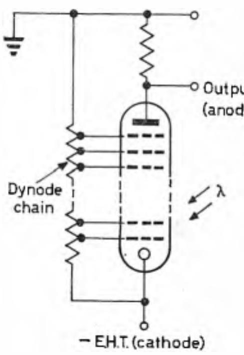
Also, the actual value of I_A must be kept very low if fatigue effects are to be minimized[2]. Most photometric instruments are designed so that I_A does not exceed 10μA, and a maximum of 1μA is common. This implies an I_D of 0·1 to 1·0mA.

These factors lead to a significant relaxation in the specification of the relevant power supply provided that the dynode chain is permanently connected to this supply. This is because the load presented is essentially constant, and the load stability factor for the supply can therefore be much lower than would otherwise be acceptable. This is a point which is valid only for custom-built supplies, for it is hardly likely that a commercial unit quoting an apparently poor load stability would gain commercial acceptance.

The mains variation and temperature stabilities must, of course, remain high.

In some forms of spectrometric instrument, it is necessary for the e.h.t. supply voltage to be capable of control by an incoming signal. For example, in some instruments employing the double-beam-in-time light chopping principle[3] it is required that the half-cycle corresponding to the reference beam be delivered to the measuring system at a constant magnitude so that the 'sample' half-cycle may be compared with this constant value. This implies that the photomultiplier/pre-amplifier channel should have a variable-gain capability, and this may be achieved by varying the e.h.t. supply.

On the other hand, if a fixed e.h.t. supply is utilized, the variable gain facility may be provided by the pre-amplifier itself. A particularly simple way of achieving this is illustrated in Fig. 2 where a f.e.t. is used below pinch-off to provide a purely resistive load whose value is a function of the gate-source voltage[4]. The p-channel f.e.t.'s currently available have channel resistances which rarely exceed 10 to 20kΩ at $V_{GS} = 0$, this implying a voltage drop of only 0·1 to 0·2V at $I_A = 10\mu$A. If a system is en-



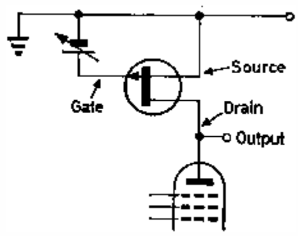


Fig. 1 (left). Photomultiplier connexion
Fig. 2 (right). Gain control using a p-channel f.e.t. as a voltage variable load

visaged wherein the signal must be measured down to 1 per cent of its maximum value, then this involves voltage levels of the order of 1mV, which approaches the noise level for such systems. Thus, the method is more applicable where the higher values of I_A are permissible.

A suitable transistor for this circuit is the Union Carbide UC40, which has a very low drain-source capacitance and whose value of R_{DS} will range over several decades, the lowest value being about 10kΩ. Similar to this is the Siliconix 2N3112.

The foregoing paragraphs indicate that there is a requirement for both fixed and variable gain e.h.t. supplies, and it will be shown that both types can use the two-transistor invertor as a source of the high voltage involved. The design of the relevant transformer will be considered later; in order to discuss voltage stabilizing circuits, it will be assumed that the invertor takes the form shown in Fig. 3 and that the output voltage is a function of the input current.

A Simple Fixed-Voltage E.H.T. Supply

Recently, the Victoreen 'Corotron' has become available in the U.K.*, this being a sub-miniature corona stabilizer which lends itself to the simple circuit of Fig. 4.

* *Advance Electronics Ltd.*
† *University of Wales, Swansea*

* *Imported by Walmore Ltd, Betterton St, London.*

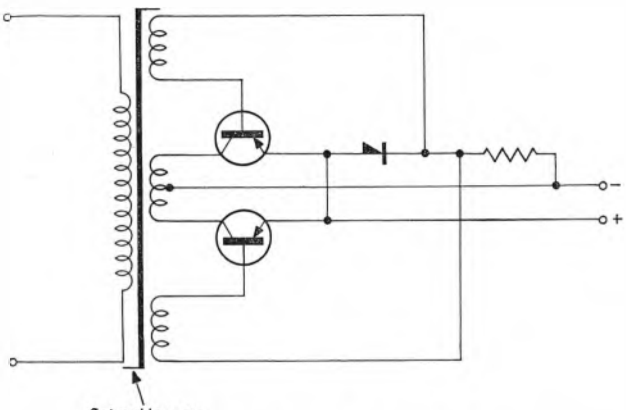


Fig. 3. Invertor with starting diode

The current which the Corotron may pass is both small and limited, ranging from $5\mu A$ to $100\mu A$. This means that when the input voltage V_{in} is at its lowest anticipated value, and I_A is at its maximum, the Corotron must still pass $5\mu A$ or more, otherwise it will cease to stabilize. These conditions can be represented as follows:

From $I_S = I_{in} - I_L$

$$I_{S(min)} \geqslant \frac{V_{in(min)} - V_S}{R} - \left[\frac{V_S}{R_D} + I_{A(max)}\right] \ldots\ldots (1)$$

where R_D is the resistance of the dynode chain.

The converse case occurs when V_{in} is maximal and I_A is minimal, under which conditions the Corotron must not be allowed to pass more than $100\mu A$:

$$I_{S(max)} \leqslant \frac{V_{in(max)} - V_S}{R} - \frac{V_S}{R_D} \ldots\ldots (2)$$

If $V_{in(max)}$ and $V_{in(min)}$ are known as numerical values, then expressions (1) and (2) can be solved to give a range of values within which R must lie. Otherwise, if the maximal and minimal values of V_{in} are known only as percentages of a nominal value, expressions (1) and (2) may be treated as equalities and solved as simultaneous equations.

An example will illustrate this.

If V_{in} is able to vary over the standard range +10 per cent and −15 per cent of nominal, and the following parameters are involved, then equations (1) and (2) can be solved for a specific case:

Let V_S = 1250V (i.e. a GV1A-1250 is used)

Let the photomultiplier be a ten stage type as exemplified by the 931A series. Here, convenient dynode resistors will be $1M\Omega$ each and $I_{A(max)}$ will be $1\mu A$. Thus, $R_D = 10M\Omega$.

$$\text{Then}\quad 5 = \frac{0{\cdot}85V_{in} - 1250}{R} - \left[\frac{1250}{10} + 1\right] \ldots (1a)$$

$$\text{and}\quad 100 = \frac{1{\cdot}1V_{in} - 1250}{R} - \frac{1250}{10} \ldots\ldots (2a)$$

The solutions of these equations give V_{in} = 2490V and $R = 6{\cdot}63M\Omega$.

If such a power supply is so installed that it is integral with the dynode resistors grouped around the photomultiplier socket, then it can be operated even if the tube has been extracted. However, if it is possible to disconnect the dynode chain, a current of at least $(2490\text{-}1250)/6{\cdot}63 \simeq 187\mu A$ can flow, which may damage the Corotron.

The variation in V_S over the full working range of the Corotron is given as 8V which implies a transfer ratio R_T of:

$$R_T = \frac{\Delta V_S}{\Delta V_{in}} = \frac{\Delta V_S}{(1{\cdot}1 - 0{\cdot}85)V_{in}} = \frac{8}{622{\cdot}5} = 1 : 77{\cdot}8$$

An actual test circuit using the calculated parameters performed considerably better than this, as is seen in Fig. 4(b).

The load stability is immaterial, for the load current can change by only $1\mu A$ in $125\mu A$. The measured noise was better than 0·1V peak-to-peak even without a smoothing capacitor across the Corotron.

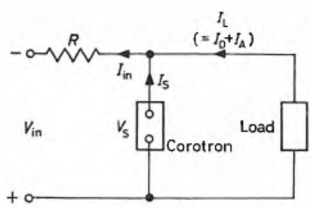
The values of both R and V_{in} are rather high in the above example, and if an invertor/rectifier unit is used as the e.h.t. source, these values can be reduced by inserting some regulation at the primary. For example, a ±5 per cent Zener regulator might be used to define the input voltage, when, for a constant load, the output voltage will also be within ±5 per cent of nominal. In other words, the invertor/rectifier unit can be regarded as a d.c. transformer.

For a ±5 per cent variation in V_{in}, equations (1) and (2) give $R = 1{\cdot}4M\Omega$ and V_{in} = 1320V. It will also be found that there is no danger in disconnecting the dynode chain under these conditions.

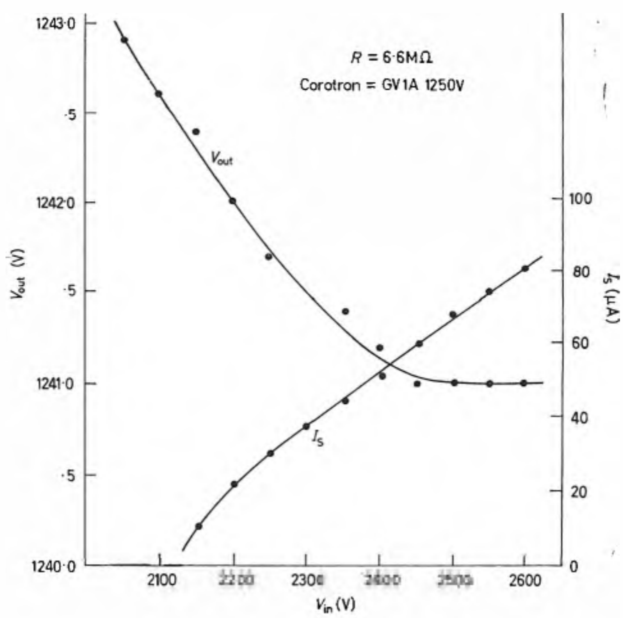
Stabilizing Circuits for Invertor Control

In the event of a variable-voltage, highly stabilized supply unit being required, it immediately becomes necessary to compare the output voltage with a reference, and to amplify any disparity, so producing a control signal which can be applied to the invertor. In general, the high

Fig. 4. (a) A simple stabilized e.h.t. supply
(b) Performance of simple e.h.t. supply



(a)



(b)

voltage must be sampled, and the obvious way of achieving this is to utilize a dropper chain as shown in Fig. 5. Here, a fraction R_R/R_T of the output voltage is compared with the reference, which means that the apparent loop gain of the stabilizer has immediately been reduced by this factor. Clearly, R_R/R_T can be very small; for example, if a 5·5V reference is used (as is provided by a Zener diode with a minimal temperature coefficient) and the maximum output voltage is 2000V, then

$$\frac{R_R}{R_T} = \frac{5{\cdot}5}{2000} \simeq 0{\cdot}0027$$

Another disadvantage of the dropper chain becomes apparent if a bi-polar input transistor is used in the comparator amplifier. Under these circumstances, the base current of this transistor must be supplied via the chain itself. Such a situation is exemplified by the long-tailed pair comparator circuit in Fig. 6. Here, the reference voltage is derived from a cascaded Zener diode network and compared with the voltage dropped across R_R.

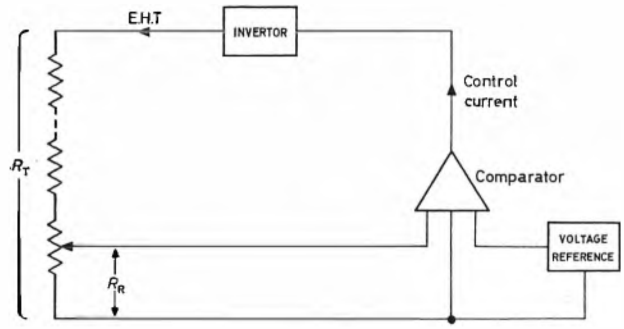


Fig. 5. Basic stabilizer system with invertor in control loop

Clearly, the base current of $VT_{1(a)}$ must flow into the dropper chain so that

$$I_{Ch} = I_{RR} + I_{B1(a)} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

Since R_R and $R_{in(VT1)} \ll R_T$, then,

$$V_{eht} \simeq I_{Ch} \,.\, R_T \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Consequently, unless $I_{B1(a)} \ll I_{Ch}$, any changes in this base current will have a significant effect upon the e.h.t. voltage, and this can be calculated using equations (3) and (4).

In point of fact, for the transistor shown in Fig. 6, h_{FE} may be as low as 10 when $I_{C1} \simeq$ 1mA, which implies a base current which forms about 10 per cent of I_{Ch} in this circuit. The base current can change as a result of three major causes:

(a) A change in V_{eht} brought about by adjustment of the voltage control RV_1.

(b) A change in load current.

(c) A change in temperature.

The change in case (a) is due mainly to a modification of the quiescent currents in VT_1, but there will be some contribution by the out-of-balance signal which must exist so that control is exercised. This latter contribution is a function of the overall loop gain in the system.

In case (b), the change in base current is entirely a function of the current gain from VT_1 to the invertor primary.

The change in I_B due to temperature is very small, for it is a common mode signal and subject to the very low common mode gain which exists when the two halves of VT_1 are closely matched.

A small concomitant of the I_B problem is that any milliammeter which is to be used as an e.h.t. meter in the dropper chain must be connected as shown in Fig. 6 rather than at the bottom of R_R, with one terminal grounded.

The complete answer to the difficulties presented by the existence of a base current is to substitute a field-effect transistor for VT_1. In Fig. 7, a dual n-channel f.e.t. is used*. Here, the gate current is negligibly small, which means that a much lower value of I_{Ch} can be tolerated. In the circuit shown, a current of 100μA has been chosen,

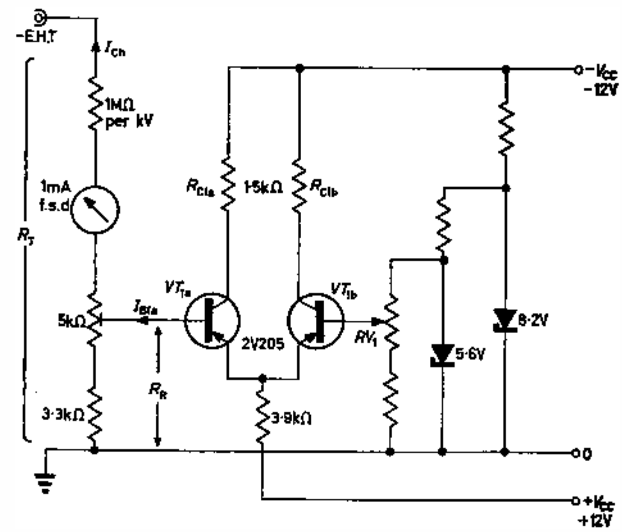


Fig. 6. Bipolar transistor comparator stage

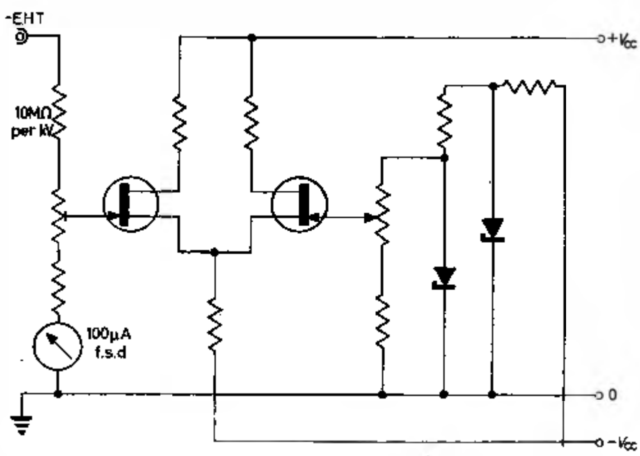


Fig. 7. F.E.T. comparator stage

merely in order that a meter having a 100μA f.s.d. might be inserted at the lower end of R_R. Otherwise, this current could have been still further reduced.

In both the bi-polar and f.e.t. comparators, temperature stability is achieved by choosing a Zener diode with a very small temperature coefficient. This implies either a 5·6V unit, or a comparatively expensive multiple device which combines two or more diodes whose composite temperature coefficient approaches zero. The former method has been found satisfactory provided that the 5·6V Zener is cascaded with a higher voltage unit so that the Zener current is kept essentially constant.

* *E.G. The Amelco 2N3935, imported by Lectropon Ltd, Wellington Street, Slough.*

The load stability is a function of the overall transconductance of the control loop. Consequently, the current gain of the following amplifier will be dictated by the stability required.

Although it is not the purpose of the present article to describe a complete e.h.t. power supply, a comparatively simple amplifier to follow the comparator stage was designed in order to demonstrate its effect upon the stability. The resulting complete circuit appears in Fig. 8. (It is hoped that a future article will describe a more comprehensive e.h.t. unit utilizing microcircuit techniques.)

The Control Amplifier

A dual-range e.h.t. unit is depicted in Fig. 8, which will provide a continuously varying voltage from 400V to 1kV and 1kV to 2kV. The two ranges are switched by means of a ceramic wafer which performs three separate functions:

(a) The dropper chain is reduced from 20MΩ for the 2kV range to 10MΩ for the 1kV range.

(b) The Zener diode network is modified to define the range limits accurately.

(c) The h.v. rectifier is changed from a doubler on the 2kV range to a simple bridge on the 1kV range.

This latter procedure allows the invertor transformer to have a maximum output of 1kV, thus reducing insulation problems.

The amplifier following the f.e.t. comparator consists of a pnp stage having an emitter resistor so chosen that the transistor presents a reasonable load to the f.e.t. drain. VT_3 is simply a current amplifier and VT_4 is the power transistor which controls the current to the invertor. It will be noticed that the emitter of VT_4 is taken directly

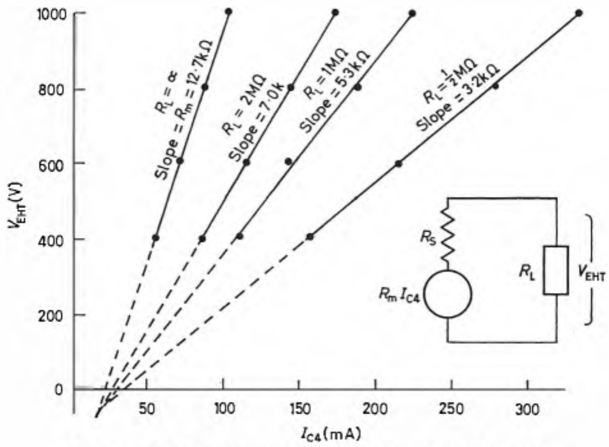


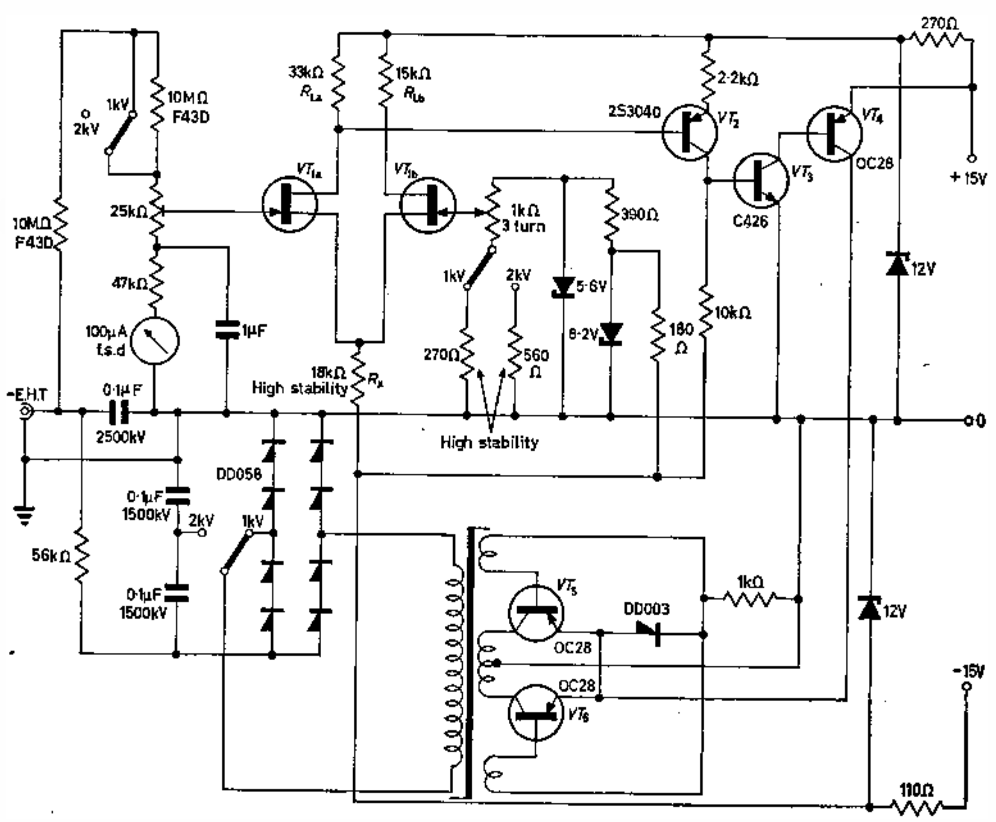
Fig. 9. Transfer characteristic of invertor/rectifier unit

to the +15V rail so that when the invertor is working at full voltage and current, some 12V are available at its primary while VT_3 still retains a V_{CE} of 3V.

The invertor and its associated rectifiers and smoothing capacitors can be treated, very approximately, as a current to voltage convertor. The justification for this is the set of transconductance plots shown in Fig. 9. The fact that each plot is a straight line confirms that, for incremental changes, the unit can be described by a simple Thevenin equivalent circuit consisting of a voltage generator $R_m I_C$, in series with an internal resistance R_S.

The fact that the plots do not pass through the origin when extrapolated backwards indicates that the full equivalent circuit should contain further generators. However, for the purposes of calculation, where incremental changes are involved, the simple Thevenin equivalent will suffice.

Fig. 8. A 2kV, 4mA e.h.t. unit using a dual comparator



When

$$R_L = \infty\ ,\ R_m I_C = V_{eht}$$

$$\text{or } R_m = \frac{V_{eht}}{I_C} = 12{\cdot}7k\Omega$$

When

$$R_L \neq \infty\ ,\ V_{eht} = \frac{R_m I_C R_L}{R_S + R_L}$$

$$\text{or } R_S = \frac{R_m R_L}{\text{Slope}} - R_L \qquad \ldots\ldots (5)$$

For the three plots, this equation gives slightly different results, but R_S may in general be taken as 1·5MΩ for the particular invertor and rectifier system used and when operated on the 1kV range. (R_S will be approximately doubled for the 2kV range.)

Calculation of the Output Impedance

The block diagram of Fig. 10 shows that the e.h.t. voltage is sampled by the dropper chain and that a suitable fraction is presented to an amplifier which contains the f.e.t. and VT_2. This represents a voltage to current convertor whose transconductance is G_m, which may be determined as follows:

Fig. 11(a) shows the dual f.e.t. followed by VT_2, and in that the gate of $VT_{1(b)}$ is held at a predetermined potential this half can be thought of as a common-gate stage fed[5] from the source of $VT_{1(a)}$. Hence, R_X is shunted by the input resistance of $VT_{1(b)}$, that is $R_{in(CG)}$.

Similarly, the drain load of $VT_{1(a)}$ is shunted by a resistance given approximately by $h_{fe(2)}R_E$. Thus, the model shown in Fig. 11(b) applies.

Using the voltage symbols shown in Fig. 11(b),

$$v_G \simeq v_S$$

$$\text{so } A_V = (v_D/v_G) \simeq (v_D/v_S) = \frac{R_{L(1a)}\|h_{fe(2)}R_E}{R_X\|R_{in(CG)}} \quad \ldots\ldots\ldots (6)$$

The value of input resistance for a common gate stage is given by[6]:

$$R_{in(CG)} = \frac{R_{L(1)} + R_{DS}}{1+\mu} \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

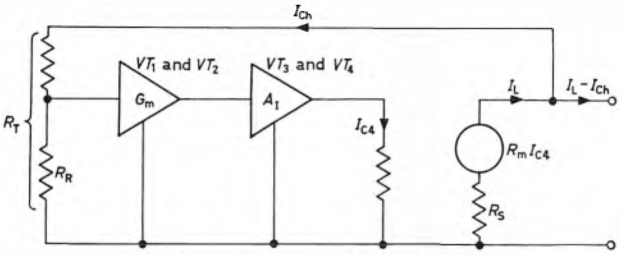


Fig. 10. Arrangement of control loop

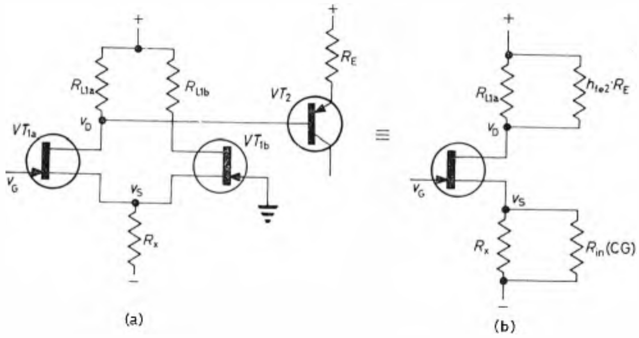


Fig. 11. Circuit and model for VT_1/VT_2 composite stage

where R_{DS} is the drain-source resistance, $1/g_{as}$, and μ is the amplification factor $g_m R_{DS}$ or g_m/g_{as}.

Finally, the emitter current of VT_2 and hence the collector current, will be given by:

$$i_{c2} \simeq \frac{v_G A_V}{R_E}$$

so that the transconductance G_m of the VT_1/VT_2 combination is:

$$G_m = (i_{c2}/v_G) = (A_V/R_E) \quad \ldots\ldots\ldots\ldots (8)$$

In the present case the following parameters apply:

For the 2N2935, $g_m = 300\mu\Omega^{-1}$, $g_{as} = 5\mu\Omega^{-1}$

For the 2S3040, $h_{fe(min)} = 40$

Since $R_{L(1a)} = 33\text{k}\Omega$, $R_{L(1b)} = 15\text{k}\Omega$, $R_X = 18\text{k}\Omega$ and $R_E = 2{\cdot}2\text{k}\Omega$, equations (7), (6) and (8) give respectively:

$$R_{in(CG)} = 3{\cdot}5\text{k}\Omega$$

$$A_V = 10$$

$$G_m = 4{\cdot}5\text{mA/V}$$

Consider again Fig. 10. The output resistance can be found as follows:

$$R_{out} = \frac{\Delta V_{eht}}{\Delta I_L} = \frac{\Delta V_{eht} R_S}{R_m \Delta I_{C4} - \Delta V_{eht}}$$

$$\text{now } \Delta I_{C4} = \Delta V_{eht} \,.\, (R_R/R_T) \,.\, G_m A_I$$

where A_I is the current gain of VT_3 in cascade with VT_4.

$$\text{so } R_{out} = \frac{R_S}{(R_R/R_T)G_m A_I R_m - 1} \simeq \frac{R_S R_T}{R_R R_m G_m A_I} \quad \ldots\ldots\ldots\ldots (9)$$

If $A_I = h_{fe3}h_{fe4}$, which in the present case is 40 . 100 = 4 000, equation (9) gives:

$$R_{out} \simeq 1{\cdot}2\text{k}\Omega$$

The measured value for R_{out} was about 1·0kΩ, which is in good agreement when it is recalled that the most pessimistic values for the transistors were taken.

Clearly, the above derivation is very approximate. For example, unless g_m is very large, then $v_S \neq v_G$; and the ratio R_R/R_T is as given only when the maximum value of V_{eht} is being generated. However, there is little point in refining the procedure, for equation (9) shows that R_{out} is a function of the various gain parameters of the transistors, and these are subject to the usual wide tolerances and variations.

R_{out} at 2kV will be approximately twice that for the 1kV level, for R_S will double and the rest of the factors in equation (9) will remain as before.

Line Variations

The e.h.t. voltage will be in part a function of the value of the line voltage, and in particular, any variation in the negative line to the long-tailed pair would be expected to have a significant effect. This is because the comparatively low resistance of a common gate f.e.t. is presented to this line in series with R_X. To a first approximation, assuming the loop gain of the system to be infinitely high, the voltage at the collector of $VT_{1(a)}$ will not change. Any variation in V_{CC}^- will therefore drive a current through $VT_{1(b)}$ given by:

$$\Delta I = \frac{\Delta V_{CC}^-}{(R_X + R_{in(CG)})}$$

where $R_{in(CG)}$ is the common gate input resistance of $VT_{1(b)}$. The voltage change across $R_{in(CG)}$ will be:

$$\Delta V_{R in(CG)} = R_{in(CG)} \,.\, \Delta I = \frac{R_{in(CG)}\,\Delta V_{CC}^-}{(R_X + R_{in(CG)})}$$

For no current change in $VT_{1(a)}$, then $\Delta V_{R in(CG)}$ must be counteracted by an equal voltage change at the gate of $VT_{1(a)}$:

$$(R_R/R_T) \,.\, \Delta V_{eht} = \frac{R_{in(CG)}\,\Delta V_{CC}^-}{(R_X + R_{in(CG)})}$$

$$\text{or } \frac{\Delta V_{eht}}{\Delta V_{CC}^-} = \frac{R_T \,.\, R_{in(CG)}}{R_R(R_X + R_{in(CG)})}$$

which in the present case works out at 29 volts per volt. This is a very pessimistic figure which will be considerably less in practice, for the real amplifier has a gain which will allow some considerable current change in $VT_{1(a)}$. The measured value was about 15 volts per volt, and this was easily reduced to 3·5 volts per volt using the very simple Zener stabilization shown.

Ripple

The invertor unit is designed to deliver a good square wave, which being so, the degree of smoothing required is minimal, for both the bridge and doubler rectification systems are essentially full-wave methods. However, the ripple—which for the present invertor/rectifier unit is better than 10mV peak-to-peak—is highly non-sinusoidal, and the higher frequency components can be amplified by the feedback loop where the relevant phase shifts permit this. A 1μF capacitor has therefore been placed across the lower end of the dropper chain to minimize this effect, and the final output ripple under the worst conditions is better than 100mV peak-to-peak.

Invertor Unit Operating Principle

The circuit is analysed on the basis of a resistive load and the effect of rectifier loads is considered later.

Take the basic circuit of Fig. 12 and assume that VT_5 starts to conduct. A voltage will be induced in the feedback winding n_f and will be in such a direction as to support conduction of VT_5. Due to this feedback action, the transistor will switch hard on and by similar action VT_6 will be turned off.

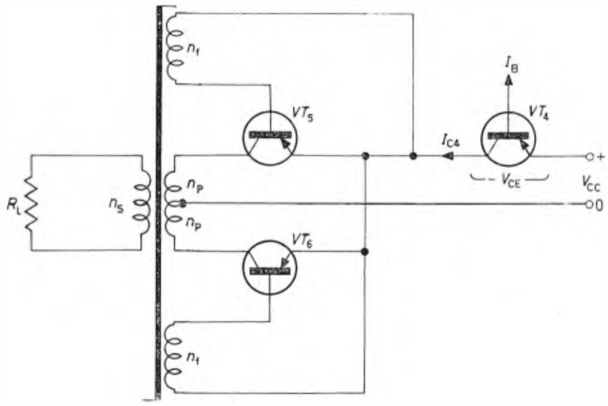


Fig. 12. Basic invertor configuration

The circuit at this time can be simplified to that shown in Fig. 13 where

$$R = R_L (n_p/n_s)^2$$

and L_p = half-primary inductance (assumed linear).

The base current of VT_4 is constant in value and its magnitude is controlled by the voltage comparator circuit of the complete e.h.t. unit.

The current I_{C4} flowing in the circuit at any instant of time T will be subject to the expressions

$$I_{C4} = f(V_{CE}) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (10)$$

this being the output characteristic of VT_4

and:

$$I_{C4} = \frac{V_{CC} - V_{CE}}{R} + \int_0^T \frac{(V_{CC} - V_{CE})}{L_p} dt \quad \ldots\ldots (11)$$

The solution of these two equations gives the operating point of VT_4 and a graphical solution for time $T = 0$ is shown in Fig. 14. As T increases, the operating point of the transistor moves towards point B. It is seen that the collector-emitter voltage of VT_4 increases with time and hence V_p $(= V_{CC} - V_{CE})$ decreases with time.

Referring again to Fig. 12, the feedback voltage is given as $V_p(n_f/n_p)$ and hence the base current of VT_5 will reduce with time. When the base current reaches the value I_{C5}/h_{FE5}, VT_5 will start to turn off, causing a further reduction in feedback voltage. This positive feedback effect turns off transistor VT_5 and turns on VT_6. The sequence of operations will now be repeated and an alternating voltage is developed across the transformer primary.

Due to energy storage in L_p, the operating point of VT_4 will vary during subsequent half cycles of operation over the range C through A to B. The first half cycle (neglecting changes in base current VT_4) will be of only half the duration of a half cycle in the steady state condition.

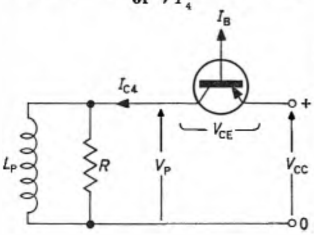
In the practical circuit, the transformer was wound on an HCR core, which exhibits a square loop B-H characteristic. The inductance of the transformer primary is very large until saturation of the core occurs, at which time this inductance falls to a very small value. Consequently, the operating point of VT_4 is close to point A in Fig. 14 during most of the half cycle of operation of the invertor, and rapidly switches towards point B when saturation of the core takes place. In this manner a good square voltage waveform is developed at the transformer.

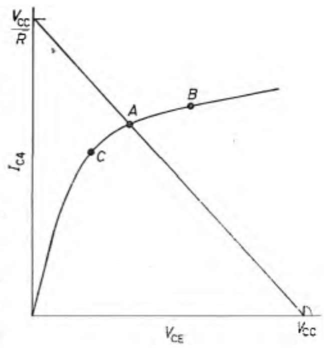
Design Considerations

The c.g.s. system of units is most commonly used when presenting data on core materials and all the following expressions are given in these units.

Fig. 13 (below). Equivalent circuit for Fig. 12

Fig. 14 (right). Operating mode of VT_4





During the time of one half cycle, the transformer core is driven from saturation in one direction to saturation in the other, and, assuming a good square voltage waveform at the transformer, this time will be given by the equation:

$$t = \frac{2B_{sat}n_pA}{V_p \cdot 10^8} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (12)$$

where A is the cross-sectional area of the magnetic circuit. Hence,

$$f = \frac{V_p \cdot 10^8}{4B_{sat}n_pA} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (13)$$

It is apparent from equations (10) and (11) that in order to achieve a square wave, I_{C4} should be substantially constant during the half cycle. That is:

$$\frac{V_{CC} - V_{CE}}{R} \gg \int_0^T \frac{V_{CC} - V_{CE}}{L_p} dt \quad (\text{where } T = (1/4f)$$

$$\text{or } (1/R) \gg (1/4fL_p)$$

Substituting for f from equation (13), and writing $L_p = \dfrac{4\pi n^2 A\mu}{10^9 l}$ gives, $(R/V_p) \ll (4\pi/10) \cdot (\mu/B_{sat}) \cdot (n_p/l)$

$$\ldots\ldots (14)$$

where l is the length of magnetic path.

It can be seen from equation (14) that a toroidal winding is advantageous, for not only does this construction yield the maxium value for μ, but it also enables the ratio n_p/l to be maximized, since for a given wire size n_p is proportional to the winding area. A toroidal winding can, however, present insulation problems for high voltage transformers and from this point of view a laminated core is preferred.

It should also be noted that equation (14) is more easily satisfied when a large transformer is employed, but this, of course, would oscillate at a low frequency. The designer is thus presented with a choice between the extremes of either providing a large transformer with a low frequency of oscillation, but good waveform, or a smaller transformer operating at a high frequency, but producing a waveform with a large amount of droop.

When the output of the unit is obtained via a bridge rectifier, the first of these alternatives is to be preferred, for a small size of smoothing capacitor is then sufficient to meet the ripple specification. In the case of a voltage multiplier circuit, however, a higher frequency of oscillation may be preferred (to keep the size of capacitors to a reasonable value) even at the expense of droop in the invertor waveform. A compromise solution based on experimental results is obtained.

The Invertor Transformer

A full theoretical design of the invertor transformer is complex and it is probably most useful to make an initial design on a simple basis and then to modify this in the light of test measurements.

The maximum output voltage should be taken to correspond to a V_p of approximately 12V. The feedback voltage at minimum output should be about 1 to 2V.

A core size is selected and a primary winding designed using a gauge of wire appropriate to the primary current. This winding should fill roughly half the available winding area. The frequency of oscillation is determined from equation (13) and an estimate of droop from equation (14). If these results are not thought to be satisfactory, a different lamination is chosen and the process repeated.

When the core size has been established, secondary and feedback windings are designed and a prototype transformer can then be produced.

The two halves of the primary should be bifilar wound to reduce voltage transients developed across the transistors during switching.

In the actual design, 0·004in thick HCR laminations of M.E.A. 225 size were used, the secondary being wound with 4 000 turns of 45 s.w.g. wire, properly distributed. The centre-tapped primary used 80 turns of 26 s.w.g. wire bifilar wound. This transformer (Belclere No. KN3383) was used in the experimental unit of Fig. 8 and now forms the basis of a further development involving microcircuit techniques.

Acknowledgments

The authors would like to express their gratitude to Prof. H. W. Gosling of the University of Wales, Swansea, for his many useful comments during the composition of the article; and to Hilger & Watts Ltd for permission to publish part of the contents therein.

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Electronic Telephone Exchanges

The first of a new generation of telephone exchanges has now entered public service at Ambergate in Derbyshire.

The exchange, known as the Post Office TXE2 (Telephone Electronic Exchange Type II), has been developed by Ericsson Telephones, part of the Plessey Telecommunications Group, in co-operation with the Post Office Engineering Department under the auspices of the British Joint Electronic Research Committee. This Committee, comprising the GPO, AEI, ATE, ETL, GEC, and STC, was set up to co-ordinate the electronic exchange development activity of the Post Office and the five principal manufacturers of telephone switching equipment.

The TXE2 system is marketed by the Plessey Telecommunications Group under the name PENTEX.

The exchange installed at Ambergate will cater initially for 700 lines and is capable of being extended to 2 000 lines.

This system is designed for use in rural and small urban areas. In contrast with the majority of existing telephone exchanges, in which electromechanical switches step in synchronism with the 'pulses' produced by the telephone dial, this new electronic exchange system 'stores' the dialled information and then switches the call via a selected path to the wanted subscriber. The switching is carried out by reed relays.

One of the main features of the exchange is the use of registers to set up calls. Apart from storing the dialled information, the register also stores the calling subscriber's telephone number and the class of service accorded to the calling and called subscribers' telephones. The class of service defines the type of telephone installation and the facilities and prohibitions associated with it. Because this information is stored, a speech path to connect the two telephones can be selected and switched after the dialling has been completed if an examination of the wanted line has shown it to be free.

The speech path is selected by silicon solid state electronic circuits whose operation is both faster and inherently more reliable than that of electromechanical circuits. The high speed of operation enables one set of control equipment to handle all the calls passing through the exchange.

Calls are supervised, both during and after setting-up, by supervisory relay sets, the register having been released after set-up to enable it to process other calls. A caller is connected via a three-stage switching network to a supervisory relay set appropriate to the kind of call which is being made.

When the register has received the dialled digits the wanted telephone line is examined and, if it is free, a path is set up to it from the supervisory relay set, thereby completing the connexion between the two telephones. The second half of the speech path between the two telephones includes an additional switching stage. This is provided to ensure adequate availability of connexions between the chosen own-exchange relay set and the wanted telephone line.

Each of the switching stages consists of crosspoint switches, in which one selected group of wires is connected to another selected group by the contacts of a reed relay.

The use of registers and high speed electronic circuits make possible the introduction of new facilities as the demand arises. Because the register is capable of receiving dialled information sent much faster than by a rotary dial, it can work equally well with push-button telephones, which are likely to be widely used in the not-too-distant future.

High-speed path selection enables a 'second attempt' to be made automatically if a call fails to be established owing to a random fault, or an 'alternative route' to be attempted if the first route is found to be congested.

'Direct dialling-in' to private automatic branch exchanges can be provided. With this facility the caller dials not only the P.A.B.X.'s telephone number but also the number of the wanted extension, thereby saving himself time and reducing the work of the P.A.B.X. switchboard operator.

A group of subscribers such as doctors with a common practice can be offered 'subscriber-controlled transfer' facilities whereby the setting-up and cancellation of the facility is achieved by dialling simple codes. Also, a facility is provided whereby subscribers are able to reject incoming calls at will. Telephone-users are advised automatically that the line they are dialling does not wish to receive calls.

A 'short-code' dialling facility can be offered for numbers which are dialled frequently.

The electronic exchange equipment at Ambergate is mounted on cards which plug into enclosed racks. Good connexions are ensured between the cards and the rack wiring by using palladium-plated contacts on plugs and sockets.

Simplified Design of the Feedback Pair

By J. C. S. Richards*, Ph.D., M.I.E.E.

The design of a two transistor a.c. amplifier stage with shunt feedback is reduced to the application of a few very simple formulae covering d.c. bias, gain and bandwidth. The approximations and assumptions involved are valid for most low power silicon planar transistors; the performance obtained is sufficiently near optimum for many purposes.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

ONE of the most useful a.c. amplifier stages is the two transistor feedback pair[1,2] shown in Fig. 1. It requires very few components, and can give a well defined current gain and a good high frequency performance. It has a low input impedance and a relatively high output impedance, so that there is no difficulty in operating stages in cascade. Because both the transistor currents and the gain are established by means of negative feedback, the stage is relatively insensitive to the particular type of silicon transistors used. If a little less than the maximum possible performance is acceptable, the design procedure can be reduced to the application of a few simple rules, which are sufficiently accurate for many purposes. They are particularly useful if very inexpensive, loosely specified transistors are to be used.

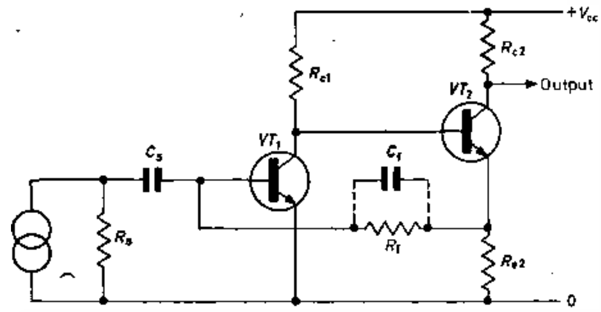

Fig. 1. Complete amplifier stage

The following are the most important assumptions made in deriving the formulae given below. The transistors are typical low power (<1W) silicon planar types having collector saturation voltages of less than 0·5V; the range of collector currents considered is 10μA to 30mA. The base spreading resistance and internal feedback (the 'Early effect') are ignored; they are significant only in limiting cases. The collector and emitter currents are assumed to be equal. The small signal current gain is assumed to be the same as I_c/I_b, and both are symbolized by β. Resistors are assumed to be within $\pm$10 per cent of their design values, and the allowed range of ambient temperature is 0 to 65°C.

Symbols with suffixes 1 or 2 refer respectively to the first and second transistor; and the suffix *d* indicates that a quantity has its design value. (Max) or (min) indicates a maximum or minimum value respectively for that quantity, taking into account the effects of both tolerance and temperature.

D.C. Conditions and Bias

The collector current I_2 in the second transistor is generally chosen on the basis of signal handling requirements, power supply economy, or frequency response (see below); in input stages, noise may be a predominant factor. The collector current I_1 in the first transistor is made I_2/n, where n can be chosen to optimize some particular aspect of performance; a value of 3 is generally suitable and is henceforth assumed.

To establish these currents it is given that

$$R_{e2}I_2 = V_{be1} + R_f I_1/\beta_1$$

For the current and temperature ranges of interest, V_{be1} lies between 0·3 and 0·9V for most transistors. If R_f is restricted to be less than $\beta_1 R_{e2}$, and R_{e2} chosen from the formulae

$$R_{e2d}I_{2d} = 0{\cdot}7\text{V} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

then the actual value of I_2 will be between 0·3 and 2 I_{2d}.

The other resistances are given by

$$R_{c1d}I_{2d} = 3(V_{cc(min)} - 1{\cdot}5)\text{ volts} \quad \ldots\ldots\ldots \quad (2)$$

$$R_{c2d}I_{2d} = 0{\cdot}4\,(V_{cc(min)} - 3)\text{ volts} \quad \ldots\ldots\ldots \quad (3)$$

where V_{cc} is in volts.

This value of R_{c2} ensures that V_{ce2} is at least 1·5V.

Gain

Provided $R_f > R_{e2}$, the gain A_i of the amplifier may be written as

$$A_i = \frac{-A_{io}}{1 + \gamma A_{io}}$$

where the feedback factor γ is $R_{e2}/(R_{e2} + R_f)$

and

$$A_{io} = A_v/[\gamma + R_{c2}/R_s + R_{e2}/(\beta_1 r_{e1})] \quad \ldots\ldots \quad (4)$$

Here A_v is the voltage gain between the base of VT_1 and the emitter of VT_2, and is given by

$$A_v[1 + R_{c1}/(\beta_2 R_{e2})] = R_{c1}/r_{e1} \quad \ldots\ldots\ldots\ldots \quad (5)$$

Since r_{e1}, R_{c1} and R_{e2} are approximately proportional to the currents in the transistors, the value of A_v is fairly independent of the value of I_{c2d}, and depends mainly on V_{cc} and β_2. If β_2 is 50, $A_{v(min)}$ is about 90 when $V_{cc(min)}$ is 6 volts and 140 when $V_{cc(min)}$ is 10 volts. For a given design $A_{v(max)}$ is unlikely to be more than $3A_{v(min)}$.

In determining A_{io}, the factor R_{c2}/R_s is often dominant. If identical stages are in cascade, then R_{sd} is equal to R_{c2d}, and R_{e2d}/R_{sd} is about $1{\cdot}8/(V_{cc} - 3)$. If V_{cc} is small, then to obtain a large value of A_{io} it is necessary to operate successive stages at progressively higher currents; by making I_{1d} of the second stage equal to I_{2d} of the first stage and so on throughout the amplifier, R_{e2d}/R_{sd} is reduced to about $0{\cdot}6/(V_{cc} - 3)$. This arrangement also keeps the total current drawn from the power supply small and makes it easier to decouple the early stages.

The term $R_{e2}/\beta_1 r_{e1}$ has a maximum value of about $15/\beta_{1(min)}$, or 0·3 if $\beta_{1(min)}$ is 50.

Hence if γ is 0·1, and $V_{cc(min)}$ is 6 volts, then $A_{io(min)}$ is about 90 when successive stages work at similar currents, and about 150 when the currents are progressively increased.

* *Aberdeen University.*

When $V_{cc(min)}$ is 10 volts, the corresponding values for $A_{io(min)}$ are about 240 and 340.

The current gain A_i has a maximum value of $1/\gamma$, and $A_{i(max)}/A_{i(min)}$ is $(1 + 1/A_{i(min)})$. The value of γ to be used is largely determined by the permitted tolerance in gain. In general terms it may be said that it is seldom possible to make $1/\gamma$ greater than 100, since the biasing arrangements then demand that $\beta_{(min)}$ be greater than 100; if $1/\gamma$ is about 30 the gain is sufficiently well defined for many purposes; and unless high quality resistors are used for R_{e2} and R_f, there is little advantage in making $1/\gamma$ less than 10.

Bandwidth

The loop response function has two real negative poles; these are well separated, and even when $1/\gamma$ is as small as 3, there are unlikely to be very large peaks in the frequency response. To simplify the rather complicated formulae for the poles it is assumed that $C_c r_e f_T$ lies between 1 and $1/A_v$. Here C_c is total collector to base capacitance, including any wiring capacitance, and f_T is the gain bandwidth product of the transistor. To a rough approximation the poles are then given by

$$\omega_A = 1/(A_{io} R_{e2} C_{c1}) \quad \ldots\ldots\ldots\ldots \quad (5)$$

$$\omega_B = \omega_{T1} R_{e2}/R_{c2} \quad \ldots\ldots\ldots\ldots \quad (6)$$

and ω_A is usually much less than ω_B.

Provided ω_B/ω_A is greater than $4\gamma A_{io}$, the response of the amplifier to a step function input exhibits no overshoot[3], and the bandwidth f_2 is very roughly given by $\gamma/(\pi R_{e2} C_{c1})$.

For most small planar transistors f_T has a value of several hundred MHz at currents of about 10mA, and a value of a few MHz at currents of 10μA. Within this range of currents it is usually not too far from the truth to take f_{T1} equal to $100\sqrt{I_1}$ MHz, where I_1 is in milliamperes. The condition for no overshoot then becomes

$$C_{c1} > 10\gamma(V_{cc(min)} - 3)\sqrt{I_2}\,\text{pF} \quad \ldots\ldots\ldots \quad (7)$$

and, if this is satisfied, the 3dB bandwidth is given very approximately by

$$f_2 = 200\gamma\, I_2/C_{c1}\,\text{MHz} \quad \ldots\ldots\ldots\ldots \quad (8)$$

where C_{c1} is in pF, V_{cc} in volts, I_2 in milliamperes.

If the inequality (7) is not satisfied, it is necessary to add a capacitance C_f in parallel with R_f in order to avoid overshoot; this is also the best way to reduce the bandwidth of the amplifier to the value required. C_f can be calculated, or found empirically. However, if $C_{c1} < 30$pF, and $V_{cc(min)} < 12$V, then a value of C_f which ensures critical damping, and the corresponding bandwidth, are given conservatively by

$$R_f C_f \geqslant 10^5/I_2 \mu\text{sec} \quad \ldots\ldots\ldots\ldots \quad (9)$$

$$f_2 \leqslant 160\, I_2\,\text{kHz} \quad \ldots\ldots\ldots\ldots \quad (10)$$

where I_2 is in mA, C_f is in pF, and R_f is in ohms.

The low frequency response is determined by $R_s C_s$.

Power Supply

If successive stages use the same polarity of transistor (npn or pnp) then unwanted signals fed in from the power supply may be troublesome. Any variation δV_{cc} in V_{cc} is equivalent to an input current of $\delta V_{cc}/R_{c2}'$, where R_{c2}' is the collector load resistance of the previous stage. To avoid disturbances due to unwanted positive feedback, the power supply impedance Z_{cc} must have a magnitude much less than $\gamma R_{c2}'$.

These effects can be very much reduced if successive stages are alternately npn and pnp transistors as shown in Fig. 2. Fluctuations in the power supply feed small currents into the bases of VT_2' and VT_2''; these currents are injected within the feedback loops and are relatively ineffective. Feedback via Z_{cc} is essentially negative (assuming Z_{cc} has an argument of magnitude less than $\pi/2$); so that even if the condition $|Z_{cc}| \leqslant A_{io}' R_{e2}' \gamma' \gamma''$ is not satisfied instability is unlikely.

Gain Control

For precise switched gain control the most useful circuit is probably that of Fig. 3. The attenuation factor n is essentially determined by R_1 and R_{c2}', which must therefore have a close tolerance. R_2 keeps the impedance in the base circuit of VT_1'' sensibly constant so that the gain of the second stage does not alter; it need not normally have a

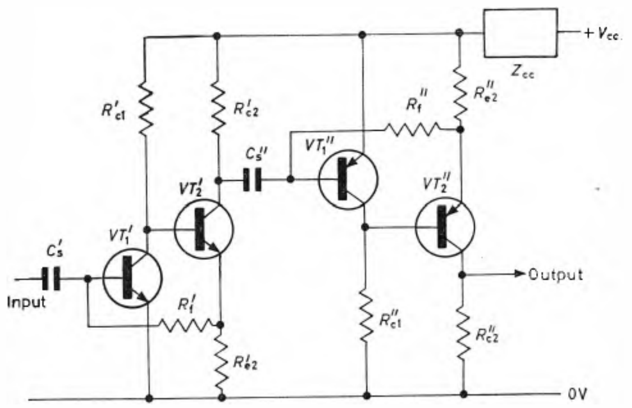

Fig. 2. Use of npn stage followed by pnp stage to avoid positive feedback via the power supply

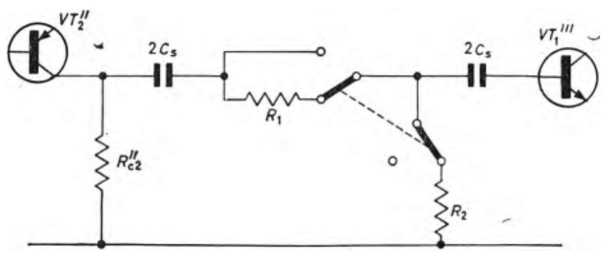

Fig. 3. Method of obtaining precise switched gain control

close tolerance. R_1 is $(n - 1R_{c2}'$, and R_2 is $nR_{c2}'/(n-1)$. The impedance seen by the collector of VT_2' changes less than 25 per cent for all n greater than 2, and this is normally unimportant.

A fine control is often needed to bring the gain of the complete amplifier to some required value. A preset potentiometer in series with R_f is convenient.

Input and Output Stages

The input stage of an instrumental amplifier is usually required to have either a very low, or a very high, or a very accurately defined input impedance. The feedback pair has a usefully low input impedance of about R_f/A_v; a moderately large and well defined input impedance can be obtained by adding a series resistance. If very large input impedance is required ($>$ 1MΩ say) it is probably best to precede the feedback pair with a field effect transistor.

The effect of feedback on the noise figure can be complicated, but for most purposes it is sufficient to use manufacturers' data for transistors without feedback and assume a resistance $(R_f + R_{e2})$ in parallel with the source resistance R_s.

The biasing data given at the start of this article assume that the stage is to feed a low impedance load; that is, that only small output voltage swing is required. An output stage is often required to give a large voltage, and this usually means that the tolerance allowed on I_2 is small. By using for VT_1 a transistor with closely specified V_{be} ($\pm$ 0·05V, say) and high $\beta_{(min)}$, it is not difficult to reduce the total variation of I_2 to less than $\pm$30 per cent, which is adequate for most purposes.

Integrated Circuits

An integrated circuit version[2] of the feedback pair can be expected to give a very good high frequency performance, since stray reactances are small and reproducible. Integrated circuits intended as very wide band or r.f. or i.f amplifiers have been available for some time.

The feedback pair uses so few components and is so easy to design that, where very high frequency performance is not important, there is not a great incentive to sacrifice the flexibility available with conventional construction; at any rate not until a wide range of integrated circuits is available at very low cost. Perhaps the most useful integrated circuit would consist of a single encapsulation containing an npn feedback pair followed by a pnp feedback pair, each with a gain of 10 or 30, and capable of giving, say, a milliampere of output current when operated from a 6V supply.

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Small Brushless D.C. Motors with Electronic Speed Control

In the past small battery operated d.c. motors such as are used for driving tape recorders, record players and similar apparatus have suffered from the inherent faults which are common with small d.c. motors, i.e. unreliable performance and short life. The West German firm AEG has now designed a small electric motor which overcomes these faults. The conventional commutator with carbon brushes has been replaced by an electronic circuit. This basic circuit can easily be extended in order to make speed control possible without using a centrifugal switch.

Driving motors for tape recorders, record players, film cameras etc. must meet the following requirements:

(a) High efficiency to keep batteries small and to ensure a long working period per battery.

(b) Small overall dimensions.

(c) High working reliability.

(d) Long life.

(e) Low level of noise.

Requirements (a) and (b) can easily be met, but (c), (d) and (e) are very difficult to realize. Certainly the d.c. motor has a very high starting torque, but the sliding contacts—commutator, sliprings in speed controlled motors—are a great disadvantage. These sliding contacts shorten the lifetime and moreover they increase the noise level of the motor.

The new motor operates as follows:

The torque originates from the interaction of the three starter coils 1, 2 and 3 with the rotating permanent magnet 4 (Fig. 1). The fixed stator coils can be made alive through the switching transistors 5, 6 and 7, as compared with the classical d.c. motor the stator and rotor have changed their function. Rectifiers are connected in series with these transistors, being fed by the control coils 8, 9 and 10 which are mounted as secondary windings on a special fixed transformer core 11. An oscillator equipped with a transistor generates an alternating voltage of 100kHz and feeds the primary winding 12. In this oscillator the coil 12 and the capacitor 13, arranged as a parallel oscillatory circuit, determine the amplitude of the auxiliary frequency. The magnetic coupling between the primary winding 12 and one control coil is provided by the control segment 14 fixed to the rotor. This arrangement ensures that independent of the angular velocity only the stator winding 1, 2 or 3 is on load.

Speed control can be arranged as follows:

Each stator winding is only on load during a third of a rotor revolution. In the on-position of the stator coil the voltage induced by the rotating magnet is so directed that the difference between the battery voltage and the induced voltage prevails over the current forming the motor torque. Between two on-positions the sign of the induced voltage in the currentless stator windings reverses for half a period so that one can measure the induced voltage caused by rotary motion during this time interval with a diode. This induced voltage which has been rectified is proportionally dependent on the motor speed.

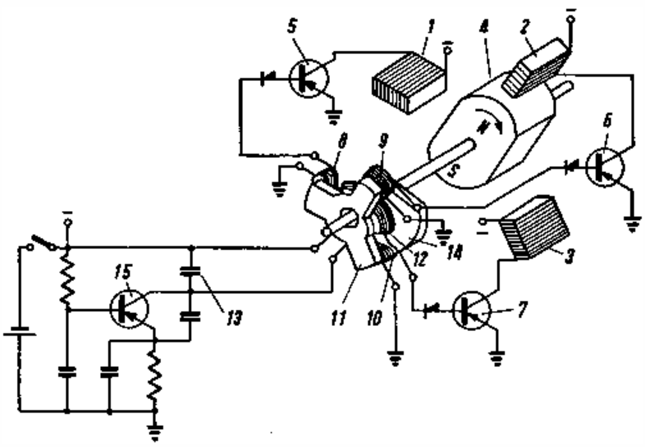
For each stator coil one rectifier is provided. A capacitor smoothes the pulsating voltage. The bridge connected in series incorporates a voltage dependent resistor. It is so balanced that near the nominal speed the diagonal voltage is zero on the other side of the bridge.

Should the speed vary the bridge becomes unbalanced. The signal thus formed is forwarded from the emitter of the transistor incorporated in the bridge to the base of the oscillator-transistor which influences the oscillation amplitude. This has its effect on the supply voltage of the switching transistors 5, 6 and 7 which then adjust the speed of the motor.

This ensures a very constant speed and the normal fluctuations are 0·05 per cent on average. By means of an adjustable resistor incorporated in the bridge the speed can easily be altered.

Fig. 1. Principle of the brushless d.c. motor

(1), (2), (3) stator windings; (4) rotating permanent magnet, rotor; (5), (6), (7) switching transistor; (8), (9), (10) control coils; (11) transformer core (ferrite); (12) primary winding; (13) oscillatory circuit capacitor; (14) control segment (ferrite); (15) Oscillator-transistor



Economy and Reliability in Nucleonic Instrument Design

By B. Shepherd*, C.Eng., M.I.E.E.

Described in this article is a series of instruments designed with value engineering applied to produce reliable but not expensive instruments. Novel circuits have been employed in the design of a fast decade board with visual indication, a stable transformerless high voltage unit, and tunnel diode pulse height discriminators.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

IT is almost too easy when designing with transistors to make an instrument more and more complex, with a higher specification, and consequently much more expensive, than is in fact really needed. The general users of nucleonic equipment, for example, are faced with so many sources of error that only the more specialized user will require instruments of high performance.

Now that epoxy encapsulated silicon planar transistors are available in both npn and pnp form for only a few shillings each, it is possible to design circuits of a complementary nature with often a consequent reduction in passive components. The proven high reliability of planar transistors does not appear to be impaired in epoxy encapsulation over the temperature range required for industrial instruments.

Throughout the designs metal film resistors were used in the critical, and metal oxide in the non-critical low power positions. The increased cost over using carbon resistors is less than 2 per cent on the instrument selling prices and is considered justified by the added reliability.

Scaling

The fundamental function of a nucleonic instrument is the ability to count randomly arriving pulses over a period of time. A measure of the efficiency of the system is the time taken to process one pulse, i.e. the dead time. During this time the scaler is incapable of counting a second pulse and thus the overall count may have 'dead time losses'. To test this dead time a scaling system is usually subjected to pulse pairs, the spacing between the pair of pulses being gradually reduced until the second pulse is not counted. The pulse pair resolution aimed for in the scaling section was 1μsec.

Wire-ended numerical indicator tubes were chosen for the visual read-out scaling portion because of their acceptable appearance, high readability in strong ambient light conditions and low cost. Their life is 25 000 hours plus and even at end of life there is usually no abrupt failure. Until a few years ago transistors to drive these tubes were quite expensive because of the high voltage required. However, several manufacturers now make numerical indicator tube drivers with a collector/base rating of the order of 70 to 120V. These transistors can be connected directly in the cathode circuits of the indicator tube with no collector resistors or collector supply as the 'off' transistors operate quite safely in the avalanche region.

Three scaling systems with visual indication were examined.

(a) Four binary divide-by-16 with feedback, eliminating the last six states, with appropriate diode decoding to the tube drive transistors.

(b) Ten state ring counter directly suitable for driving the tube drive transistors.

(c) A bi-quinary system with appropriate odd and even decoding to the tube drive transistors.

The expected component quantities and costs are given in Table 1 based upon reliable and wide temperature range circuits. This shows preference for the bi-quinary system.

TABLE 1

Decade Counter Cost Comparisons

COMPONENTS	BINARY		DECIMAL		BI-QUINARY	
	Quantity	Cost (s)	Quantity	Cost (s)	Quantity	Cost (s)
Counting Transistors	8	27	20	67	12	40
Tube Drive Transistors	10	33	10	33	12	40
Counting Diodes	18	27	0	—	4	6
Decoding Diodes	30	40	0	—	0	—
Reset Diodes	4	4	1	1	2	2
Resistors	35	18	33	17	28	14
Capacitors	16	8	12	6	11	5
Numerical Indicator Tube	1	25	1	25	1	25
Totals	122	£9 2s.	77	£7 9s.	70	£6 12s.
Number of Soldered Joints	271		193		173	

* *Isotope Developments Ltd.*

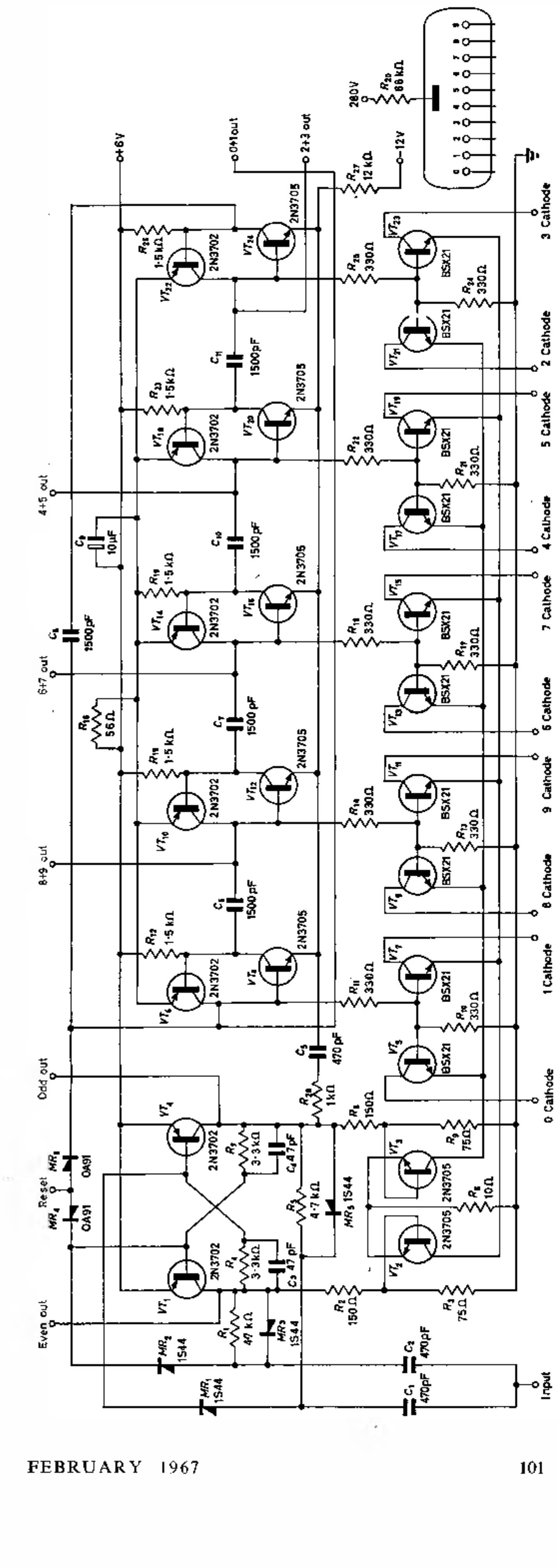


Fig. 1. Decade counter

The circuit diagram of the decade counter is shown in Fig. 1. The supply voltage used is limited to 6V to protect the base-emitter junctions of the planar transistors during transients. The binary section VT_1 VT_4 is a fairly conventional fast binary having speed-up diodes. The ring of five, however, will not be so familiar.

The pulse from the binary is connected via C_5 to the common emitters having a long-tail to a negative line. By virtue of the d.c. conditions any one stage being 'on' will cut off the other four stages. A positive pulse to the common emitters will cut off the 'on' stage as well. This stage going 'off' will pass a negative pulse to the base of the next stage priming it. When the input pulse disappears the following stage will come 'on'.

One disadvantage of a ring system of counting is that it is naturally slower. This is because the transfer does not take place on the leading edge of a pulse but on the falling edge and the time-constants of the circuit have to be such that the input pulse disappears while the priming pulse is still available. The circuit shown works satisfactorily at a regular input frequency of 500kHz over a fairly wide range of components.

Resetting can be accomplished by taking the appropriate line to approximately +7V. By means of two diodes this returns the binary to a particular state and brings 'on' the 0-1 position of the quinary, cutting all of the others off.

For decoding purposes only two extra transistors, VT_2 and VT_3 and no diodes are required. The quinary is advanced once for every two input pulses and each stage represents 0 and 1, 2 and 3, 4 and 5, and so on. Each of the five quinary outputs is coupled to two bases of the appropriate indicator tube drive transistors. The emitters of these drive transistors are commoned in odd and even form to the two extra transistors which are switched alternatively by the binary. Thus for any one state there is only one cathode-to-ground path. Both the binary and all five quinary outputs are brought out for use in automatic stop conditions. In the instrument six decades are connected in series giving a maximum count capability of 999 999.

A gating circuit is included prior to the first decade which will automatically remove the input pulses to the first decade upon some signal, i.e. a predetermined count, being received. A switch can select the predetermined count in nine ranges between 100 and 10^6 counts, the pulse being derived from the appropriate 0 and 1 or 4 and 5 section of the quinary. The 'stop' pulse operates a simple fast bistable. The output of this bistable switches off the emitter current of a pulse inverting stage as indicated in Fig. 2. VT_3 is included to give a fast returning edge to the output pulse. The bistable can also be operated manually by 'start' and 'stop' press keys on the front of the instrument.

Timing

In order to determine reliably the rate of arrival of the input pulses it is necessary to determine the time over which a count has been accumulated. It is normal, therefore, to run a second scaling system counting pulses derived from a time reference which is gated to 'start' and 'stop' at the beginning and end of a counting period.

The counting section of the timer used consists of the standard decade board, already designed, in view of its superior appearance and also greater reliability over mechanical timers. A low voltage mains frequency signal operates a Schmitt trigger which produces a sharp edged square wave at the same frequency as the input.

quantity of electrons at the final anode. The current gain of a photomultiplier tube is typically 10^6 to 10^8 and is highly dependent upon the voltage across it. For this reason the high voltage must remain stable to within approximately 0·1 per cent usually over a period of many hours.

Stable e.h.t. circuits giving outputs of the order of 2kV have in the past normally used an invertor circuit with a transformer and high voltage secondary coupled into two or three stages of a Cockcroft-Walton multiplier. Invertor circuits of this type have many problems associated with the design, particularly when used for variable supplies.

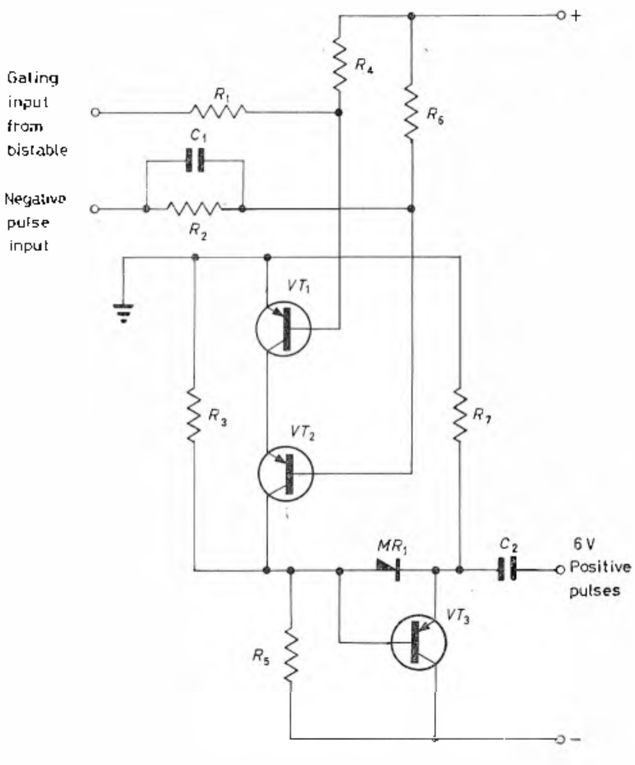


Fig. 2. Pulse gate

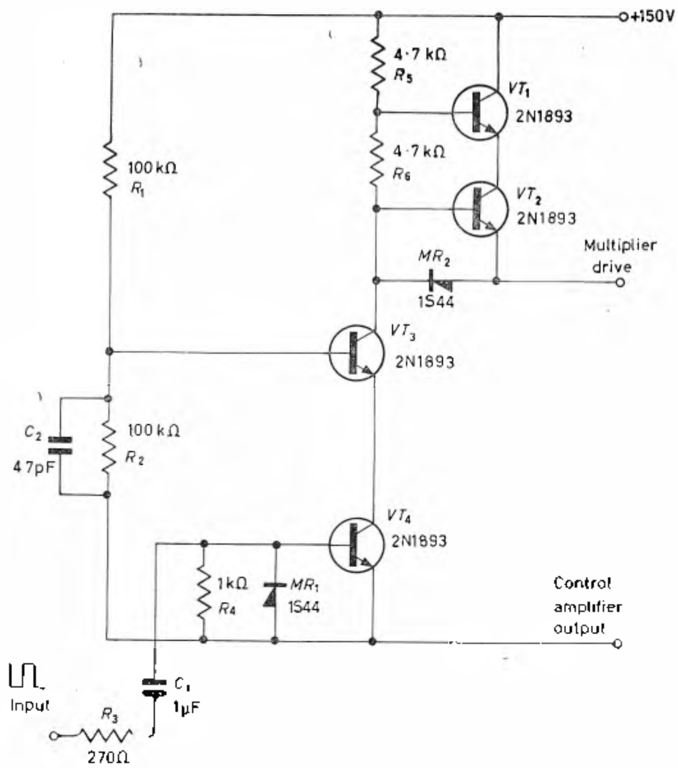


Fig. 3. H.V. switch circuit

The number of decade boards used is kept to a minimum of four by preceding them by a prescaling system having four ranges. This prescaling system has a total division capability of 300. Using the binary and ring counting system explained before without, of course, any read-out transistors, etc., the divide by 300 was split into its lowest common multiples, i.e. 2, 2, 3, 5, 5. These division ratios are combined by switching, to give output pulses of 10 per second, 1 per second, 100 per minute and 10 per minute. Small neon indicators included in the appropriate positions between decade boards are also switched to act as decimal points. The four timing ranges therefore obtained are 0·1 to 999·9sec, 1 to 9 999sec, 0·01min to 99·99min, and 0·1 to 999·9min.

High Voltage Supply

Many kinds of radiation detectors used with nucleonic instruments emit very small flashes of light or scintillations. The emitted photons are detected and amplified by means of photomultiplier tubes to produce a charge pulse or

Until recently high voltage transistors were difficult to obtain, costly and usually had a fairly low upper frequency cut-off. Several high voltage silicon planar devices are now, however, available and it was therefore decided to design a high voltage square wave switching circuit to drive straight into a multi-stage Cockcroft-Walton and thus eliminate the transformer. Above 20 stages the efficiency falls off unless exceptionally high value capacitors are used, and also the arrangement is economically unsound in view of the large number of components required. Therefore the square wave amplitude required to produce an output voltage of 2kV must be at least 100V peak-to-peak.

When driving into a Cockcroft-Walton the power is being drawn from the switching circuit in both states of the square wave and thus a low output impedance is required for each state. To keep the capacitors in the Cockcroft-Walton to a reasonable value and to reduce the filtering problems the operating frequency should be as high as possible. To reduce the power wasted during the switching transients the frequency should be kept as low as possible. A frequency of 50kHz appeared to give optimum results.

The switching circuit giving the required waveform is shown in Fig. 3. Transistors VT_3 and VT_4 in their 'off' condition share the supply voltage by means of the bias resistors R_1 and R_2. The required low output impedance is obtained from the emitter follower VT_2. The low impedance during the 'on' stage is derived by the bottomed transistors VT_3 and VT_4 and the diode MR_2. The supply voltage is now equally divided between VT_1 and VT_2 due to the split in the collector load resistor.

in Fig. 4. This amplifier is fed with a constant current derived by the large voltage across R_{11}. The gain of this stage is set to 50 by resistors R_7, R_8, R_9 and R_{10}.

A second amplifier with a gain of approximately 40 set by R_{14} and R_{21} is used to vary the output voltage applied to the supply line of the 50kHz switching circuit. This encloses transistors VT_5 to VT_{11}. Three transistors VT_9, VT_{10} and VT_{11} are cascaded to withstand a possible 150V and to share the power dissipation.

It is necessary that the amplifier gains are known fairly accurately because the whole system represents a closed

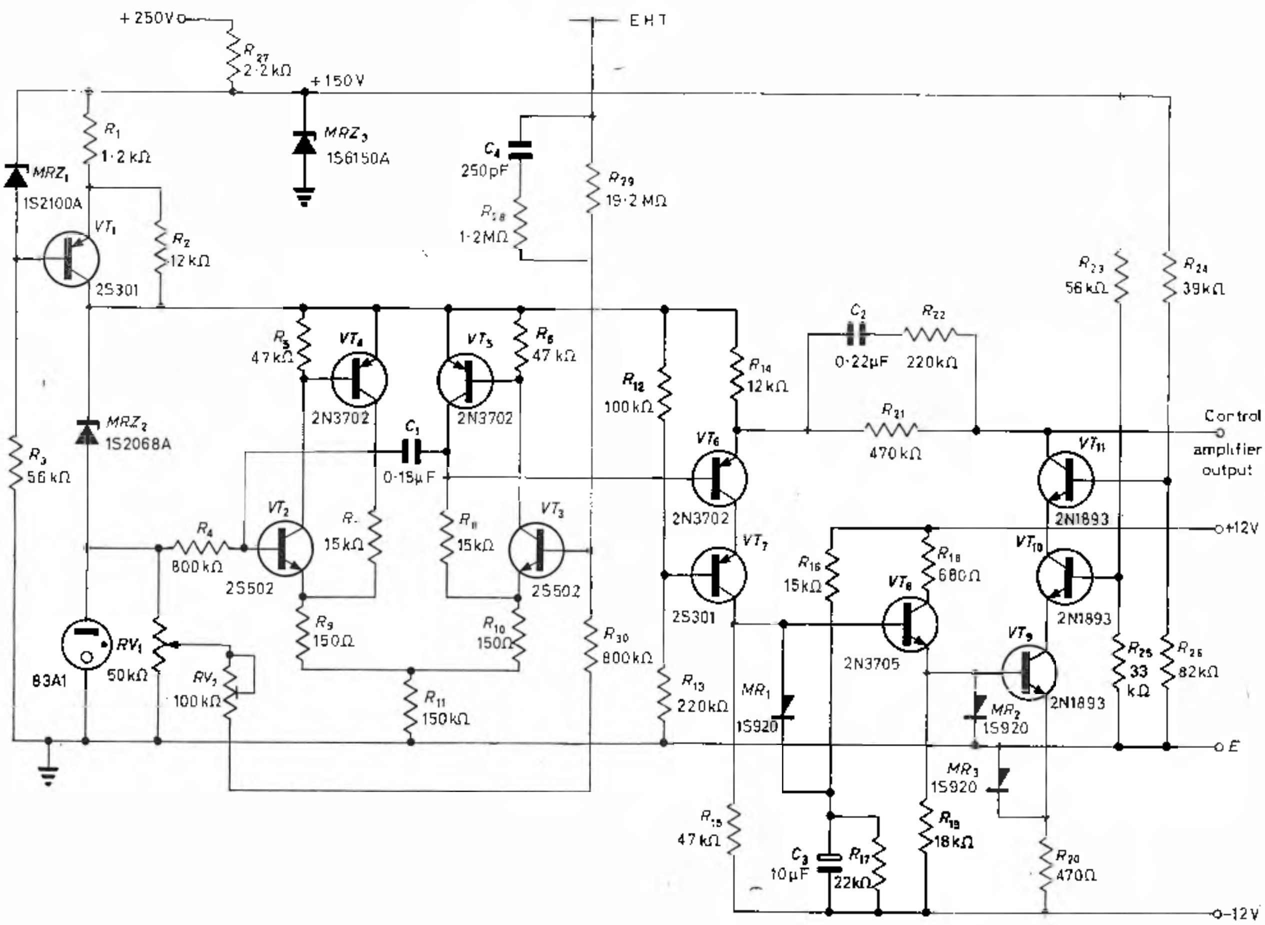


Fig. 4. E.H.T. control amplifier

This circuit operates satisfactorily with a supply voltage varied from about 5V to 150V. (The amplifier output varies from earth to 145V). C_2 improves the voltage sharing between VT_3 and VT_4 during the switching transients. A big advantage of this type of circuit is that the square wave may be derived from some free-running square wave oscillator and thus the frequency will remain constant regardless of the output voltage or the load current.

In practice 16 stages were required for the Cockcroft-Walton multiplier. Two stages of resistor capacitor filtering produced a 2kV output with less than 2mV of 50kHz breakthrough.

The final output of the e.h.t. unit is monitored by high value metal film resistors and compared with a neon stabilizer by means of a differential amplifier as indicated

feedback loop with a very large inherent lag introduced by the Cockcroft-Walton multiplier and filter arrangement. Frequency compensation is employed in three places, namely the dominant lag C_1 in amplifier 1, the limited lag arrangements of R_{22} and C_2 in amplifier 2, and the phase advance network C_4 and R_{28} across the high value monitoring resistors.

Transistor VT_1 has been employed for two reasons. The first is that it provides a sufficiently large voltage to fire the neon stabilizer when the unit is first switched on. Secondly with MRZ_1 it acts as a constant current generator stabilizing the reference tube current.

As the lower end of the monitoring chain of resistors is varied by the ten turn potentiometer RV_1 connected across the reference tube the output voltage will also

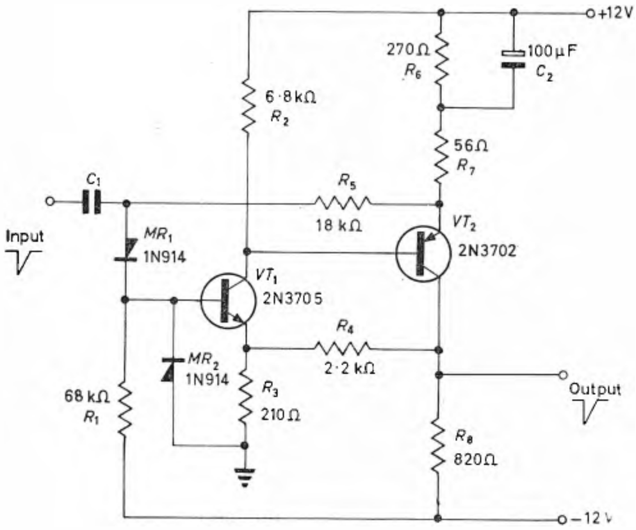


Fig. 5. Gain-of-ten amplifier

vary. The required output range of 300V to 2kV is easily obtained.

The output current capable of being drawn from the high voltage terminal is limited in case of inadvertant short-circuit because the amplifier output via VT_9 cannot supply more than a certain current defined by R_{30}.

Pulse Amplification

As explained previously the pulse from the photomultiplier tube assembly is essentially a quantity of charge. The time distribution of this pulse as measured at the photomultiplier tube anode is approximately 100 to 400nsec. The charge, however, is not wholly dissipated in the anode resistor in this time. A certain amount of it is used to charge up associated stray and cable capacitance which then discharges into the load resistor at a rate depending upon their *RC* time constant. The voltage pulse derived therefore consists of a negative pulse having a rise time of between 100 and 400nsec and an amplitude and fall time depending upon the load capacitance and the anode and following circuit resistances. A measure of the resolution of an amplifier is the time taken to recover from some large overload and return to its no-pulse base line. Techniques have been evolved for reducing the recovery from overload time e.g. double differentiation and delay line shaping[1].

In view of the required simplicity of design, however, it was decided to remove as many causes as possible of long recovery time inside the amplifier and at the same time to give the amplifier as large an unsaturated swing as possible.

The normal cause of a long recovery time is the charging of some capacitors inside the circuit due to the changing d.c. levels of the circuit upon saturation e.g. a coupling capacitor frequently finds a drastic change in impedance presented to it when the following stage saturates.

The standard gain-of-10 amplifier is shown in Fig. 5. The two transistors are connected in a feedback ring and the gain is set by resistors R_3 and R_4. The circuit is designed to receive only negative pulses and is appropriately biased. The output does not saturate until the pulse exceeds 10V. A larger voltage pulse than the unsaturated 1V maximum input applied to the input capacitor C_1 only serves to cut off the diode MR_1 and the charging current into the capacitor is very small.

Two such gain-of-10 amplifiers are cascaded with appropriate decoupling to give a gain of 100 and preceded by a buffer stage with a gain of one to obtain an input resistance of approximately 50kΩ. Tests on the amplifier at maximum gain show that the output follows the input within 1 per cent and upon overload clips the top of the pulse cleanly and follows the portion of the pulse which is within its capability.

Tunnel Diode Discrimination

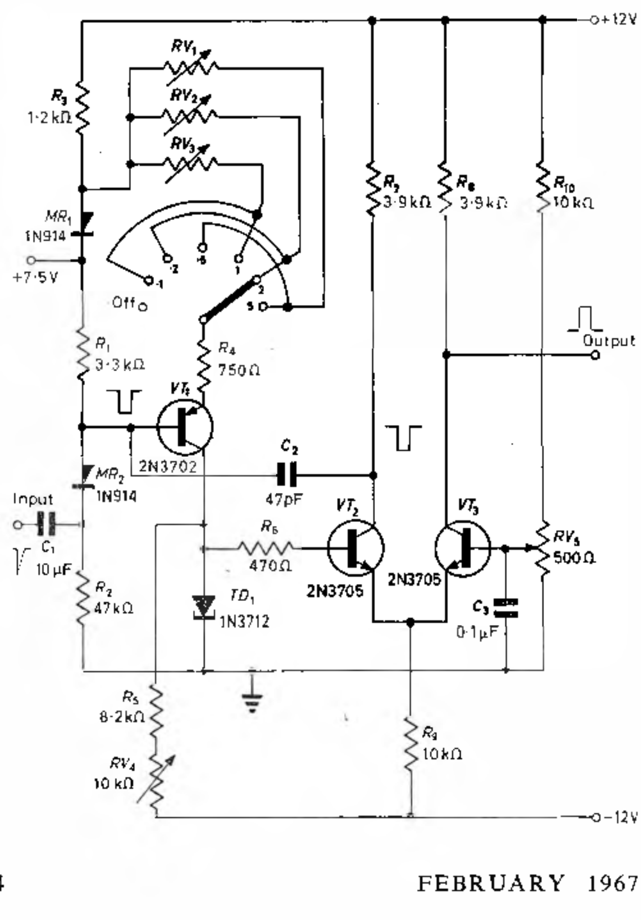
A window technique is used to analyse the pulse height. A lower threshold, controlled by a potentiometer, passes pulses which are greater than some particular value as set on the control, over a range of 0 to 5V. As soon as the input pulse exceeds this value the portion in excess is amplified by another gain-of-10 amplifier and fed to a lower discriminator having a threshold of approximately 0·5V. Referred to the threshold control this represents 50mV. A second discriminator operates if the amplitude of the pulse exceeds another upper threshold level.

The two discriminator outputs are combined in an anticoincidence system such that an output pulse is derived only when the input pulse lies within the two levels.

The upper threshold tunnel diode discriminator is shown in Fig. 6. Tunnel diodes are used for discrimination because although fairly expensive they have the ability to operate very quickly. A pulse which exceeds the discriminator level by a few millivolts with a total rise time of a few hundred nanoseconds may only be in excess of this level for a few nanoseconds. If the pulse is not to be stretched then discrimination must take place during this time.

Transistor VT_1 normally passes 1mA as defined by the bias resistors R_1 and R_2 and the emitter resistor R_4 R_5 and

Fig. 6. Tunnel diode discriminator



RV_4 to the negative line require 1mA and therefore the tunnel diode current is nominally zero. The input signal applied to the base of the transistor will increase the collector current and this increase will pass into the tunnel diode. When this extra current exceeds 1mA the tunnel diode will change to its high voltage state in approximately 5nsec. This change switches VT_2 'on' which is normally cut off by the common emitter of VT_3. The negative pulse at the collector of VT_2 is passed via C_2 to the base of VT_1 bringing VT_1 harder on and thus holding the tunnel diode in its high voltage state. The capacitor

If the upper threshold input pulse is not present VT_3 will come 'on', bringing 'on' VT_5.

Ratemeter

To limit the length of the article it is not proposed to describe the circuit arrangements of the ratemeter because this is very similar to that already described[2].

Conclusions

A range of simple nucleonic instruments has been described. Component costs have been kept to a minimum

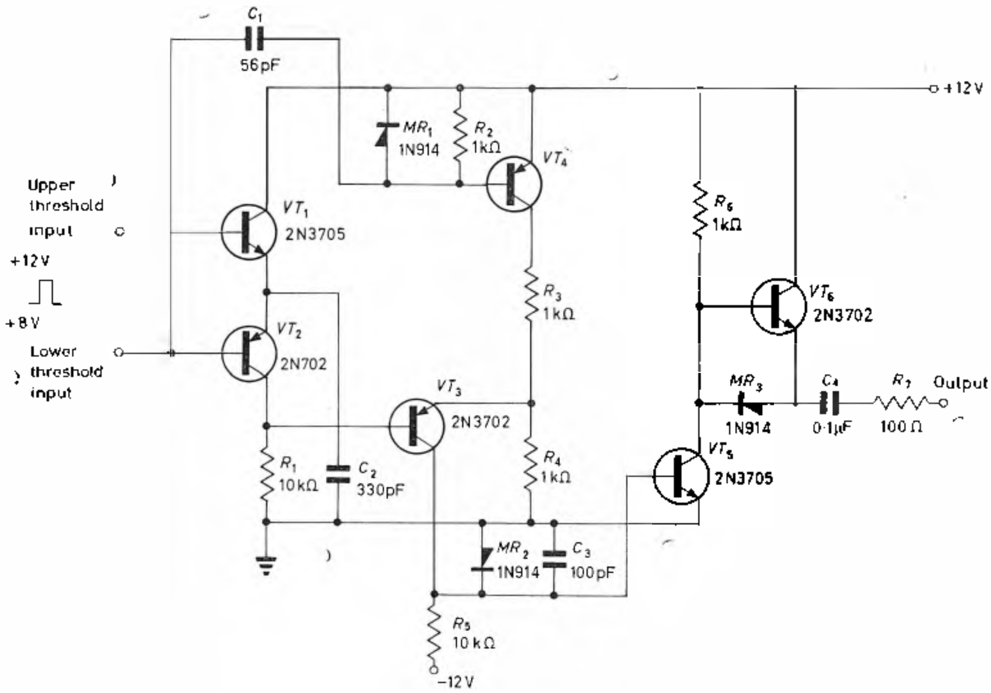


Fig. 7. Anti-coincidence circuit

C_2 discharges, gradually cutting VT_1 'off' until the current through the tunnel diode falls below the valley current causing it to revert to the low voltage state. It then switches 'off' VT_2, and VT_3 back 'on'. This monostable action has a width defined by C_2 and is set nominally to 0·5μsec. A much wider overload pulse will, however, hold the tunnel diode circuit on until the pulse itself falls back to the base line. Thus the output of the discriminator can be as wide as the maximum width input pulse but cannot get narrower than 0·5μsec.

The method of obtaining anticoincidence is shown in Fig. 7. The positive edge of the lower threshold input pulse reverse biases VT_3 and a pulse derived from this leading edge via C_1 is absorbed in MR_1. If the upper threshold input is now applied it will charge C_2 via VT_1 to plus 12V. The falling edge of the upper threshold pulse occurs before that of the lower threshold pulse and VT_1 is now reverse biased. The negative edge of the lower threshold pulse now brings 'on' VT_2 and discharges C_2 into R_1 producing a positive pulse across R_1. The negative edge of the lower threshold pulse differentiated via C_1 brings 'on' VT_4 momentarily. VT_3 is thus presented with a positive pulse on its emitter and a positive pulse on its base which are coincident in time. As a result it does not give an output pulse.

without unduly effecting the reliability. New techniques have been used in the design of divide-by-10 boards, stabilized e.h.t. design and pulse discriminator circuits.

Apart from the power supply shunt transistors, all the approximately 300 transistors and nearly all the diodes are silicon. The effects of temperature are therefore small.

One disquieting thing is that whereas several American transistor manufacturers have epoxy-encapsulated ranges, at the time of writing no British manufacturer has entered this field. Many British manufacturers offer npn silicon planar transistors at competitive prices but few offer the pnp complement.

If transistor instruments are to take over fully from valve equipment in the industrial and entertainment markets cheap and reliable components are required with which the designer can produce novel circuits.

Acknowledgments

Acknowledgment is made to the Directors of Isotope Developments Ltd for permission to publish this article and to Mr. H. W. Finch for his assistance.

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A Digital Instrument for Measuring the Slippage of Asynchronous Motors

By V. Böhm*, V. Popescu* and T. Teitel*

This article describes a fully transistorized digital instrument which provides a direct reading, in decimal form, of the slippage of induction motors, by coupling a photo-electric pick-up to a disk mounted on the shaft of the motor. Using throughout pulse-counting techniques, the method is completely free from errors associated with other methods for the measurement of this parameter.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

THE slippage of induction motors, defined as the difference between the synchronous speed and the actual rotating speed of the motor shaft, expressed in percentage of the synchronous speed, is an important parameter of the motor in operation, as it is closely related to the momentum, the delivered power, the power-factor, etc. The accurate measurement of this parameter in different working conditions is important not only in laboratories in which electric motors are being tested in order to determine their characteristics, but also in other applications where two shafts are rotating at nearly the same speed, and the relative difference of these speeds has to be evaluated.

The conventional methods for the measurement of slippage usually involve a time measurement; it is assumed that the mains frequency remains constant and has the nominal value during the measurement, the slippage being determined by stroboscopic methods. It is obvious that such methods introduce large errors, and, since this measurement cannot be performed on a direct instrument, a computation has to be undertaken to acquire the desired figure.

The digital instrument described here is a direct reading device, which eliminates the disadvantages of the conventional methods, since it is based on a differential measurement carried over two slightly different frequencies; the mains frequency and the frequency derived from the actual rotating speed of the motor shaft.

Basic Principle

The instrument consists of a tachymetric pick-up device, coupled to the rotating shaft of the motor, and an electronic assembly, containing the logic circuits, connected to the same network as the one which feeds the induction motor. The amplified block-diagram of the instrument is shown in Fig. 1.

The tachymetric pick-up T.P.U., coupled to the motor under test M, may be of the inductive or photo-electric type, and generates a certain number of pulses, for instance ten, for each shaft revolution. These pulses are fed through a flexible cable, to the electronic unit, where they are gated in the gate G before reaching the decade counting units N_1 and N_2.

The gate G is periodically operated by a flip-flop B.S.G., put into action by the mains frequency after division in the flip-flop cascade D.F.R. If, for instance, the mains frequency is 50Hz and D.F.R. divides this by 100, the gate G would be opened and shut every two seconds, this time interval T being the measuring interval.

The reset stage R.S. is put into action by the same flip-flop B.S.G. which also acts upon the gate, and generates pulses which reset the decade counting units N_1 and N_2 shortly before each opening of the gate.

The decimal scaling units N_1 and N_2 count the pulses delivered by the pick-up device during the measuring interval T; since they are reset before each gate opening, at the end of the measuring interval, when the gate shuts, N_1 would have scaled the number of units, and N_2 the number of tenths of the input pulses.

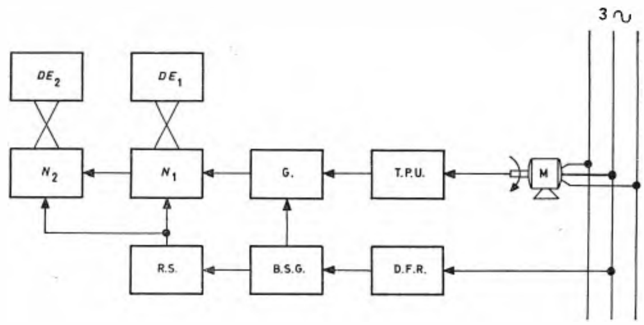

Fig. 1. Simplified arrangement of the instrument

Instead of displaying the actual number of counts registered in each of the decimal scaling units, the display elements DE_1 and DE_2 are connected in a way to read out the decimal complement of this number as indicated below:

Registered Number of Counts	*Displayed Figure*
0	9
1	8
2	7
3	6
4	5
5	4
6	3
7	2
8	1
9	0

It is thus possible to use standard decimal counting units, instead of reversible counters, to provide the necessary subtraction.

The operation of the instrument can be followed from the diagram in Fig. 2.

At the initial moment, synchronous with the reset pulse C, the gate G opens and stays open the whole measuring interval T determined by the mains voltage pulses divided in the D.F.R. pulse demultiplier. In the chosen example, the mains frequency being 50Hz and the D.F.R. circuit dividing by 100, this time interval T would be 2sec.

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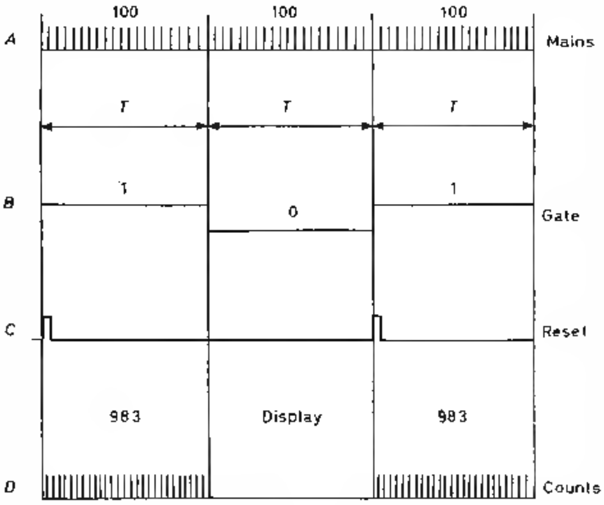


Fig. 2. Diagram of operation

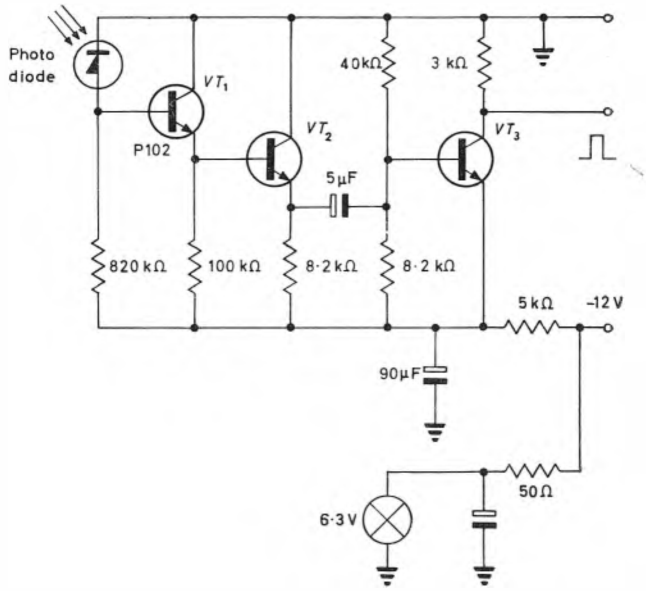


Fig. 5. Pre-amplifier circuit of the tachymetric pick-up

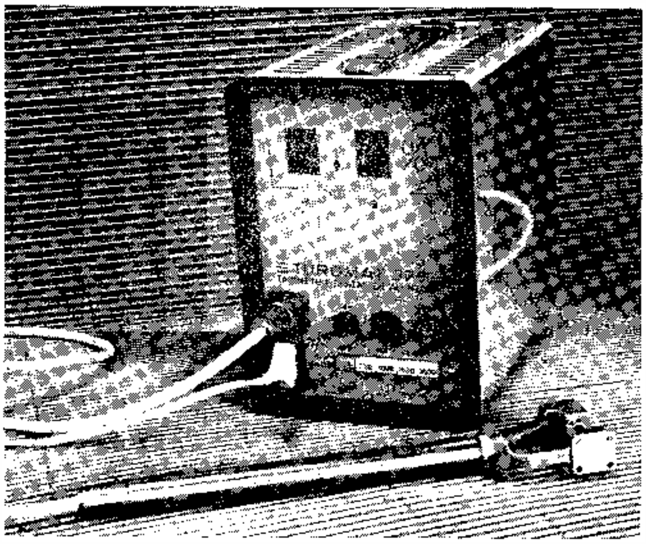


Fig. 3. View of the instrument

During the period T, the pulses delivered by the tachymetric pick-up which is coupled to the motor shaft are counted by the decimal scaling units until gate G closes again, passing from state 1 to state 0 (diagram B, Fig. 2).

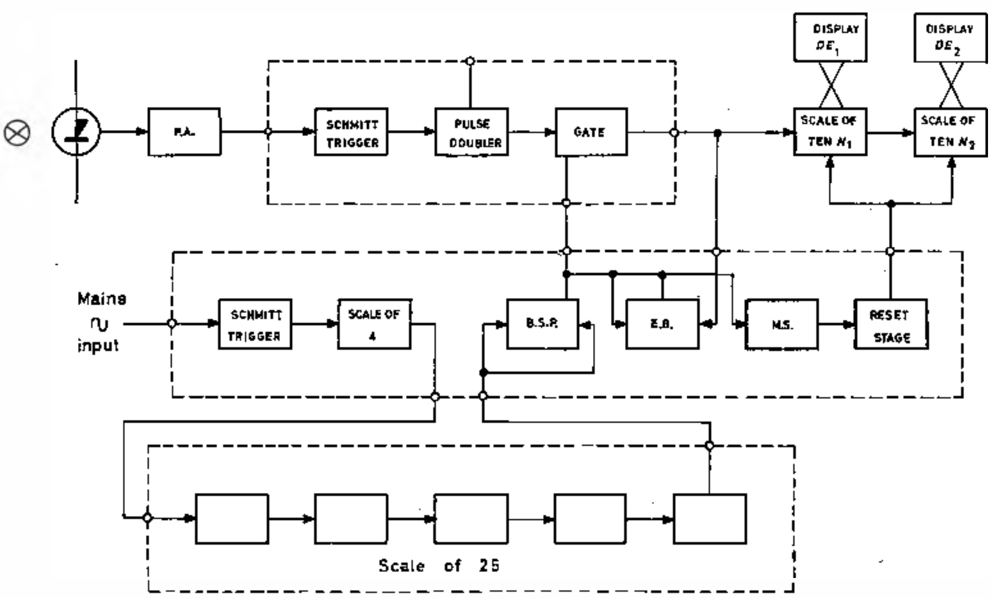
The number of pulses thus registered is a function of the slippage of the motor. Assume first that the 3 000rev/min motor shaft revolves at synchronous speed, the slippage being in this case zero; in this case the number of pulses delivered by the pick-up T.P.U. would be 1 000. If the actual slippage of the motor is less than 10 per cent, during the time interval T, the number of pulses would range from 900 to 1 000. In the example that follows, assume this number as being 983.

As the decimal scaling circuits of the instrument count only the units and the tenths, at the closing of gate G, the scaling circuit N_1 would have scaled the figure 3 (units) and N_2 the figure 8 (tenths). Following the complementary indication of the read out elements, the displayed figures would be 1 instead of 8, and 6 instead of 3, so that the final indication of the slippage would be 1·6 per cent.

The actual slippage figure in the above given example is 1·7 per cent instead of 1·6 per cent which is indicated by the instrument. This systematic error of −0·1 per cent introduced by the instrument can be eliminated by the arrangement that will be described below.

The displayed figure can be read out by the operator during the two second pause that follows the measuring interval, then the cycle continues. In that way, on the front

Fig. 4. Detailed arrangement



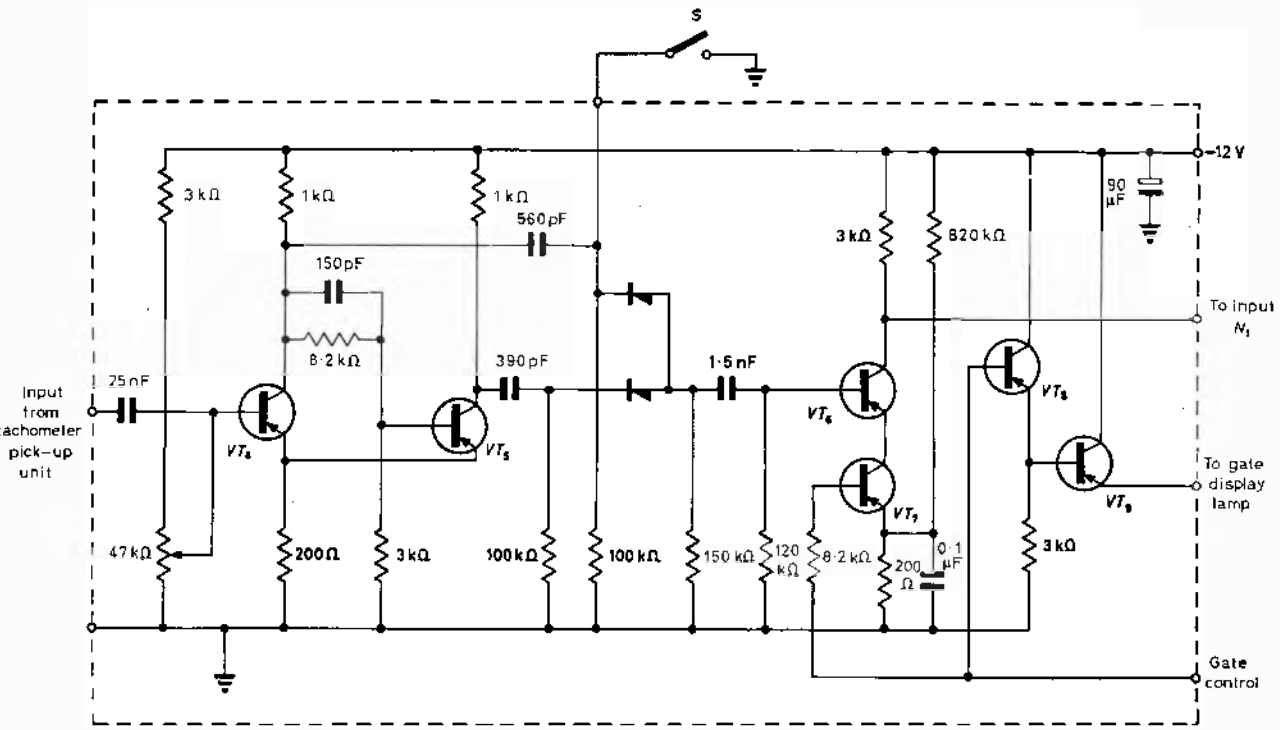


Fig. 6. Input circuit and gate

panel of the instrument the figures indicating the mean slippage of the motor during the period of 2sec appear periodically, and can be read by the operator.

This basic principle of operation imposes no limitations regarding the accuracy of the instrument; using a greater number of pulses generated each revolution, or increasing the time interval, it may be possible to measure the slippage with any desired precision.

Description of the Instrument

A digital instrument for measuring the slippage of asynchronous motors according to the above described principle is shown in Fig. 3 and its detailed block diagram is shown in Fig. 4.

The tachymetric pick-up head, being connected to the main instrument by a flexible cable, has the form of a fork and contains the optical system, lamp and condenser lens in one branch, and the photo diode in the other. The

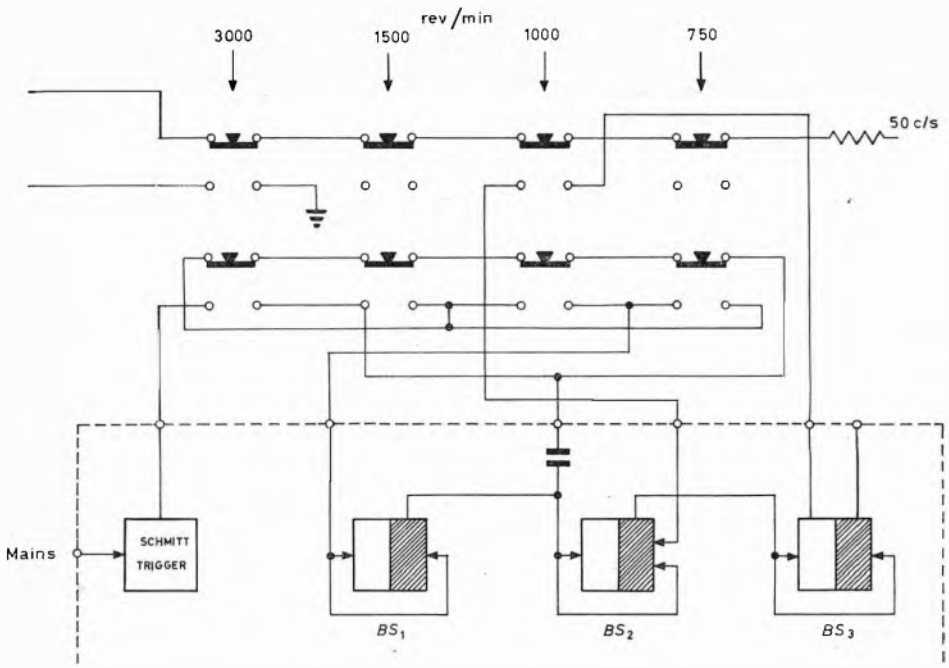
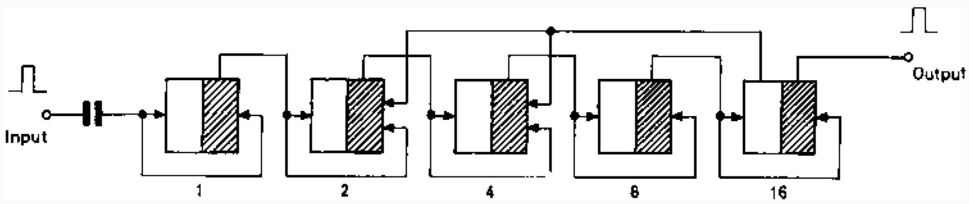


Fig. 7. Switch arrangement of the scale-of-eight circuit

Fig. 8. Schematic diagram of the scale-of-25 circuit



electronic circuit is a pre-amplier shown in Fig. 5, located in the handle of the fork. This amplifier uses silicon transistors and provides output pulses on a low impedance. This circuit also supplies the filtered d.c. voltage for the electric lamp.

The operator has to keep the tachymetric pick-up head in a position relative to the ten-hole disk mounted on the motor shaft, in a way to ensure the modulation of the light beam by these holes as the disk revolves, thus generating electric pulses which are fed to the electronic circuits of the instrument.

In the input circuit, the pulses are shaped with the Schmitt trigger VT_4-VT_5 of Fig. 6 and pass through the gate VT_6-VT_7 before reaching the first decimal scaling unit N_1. The switch S permits the doubling of the number of pulses, by an arrangement of the contacts of the push-

decimal scaling units, consisting of four cascaded flip-flops, working according to the 1-2-4-2 scaling scheme, and associated with a magnetic binary-to-decimal decoding matrix which drives an in-line projection-type numerical indicator.

A special test facility of the instrument is provided by feeding at the input the mains frequency; in this case the instrument should indicate a slippage of 0·0 per cent.

The operation of the instrument is extremely simple: the electronic block is connected to the mains supply and the synchronous speed of the motor under test is selected by pressing the corresponding push-button in the key-board switch. The photo sensitive fork is then held over the revolving disk on the motor shaft, and the actual slippage is periodically displayed on the luminous indicators on the front panel.

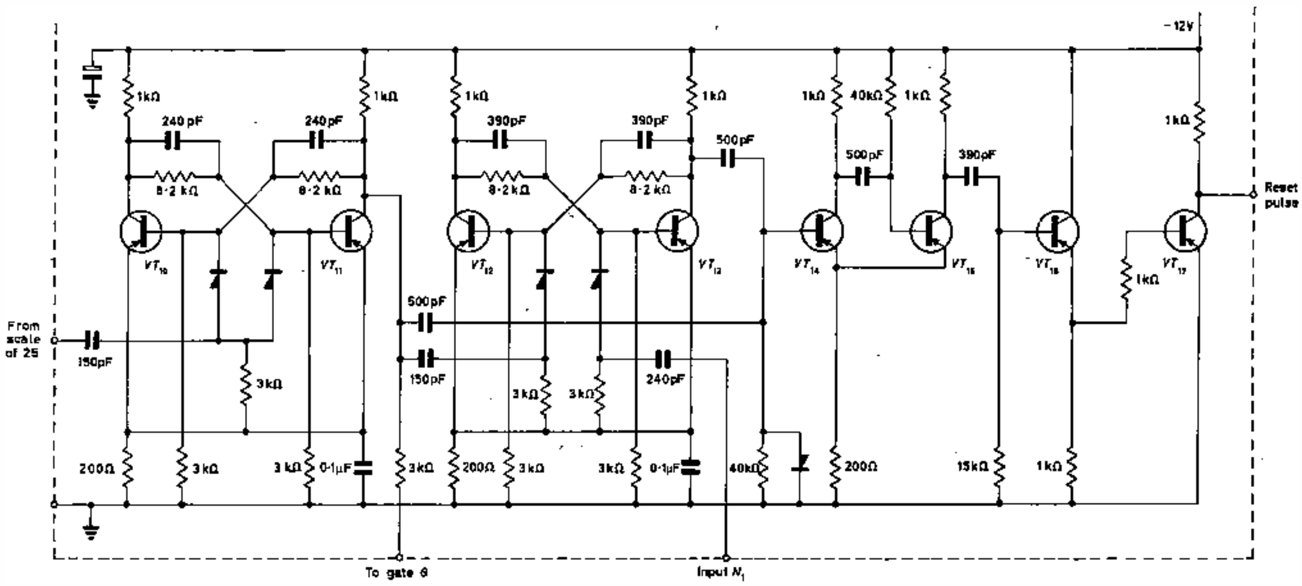


Fig. 9. Gating, error-compensating and reset circuits

button switch on the front panel, thus enabling the use of the same ten-hole disk for motors having 750, 1 000 1 500 and 3 000 rev/min synchronous speed.

These contacts, shown in Fig. 7 also affect the measuring interval, establishing it at 2sec for the 3 000 and 1 500rev/min range, at 3sec for 1 000 and 4sec for 750rev/min synchronous speed, by dividing the mains frequency by 100, 150 and 300, in order to bring the theoretical number of pulses, at zero slippage, to 1 000, no matter what kind of motor is being tested.

The division of the mains frequency is achieved by the cascade of the three flip-flops $BS_{1\text{-}2\text{-}3}$ in Fig. 7 and the scale-of-25 frequency divider consisting of five flip-flops connected as in Fig. 8.

The gate G is operated by the flip-flop VT_{10}-VT_{11} in Fig. 9 which also transmits a pulse to the monostable multivibrator VT_{14}-VT_{15}, thus generating over VT_{16} and VT_{17} the reset pulse for the counting circuits.

In order to eliminate the systematic error of $-0{\cdot}1$ per cent due to the method of subtraction used in the instrument, the flip-flop VT_{12}-VT_{13}, driven asymmetrically, generates a supplementary reset pulse at the first pulse that reaches the decade counting unit N_1, and thus lowers the total number of counts by one.

The counting circuits N_1 and N_2 are binary-coded

Conclusion

The instrument described in this article has been successfully tested in laboratories and factories, and the results obtained were compared with figures given by conventional methods; no discrepancies were noted. The use of this device presents the following advantageous features compared to other methods:

The direct and objective evaluation of the slippage in numerical form, excludes the errors associated with the time measurement involved in other methods.

The automatic and periodical measurement of the slippage, at small time intervals, permits the study of dynamic processes in the behaviour of the motor.

The instrument is not sensitive at mains frequency variations, giving accurate results even if the frequency shifts.

Acknowledgments

The authors wish to express their gratitude to the Directors of the Institute for Atomic Physics, for facilities for carrying out the experimental work, and permission to publish the present article. Since writing this article it has become known that there is a French patent[1] on a similar device.

REFERENCE

1. Patent No 1 169 422/1958 (France).

High Frequency Multiplication Using Logarithmic Diodes

By R. P. Ingram, B.Sc.(Eng.), C.Eng., M.I.E.E.

A method of multiplying at high frequencies, using a bridge-ring modulator of logarithmic diodes, is described. The theory indicates that, provided certain drive and output conditions are fulfilled, then a very good approximation to true multiplication results. Such multipliers can be constructed to function up to 100kHz with little difficulty, and with a suitable choice of diodes fairly good results may be obtained at higher frequencies. Devices of this kind may well be useful for correlation analysis, and for use in scanning systems.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

BRIDGE-RING modulators have been used before for multiplication at high frequencies, but this article presents a new approach to their design, and the theory evolved demonstrates both the usefulness of this method, and its limitations.

Theory

For a typical logarithmic diode the relationship between current and applied voltage is accurately given, over a wide range, by:

$$I = 10^{-A}\,(e^{BV} - 1) \quad \ldots\ldots\ldots\ldots\ldots (1)$$

where I is the current in amperes.

V is the voltage in volts.

A is a constant, typically 10.

B is a constant, typically 23.

Over the logarithmic range of the diode, the simplification $I = 10^{-A}\,e^{BV}$ is sufficiently accurate ....(2)

Hence $\log AI = BV$

and $BV = \ln I + A \ln 10$

The dynamic resistance of the diode, r_d is given by dV/dI

$$dV/dI = 1/BI \quad \ldots\ldots\ldots\ldots\ldots (3)$$

Thus it can be stated that, for a logarithmic diode, the dynamic resistance is inversely proportional to the current, over the logarithmic portion of the characteristic.

Consider now a bridge ring modulator made up of four logarithmic diodes as shown in Fig. 1. The input signals to the modulator consist of a small voltage signal E_M, supplied from a low impedance source, and a large current signal I_c, which is supplied from a high source impedance. The diode resistance, for a single diode carrying a given current, is denoted by R_d, and its dynamic or slope resistance by r_d.

The gross transformer turns ratios are taken as 1 to 1 for simplicity.

The effect of the current signal or carrier is alternately to pass a large current through diode paths *fce* and *edf*. During each half cycle of the carrier current excursion, point *b* is, in effect, connected alternately to *c* and then to *d*, without actually inducing any signal into the paths *bc* and *bd*. The signal input is thus routed through to the output via paths *acb* and *adb*, so that the signal phase direction is alternated by the carrier. The output current magnitude may be deduced by considering the situation at an instant during one half carrier cycle.

The equivalent circuit is shown in Fig. 2. Notice that although the diodes appear in series to the carrier current path (with a total resistance of $2R_d$), they are effectively in parallel to the signal and output path.

The current through the signal and output path is given by:

$$I_o = (E_M/2)\,.\,\frac{1}{((R_s/4)+(r_d/2)+R_o)} = \frac{E_M}{(R_s/2)+r_d+2R_o} \quad \ldots (4)$$

If the source and output resistances are made very low; that is $r_d \gg R_s/2 + 2R_o$ then:

$$I_o \simeq E_M/r_d \quad \ldots\ldots\ldots\ldots\ldots (5)$$

but for logarithmic diodes $r_d = 1/BI$

so that

$$I_o \simeq BE_M I_c \quad \ldots\ldots\ldots\ldots\ldots (6)$$

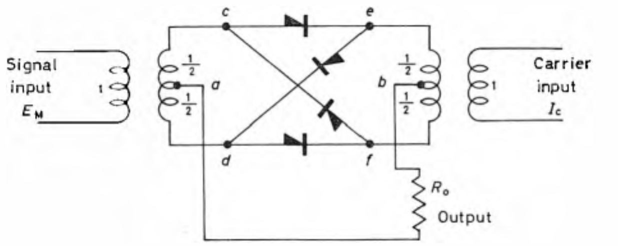


Fig. 1. The bridge ring modulator configuration

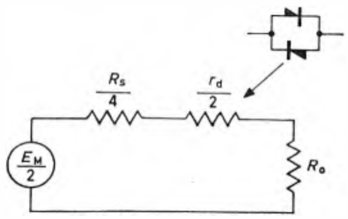


Fig. 2. The instantaneous equivalent circuit for one direction of carrier current flow

Hence the device multiplies for both phase and magnitude correctly. It is assumed in this simplified theory that the diodes are matched, that the signal current is small compared with the carrier current, and that the transformers are ideal. By 'ideal' is meant that the transformers are perfectly balanced, have zero leakage inductance and resistance, and that their magnetization current is zero. In fact, the transformer characteristics can, in practise, be controlled so that their properties impose no significant limitations on the multiplier.

Limitations

The main limitations stem from the facts that:

(a) The carrier current must to some extent be larger than the signal current.

(b) Unwanted products appear at the output, such as carrier and signal harmonics; some of which are increased relative to the desired output as the ratio of inputs is increased.

(c) The output and source resistances must be kept very low.

Upon examination it appears that (a) is not such a limitation as it would, at first, appear. As the signal current flows through the parallel combination of diodes, the total of carrier plus signal current is increased in one diode, and decreased in the other diode, so that the resulting value of $r_d/2$ is not affected until the signal current actually exceeds the carrier current. The effective resistance r_t of the combination is given by:

$$r_t = \frac{r_{d1} \times r_{d2}}{r_{d2} + r_{d2}} = \frac{\dfrac{1}{B(I_c+I_s)} \times \dfrac{1}{B(I_c-I_s)}}{\dfrac{1}{B(I_c+I_s)} + \dfrac{1}{B(I_c-I_s)}} = 1/2BI_c \quad \ldots . \ (7)$$

have an output range of $2X$dB if no output is to be discarded.

The dynamic range of the equipment following the multiplier, the logarithmic range of the diodes, and the relationship of signal to carrier current levels stemming from (a) and (b), are likely to be the factors setting the values of drive to the multiplier.

In practice unwanted products of the kind mentioned, due to strays and unbalanced diodes, can further be reduced by filtration. If each input frequency band is say 100 to 110kHz, then the multiplier true outputs will be in the bands 0 to 10kHz and 200 to 220kHz. It is usually the low frequency band which is of interest, and this can be separated from the higher output frequency by means of a low-pass filter. Fundamentals and harmonics of the input signals also will be attenuated by the filter.

With regard to (c), the requirement that $r_d \gg (R_s/2)+2R_o$

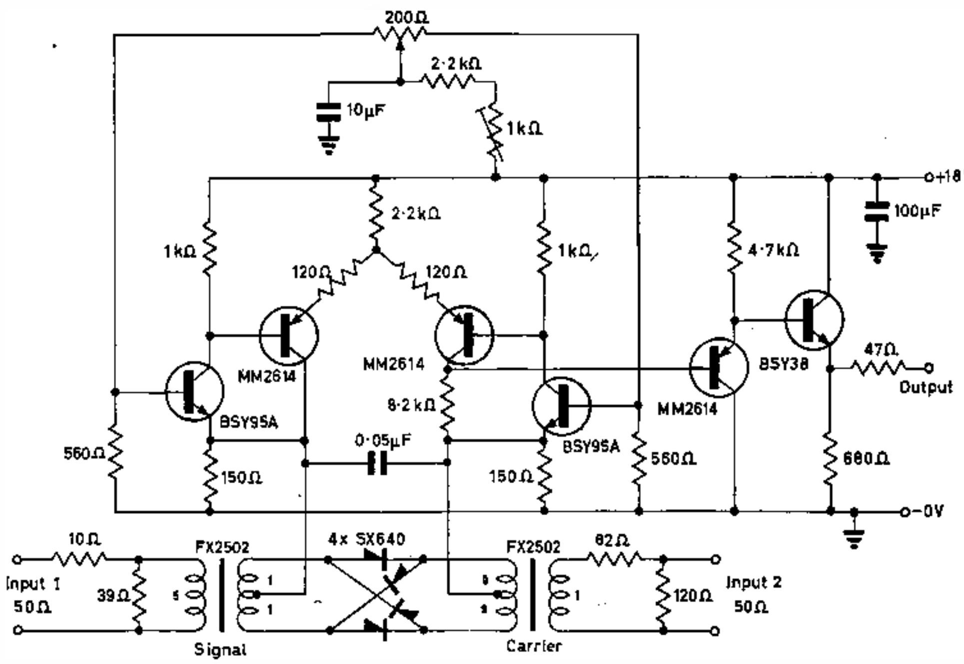


Fig. 3. A practical multiplier with driving transformers and a directly coupled output amplifier having low input impedance

where r_{d1} and r_{d2} are the separate diode dynamic resistances.

If the signal voltage is increased to the extent that the signal current exceeds the carrier current then the multiplying action tends to interchange; the signal acts as a switching carrier, and the carrier acts as a signal. Although the direction of output current flow remains correctly phased, the magnitude is no longer multiplicative, but is equal to twice the input carrier current. The device has then become a normal switching modulator.

As (b) suggests that the signal and carrier current levels should be as nearly equal as possible, the maximum signal current should then equal the smallest carrier current likely to be required to produce a noticeable output. For the applications mentioned the multiplier outputs are usually only of interest when there are two appreciable coincidental inputs of the same phase; the signal and carrier currents through the diodes can be made almost equal then. It is interesting to notice that for a given dynamic range of say XdB for each input signal a multiplier should ideally

indicates that the signal source resistance should be low compared with the minimum diode resistance, and that the output load resistance should be even lower by a factor of four times. To be more exact, it is the source and load impedances which are concerned, but these are usually mainly resistive.

The low source impedance may be obtained with the use of emitter followers or step-down transformers. In some cases, where zero frequency is of no interest, the output may also be taken through a suitable transformer in order to obtain the low load impedance. In general a more satisfactory method is to use a direct coupled circuit of the form shown in Fig. 3, which presents a low input resistance, and operates over a wide frequency range extending to zero. The input resistance of this particular arrangement is approximately 1·5Ω, and a typical diode dynamic resistance, for a current of 2mA, is 22Ω.

A high carrier source impedance is desirable in order to retain the correct carrier current in spite of the considerable variations in loading by the diodes.

It would appear that of the logarithmic diodes which have been available so far, capacitance and storage effects limit their use to about 100kHz. Ordinary high frequency diodes have successfully been tried at frequencies of 500kHz, and 1MHz with slightly reduced performance. A typical form for the dynamic resistance of an ordinary diode is: $r_d = 1/7{\cdot}71^{0{\cdot}9}$, which is only slightly different from the true logarithmic relationship.

Practical Results

The circuit shown in Fig. 3 has been constructed and tested. The carrier current is obtained from a high impedance step-up transformer having a flat response from 200kHz to 2MHz. The signal drive voltage is similarly supplied, but from a low impedance step-down transformer.

The output from the multiplier is fed directly into the emitters of a pair of transistors in the amplifier. The emitter input resistance is low using the grounded base connexion, but the input resistance is further reduced by the application of current feedback which emanates, in a symmetrical fashion, from the collectors of a long-tailed pair of transistors. The amplifier output voltage is obtained from the potential difference created by this feedback current flowing through a collector resistance. Emitter-followers are used to reduce the output resistance, and consequently to enable a usable output current to be taken.

Using different types of diodes and drive levels, in the multiplier, good results were obtained at frequencies up to 500kHz, but at 1MHz the sensitivity was reduced. When the same input signal was applied to both input points the multiplier 'squared' with an output error of within 7 per cent of the exact squared value over an output range of 7 to 1 at 1MHz. Virtually no filtration was used, and particularly with a lower level of signal input, components of output were detected which resulted directly from the inputs. Measurements were mainly confined to 1MHz, and it is believed that considerably more accurate results would be obtained at the lower frequencies with increased high frequency attenuation.

The output varied in proportion to individual inputs, and when inputs of slightly differing frequencies were applied, the characteristic beat pattern was seen on a display of the output signal.

Reversed Connexion

There is a dual arrangement to that which is indicated in Fig. 1. The current drive may be applied at terminals *ab*, and the output taken from the right-hand transformer, provided that a direct current output is not required.

The principle is the same as before except that alternate pairs of diodes in paths *ce*, *df*, and *ed*, *fc* conduct the carrier current in parallel pairs. One advantage of this arrangement is that the low source and output resistances are relative to the forward resistance of two diodes in series with the signal path.

If the diodes are matched, then the carrier current will be shared equally between each pair of diodes.

Conclusion

The low source and output impedances necessary to allow the correct multiplicative action to occur can be obtained by the techniques outlined. By setting the signal and carrier current levels to be as nearly equal as the application permits, and with the use of frequency separation, unwanted output products may be reduced to an insignificant level. Thus by careful design the effects of limitations, which are not stringent, may almost be eliminated, and this basic arrangement using logarithmic diodes will function satisfactorily at selected frequencies, up to 100kHz.

Acknowledgments

The ideas for this multiplier were stimulated by consideration of the projects of Messrs. Coastal Radio Ltd, and the author would like to thank them for their permission to publish this article.

New Input/Output Equipment for the Argus Computer

A new range of microminiature, modular input/output and peripheral equipment together with a new range of c.r.t. display systems has been developed by the Automation Systems Division of Ferranti Ltd, Wythenshawe, Lancashire, for use with the company's ARGUS 400 and 500 microminiature digital computers.

The following is a list of the input/output modules:

(1) On-line analogue and digital inputs and outputs.
(2) Backing stores—drums, disks and magnetic tape.
(3) Peripherals—printers, tape punches and readers.
(4) Special computer control units such as operators' panels, direct digital control output units, back-up controllers, timing and isolate systems.
(5) Data link systems and computer/computer links.

The range, which must cater for all the possible types of input and output signals encountered in process control, comprises over 100 different modules.

The form of construction adopted is based on plug-in module cards. High reliability and low cost are achieved by the use of integrated circuit logical elements, and by the extensive use of printed circuit wiring.

Maximum flexibility is ensured by the use of a standard interface highway to connect the input/output modules to the computer. This interface is identical for both the Argus 400 and 500 machines, and is not modified in any way by the number or types of input/output module employed in any given system.

Six different types of computer controlled c.r.t. display systems have also been produced by Ferranti Ltd, to provide a more economical solution to the wide variety of display requirements encountered in both industry and commerce, and in the military field.

Designed to provide high quality presentation of alpha/numeric characters, special symbols and pictorial information in a wide range of tabular and graphical formats, the display systems work in conjunction with any of the standard Ferranti Argus range of digital computers.

The six basic display systems are as follows:

System 10. An extremely simple, programme-controlled point plotting display system suitable, for example, for laboratory use where a rapid means of displaying experimental results is required.

System 20. A general purpose point plotting display system for presenting complex data. Where spectrum analysis of experimental data is required it is particularly useful.

System 30. A system providing comprehensive facilities for the presentation of alpha/numeric tabular formats and information in pictorial form for process monitoring and control, information retrieval and man-computer communication.

System 40. The facilities and applications are similar to System 30 except for the incorporation of a buffer store to reduce computer loading where a large amount of information is to be displayed.

System 50. An alpha/numeric display utilizing a fixed format of 13 lines with 64 characters per line which is particularly suitable for two-way communication with a central processor, locally or remotely situated, as it incorporates an alpha/numeric keyboard for data entry, verification and message acknowledgement.

System 60. The facilities and applications are similar to System 50 except that buffer storage is incorporated to reduce computer loading where a large number of display outlets is required.

A Novel Voltage Reference Source

By T. P. Dawson*, Dip.Tech., and G. C. Taylor*, D.I.C., A.C.G.I., B.Sc.

A bridge configuration voltage stabilizing circuit employing semiconductor devices has been developed to give stabilization ratios in excess of 700, compared with ratios of about 25 for conventional stabilizing circuits using two Zener diodes.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

THE use of the bridge configuration in stabilizing circuits is well known[1,2], but little information is available for such circuits employing semiconductor devices.

It can be shown[3], that if two opposite arms of a bridge circuit contain 'constant current type' devices while the other two contain 'constant voltage type' devices, then the output voltage of the bridge is independent of the input voltage over the range of operation for which the bridge remains balanced with respect to the incremental resistances of the arms. A typical circuit is shown in Fig. 1, Zener diodes and suitably biased transistors providing the 'constant voltage' and 'constant current' arms respectively. The resistors R_B are necessary to balance the circuit. If the slope resistance of the Zener diodes is small, then the value of R_B is approximately equal to the incremental resistance of each transistor arm. With Zener diode biasing this resistance is of the order of 1kΩ: but with battery biasing it is so high that it is necessary to use an emitter-collector shunt to reduce the slope resistance to a reasonable level in order to avoid a very large voltage drop across R_B.

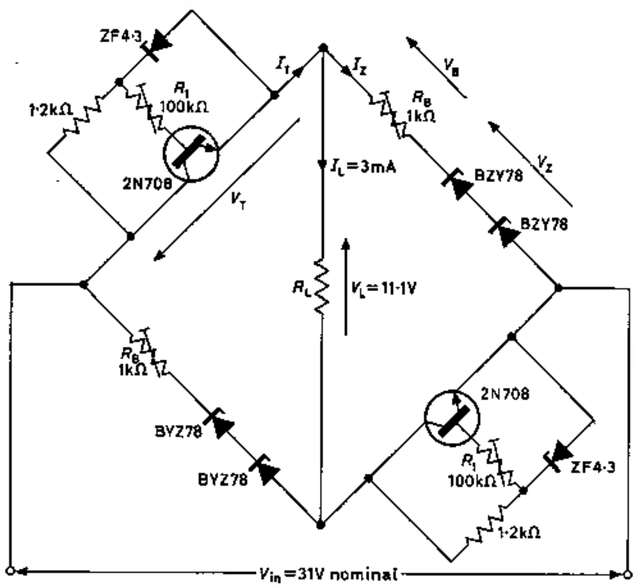


Fig. 1. Typical circuit arrangement
R_1 adjusts transistor bias

Due to the presence of the balancing resistors, R_B, the output impedance of the circuit is high, and this restricts its use to that of a voltage reference source as V_L is not stabilized against changes in R_L.

To obtain a required output voltage, V_L, the operating conditions of the transistors (V_T and I_T), are chosen to suit the nominal input voltage. V_Z is determined by the available Zener diodes and hence I_z must be adjusted to produce the value of V_{RB} which gives the required output voltage, $V_L = V_Z + V_{RB} - V_T$. The necessary load resistance is readily calculated after the load current has been found by subtracting I_Z from I_T.

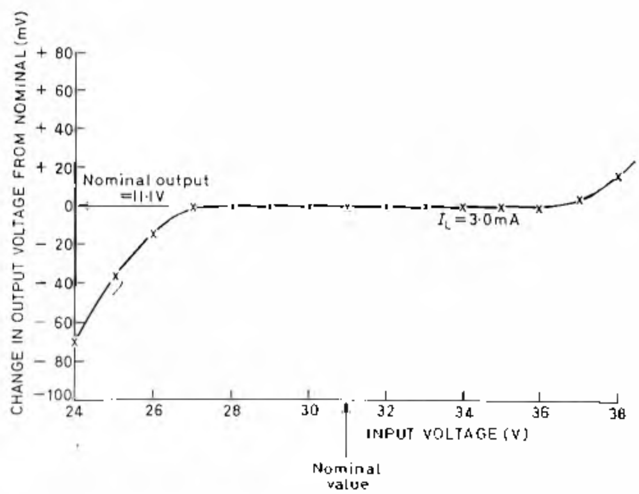


Fig. 2. Change in output voltage for varying input voltage for circuit of Fig. 1

Typical results for the circuit of Fig. 1 are shown in Fig. 2. Over the range of linear operation a ±4V (13 per cent) change in input voltage resulted in an output voltage change of less than ±2mV: giving a stabilization ratio $(\delta V_{in}/V_{in} \div \delta V_o/V_o) \simeq 720$. A similar circuit using fixed resistors in place of the transistor arms also gave good stabilization but this was restricted to a ±4 per cent range of input voltage only.

These results compare favourably with those obtained from two Zener diodes in cascade which, when working under similar conditions, gave a stabilization ratio of 23.

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* *University of Surrey.*

A Voice-Operated Watch

By K. J. Kapota*

The timing component is an electric watch and this is switched on and off via a throat microphone and electronic circuit to provide a measure of the time which a person spends in talking. The device is intended for use in the fields of psychology and psychiatry.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

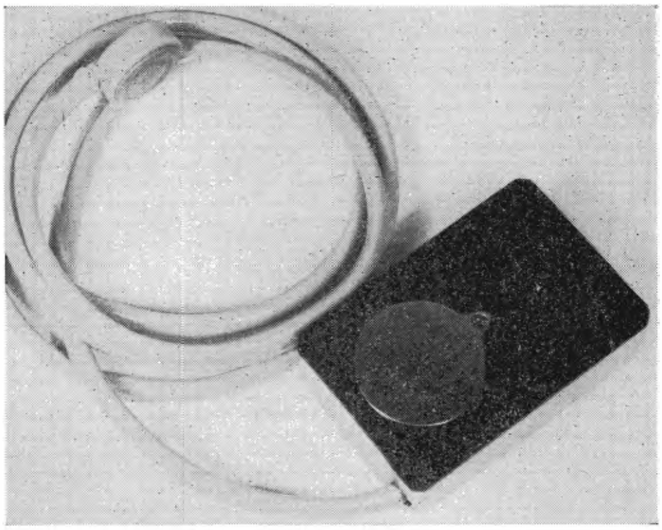
THE amount of time a person spends talking during the day is at present unknown. This aspect of human behaviour is of interest in a number of fields in psychology and psychiatry, and a device has been developed which provides a measure of time spent in speech. A small throat microphone[1] is used to transmit the speech to a unit about the size of a cigarette packet, containing the watch. As seen from Fig. 1, the timing unit can conveniently be kept in the subject's pocket, and is easily inspected at any time during the day.

The main component of the timing system is an electric watch[2], which can be controlled externally. A circuit, using hearing aid transistors[3] was developed, and is shown in Fig. 2, consisting of a microphone amplifier, rectifier and control switch. The amplifier is a three transistor d.c. feedback type, having a minimum gain of 40dB, and a frequency response from 500Hz to 10kHz. The output of the amplifier is rectified and used to drive VT_4 which acts as a switch for the collector current of VT_5. The oscillator VT_5 is an integral part of the watch mechanism, only the diode connexions being available externally.

Discrimination against noise, due to skin movement, swallowing or eating, is obtained by the bias applied to VT_4. Also the time-constant RC reduces transients outside the speech range, which would normally overcome the bias of VT_4.

Preliminary tests showed the device operated satisfactorily, was comfortable and unobtrusive to wear. The 2·5V battery life was approximately six weeks while the watch battery should operate for at least 12 months.

* *University of Oxford.*



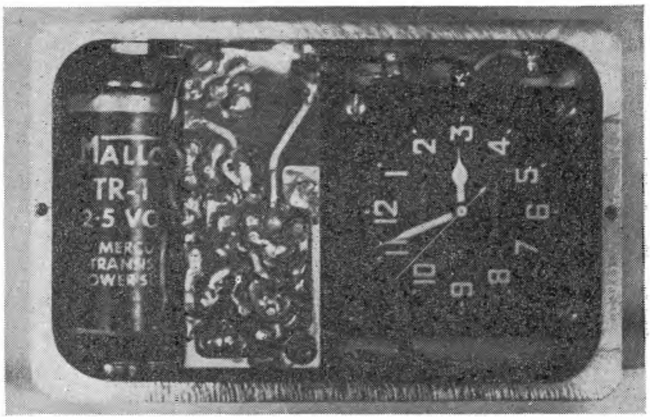


Fig. 1. The complete instrument and an interval view of the timing unit

Acknowledgment

The voice operated watch was designed and built for Mr. R. C. Oldfield, at the Institute of Experimental Psychology, University of Oxford. The author wishes to express his thanks to Mr. Oldfield who will publish details of the use of the device at a later date.

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3. HAIO selected transistors, S.G.S. Fairchild.

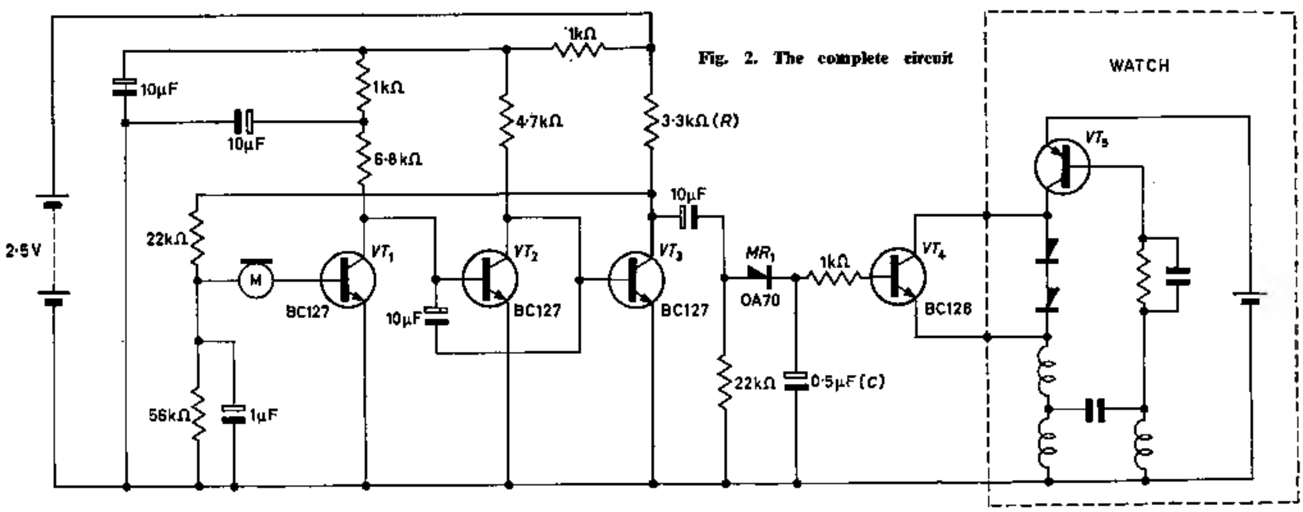


Fig. 2. The complete circuit

SHORT NEWS ITEMS

A conference on integrated circuits, sponsored jointly by the I.E.E. Electronics Division, the Institution of Electronic and Radio Engineers, and the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section), will be held at the Congress Theatre, Eastbourne, on 2-4 May this year.

The conference will discuss all current aspects of microelectronics whether semiconductor, thin film, thick film or hybrid under the following headings:

(a) Basic technological processes and design (including materials).
(b) Performance characterization and properties.
(c) Applications.
(d) Advanced techniques and devices.

Registration forms and further information are available from the Conference Section, I.E.E., Savoy Place, London, W.C.2.

A one-week course is to be held on 10-14 April this year in the Department of Glass Technology of Sheffield University on 'Glass Technology for Workers in the Electronics Industry'. The purpose of the course is to provide an introduction to those aspects of the science and technology of glass which are relevant to the industry. Topics to be covered include glass-to-metal sealing, the physics of annealing, electrical properties, strength, infra-red and ultra-violet transmission and new materials such as 'pyroceramş' and laser glasses. Some practical work will be included.

Accommodation will be provided in one of the University Halls of Residence.

Further details may be obtained on application to Professor R. W. Douglas, Department of Glass Technology, 'Elmfield', Northumberland Road, Sheffield 10.

A conference on 'On-line Measurement and Inspection—their Impact on Quality', organized by the Society of Instrument Technology, is to be held at St. Andrews, Scotland, on 20 to 22 April this year.

Topics expected to be covered include: Moisture determination and control in the food industries; particle size determination; flaw detection; instrumentation in the leather industry; automatic inspection of surfaces; colour measurement in the food industries; sampling of paper and pulp; control of fermentation; analysis for mineral processing; viscometry in food; fat determination; automatic inspection of discrete objects.

Papers will be accepted up to the end of March providing they cover new techniques or extend known techniques to new areas of industry.

Registration forms are available from the Secretary, Society of Instrument Technology, 20 Peel Street, London, W.8.

The Scottish Sections of the Institution of Electronic and Radio Engineers and the Institution of Electrical Engineers are holding a symposium on 'Telemetry in Medical and Biological Research' at the Royal Infirmary, Lauriston Place, Edinburgh 3, on 9 March this year.

Further information and delegates' registration forms may be obtained from P. M. Elliott, Honorary Secretary, Scottish Section, I.E.R.E., 21 Craigmount Loan, Corstorphine, Edinburgh 12.

The Fourth International Measurement Congress (IMEKO IV) is to be held in the Palace of Science and Culture, Warsaw, Poland, on 3-8 July this year.

Further information regarding registration is obtainable from the IMEKO IV Conference Committee, NOT, Ul. Czachiego 3-5, Warsaw, Poland, and hotel accommodation from the ORBIS Travel Bureau, Krakowskie Przedmiescie 13, Warsaw.

The International Measurement Show (IMIS 1967) will also be held at the Palace of Science and Culture on 3-11 July.

A Chair in Telecommunications Systems in the Department of Engineering Science at the University of Essex is to be established following an agreement between the University and the Post Office.

The Post Office will pay the costs of the Chair and some supporting staff for ten years, at a cost which is likely to be in the region of £100 000.

It is hoped that the Chair will be filled by October this year, in time for the beginning of the next academic year.

The Chair will provide teaching at both undergraduate and post-graduate levels and conduct research in telecommunications systems engineering. It is hoped that the Chair will take a broad view of its task and develop, as fully as possible, inter-disciplinary courses with other departments.

It is expected that close ties between the Chair and the Post Office Research Station, which will be moving from London to Martlesham Heath, near Ipswich, will be developed which could lead to a valuable interchange of men, material, support and ideas.

The Plessey Co Ltd has announced the formation of the Dielectric, Magnetic and Capacitor (DMC) Divisions in its Components Group. The DMC Divisions have been formed from the existing Towcester Divisions, the United Insulator Company Limited, and the Telegraph Condenser Company Limited (TCC) Divisions at Acton and Bathgate.

A data analysis centre, to be set up at Southampton University's Institute of Sound and Vibration Research with the aid of a Science Research Council grant of £55 000 to Professor B. L. Clarkson and Dr. C. A. Mercer, will provide new facilities for university and industrial research. The new equipment is required for the rapid high-resolution analysis of random signals arising from the Institute's own research programme and from the collaborative work it undertakes jointly with industry and research establishments.

An on-line digital computer will be used for work on the detailed structure of turbulence, noise from turbulence and machinery, random excitation of lightly damped structures and fatigue under random loading. The computer will have two high-speed sampling units and an exchangeable disk storage system. The results of each analysis will be plotted automatically by the computer and its associated equipment. Some important fields of application are the environmental and reliability testing of a wide range of products from missiles and aircraft to consumer appliances, investigations of motor vehicle and ship noise and vibration, machine tool performance, wind loading of buildings and many aspects of nuclear power station design and operation.

The Columbia Broadcasting System has placed a new order with The Marconi Company for 27 of the latest Mark VII colour television cameras. This order, which includes colour coding equipment using the American NTSC system, follows an earlier one for six cameras which have already been delivered.

The Mark VII a fully transistorized colour television camera, employs four photo-conductive camera tubes, which provide high quality colour or black-and-white pictures which are less dependent on accurate tube registration than with three tube cameras. Advanced 'Thin film' circuits are used to provide sufficient stability for 'hands off' operation, using only a simple control panel mounted in the studio suite.

'Staffing a computer installation', is a new brochure published by English Electric Leo Marconi. It describes the five basic functions essential to a successful installation (systems analysis, programming, data preparation, computer operating and maintenance) and sets out in some detail the staffing requirements of each.

It offers advice on the sort of qualities to look for in selecting staff and discusses the use of aptitude tests. The company has devised a series of special written tests of this type and arrangements can be made to hold these on a users premises, with experienced interviewers assisting the user in selecting staff.

Copies of the brochure are available from English Electric-Leo-Marconi Computers Ltd, Portland House, Stag Place, London, S.W.1.

The Solartron Electronic Group has received a £15 000 order from the United Nations in New York for a Marine Radar Simulator which has now been shipped to Nationalist China where it will be used by the U.N. Maritime Development Unit to teach navigation in the China Sea.

The Company has also received a similar order from the General Botha Nautical Academy in Cape Town. Due for delivery next May, this merchant marine aid will provide two 'own' ships and four 'targets'.

The Office for Scientific and Technical Information (OSTI) has awarded a grant of nearly £33 000 to the Hatfield College of Technology for the establishment and operation of a National Reprographic Centre for Documentation.

This Centre will be particularly concerned with photographic methods of reducing original documents to microforms (such as microfilm and microfiche) for storage, handling, retrieval and enlargement.

Its main functions will be:

(1) To act as a national clearing house for information on microrecording and associated reprographic techniques. All relevant published information will be collected, evaluated and abstracted and will be disseminated to interested users.
(2) To test and evaluate equipment on the British market and to coordinate this evaluation with similar work being undertaken by sponsored organizations in the U.S.A. and the Netherlands.
(3) To examine and evaluate users' needs and encourage research and development to meet them.
(4) To identify areas where further research and development is necessary and to sponsor and, where appropriate, carry out such work.

The activities of the centre will be carried out by a small team with appropriate library, design and photographic experience and they will be guided by an advisory committee which will include representatives of ASLIB, the Library Association, the Institute of Reprographic Technology, the Microfilm Association of Great Britain and the Office for Scientific and Technical Information.

The Royal Swedish PTT has ordered from Telefunken of West Germany five automatic h.f., s.s.b. transmitters—one rated 100kW and four 30kW each—as well as a transmitter remote control system for the overseas radio station at Grimeton.

The transmitters will be remotely controlled from the Stockholm telegraph office 500km distant, i.e. all the adjustments required in operation such as type of service, frequency, antenna pattern direction, etc., will be undertaken there.

When the transmitters have been installed two Telefunken transmitters of the same type which have been operated at Grimeton for some years will be dismantled and installed in the overseas radio station Karsborg, which will then have a total of eight Telefunken transmitters in service.

The British Electrical and Allied Manufacturers' Association, the Electronic Engineering Association and the Scientific Instrument Manufacturers' Association have decided to establish a joint organization to co-ordinate the activities of their member companies and other manufacturers concerned with technological training and educational aids.

The purpose is to promote this development market at home and abroad, and to provide a liaison with government departments and close co-operation with education and training bodies throughout the world.

Figures published by the Electronic Engineering Association for the first six months of 1966 show that manufacturers' deliveries of electronic capital equipment have broken previous records.

Provisional figures show that total deliveries climbed to an all-time record of over £162M, more than 30 per cent higher than in the first half of 1965. Over the same period, exports attained £52M, well over a third higher than the comparable 1965 figure. The export proportion was edged up to 32 per cent of total deliveries.

Deliveries of radio communications equipment attained £21M, more than a third higher than in the first six months of 1965 and exports of £8·6M were up by more than 50 per cent. Deliveries of radar and navigational aid equipment reached an all-time high of almost £40M, 20 per cent higher than in the same period of 1965, and included export deliveries of £15M.

Computer deliveries were increased by 50 per cent and attained £24M in the six months' period. Computer deliveries of £8M in the six months were well in excess of anything previously achieved in a whole 12 months. Deliveries of electronic measuring and testing equipment were up by a fifth to almost £15M with exports increased by 30 per cent to over £6M. Electronic industrial control equipment deliveries achieved £31M, about a sixth higher than in the same period of 1965.

Hilger & Watts Ltd has prepared a set of guide wavelength tables for its associated company, Microwave Instruments Ltd. The tables are computed for twelve waveguide sizes between WG10 (S band, 10cm wavelength) and WG32 (1mm). They give guide wavelength against free-space wavelength, calculated for the cut-off wavelength of the dominant H_{01} mode. All wavelengths are given to three places of decimals.

The figures given are for rectangular waveguide sections of all sizes. In addition the tables have been computed for circular guides in the dominant H_{01} mode for the Q band (8mm wavelength) and O band (4mm). The tables are applicable to all calibrated wavemeters manufactured by the Hilger and Watts organization.

Information is also given on standard U.K. and U.S. designations for all waveguide sizes within the ranges listed, full waveguide dimensions for all bands, cut-off frequencies for the dominant H_{01} mode, recommended operating ranges, attenuation per unit length of guide, and power rating of the standard waveguide.

The price is £5 5s. 0d. (post free) from Microwave Instruments Ltd, Park Lane, Shiremoor, Northumberland.

The British Standards Institution has published a new standard B.S. 4061 'Dimensions of pot cores made of ferromagnetic oxides for use in telecommunications and allied electronic equipment' which specifies dimensions and tolerances necessary to ensure dimensional interchangeability of two ranges of pot cores made of ferromagnetic oxides.

Range 1 is an established British design, range 2 is in accordance with the international standard given in IEC Publication 133. Dimensions of gauges for both ranges are specified.

Appendices give the method of calculation of the effective parameters of pot cores, notes on the use of the effective parameters and a list of symbols.

These Appendices are in agreement with IEC Publication 205: Calculation of effective parameters of magnetic piece parts.

Copies of B.S. 4061 may be obtained from the B.S.I. Sales Department, Newton House, 101/113 Pentonville Road, London, N.1. Price 12s. each (postage 1s. extra will be charged to non-members).

The Nucleonics and Instrumentation Unit of the Plessey Automation Group's Data Equipment Division has set up its research and production facilities at the Group's headquarters at Poole, Dorset.

The unit is currently engaged on developing and producing advanced systems. Future expansion plans allow for activities in other fields including automatic weighing and process and material analysis.

The first American Hercules transport aircraft containing British electronic equipment has now been delivered to the Royal Air Force.

The British equipment, worth about £70 000 in each aircraft, replaces American equipment used in the earlier version of the Lockheed Hercules. It is mainly transistorized, thus increasing reliability while reducing maintenance costs and on-the-ground time.

The British equipment is as follows:

Cossor Electronics Ltd
- I.F.F./S.S.R.

Decca Navigator Co. Ltd
- Decca Navigator, Loran C/A, Doppler Navigator, Navigation Computer, Roller Map Display

Ekco Electronics Ltd
- Weather Radar

The Marconi Co. Ltd
- Automatic Direction Finder, V.O.R./I.L.S. Receiver, v.h.f. Communications

The Plessey Co. Ltd
- u.h.f. v.h.f. communications

Smiths Industries Ltd
- Automatic Pilot, Periscope Sextant, Clock & AS I

Standard Telephones & Cables Ltd
- Radio Altimeter

Ultra Electronics Ltd
- Interphone and Cabin Address

The Post Office is exploring the possibilities of television links for conferences between two distant centres and also a vision-telephone for calls between individuals.

Already the GPO is conducting a market research, with the co-operation of the Institute of Office Management, among nearly 5 000 large industrial and commercial undertakings to assess the demand for both these services.

Both facilities would make use of the existing co-axial cable network and the microwave radio signal system now being developed to operate between a nationwide network of towers, of which the 620ft high Post Office Tower in London will be the hub.

The GPO is thinking of providing television conference links and vision-telephone services over the microwave system in the 1970's, and establishing at principal centres fully-equipped studios for television conference calls and booths for vision-telephone calls.

The GPO is to set up an experimental closed circuit television conference link in the near future between the Engineering Department headquarters in the City of London and the Research Station at Dollis Hill. Businessmen interested in the practical application of telephone vision conference facilities will be invited to take part in experimental, live conferences.

Individual person-to-person vision-telephone calls over the public telephone network would employ the equivalent of about 150 telephone circuits and give a head and shoulders picture on a screen about the size of a postcard.

Calls of this kind are possible now between telephone extensions of a private automatic exchange, and the first vision-telephones for connexion with private automatic exchanges are expected to be available from private manufacturers this year.

EMI Electronics Ltd has concluded a marketing arrangement with Compagnie Francaise Thomas Houston-HB to supply on an exclusive basis in the United Kingdom and certain other territories CFTH-HB flying spot telecine and slide scanning equipment; provision is also made for the manufacture under licence of this equipment in the United Kingdom. The arrangement means that not only can EMI Electronics Ltd install and commission this equipment but can also provide a comprehensive maintenance and after-sales service.

The Compagnie européenne d'Automatisme Electronique (CAE) has signed an agreement with Telefunken AG for the sale and maintenance of hybrid computing systems, including digital computers C 90-40 and C 90-80 manufactured by CAE and analogue computers RA 770 manufactured by Telefunken.

The agreement provides for each firm the marketing of hybrid systems in its own country and allows for a large collaboration in the design and implementing of this technique. The first system will be set up in Germany at the Institute for Mathematical technique of the Berlin Technical University.

This agreement underlines the European trends of Compagnie européenne d'Automatisme Electronique, an affiliate of the CITEC Group and of INTER-TECHNIQUE. CAE has already installed information processing systems in Germany, Belgium, Italy, Norway, the Netherlands and Switzerland.

A sound reinforcement system that will ensure maximum intelligibility in all parts of Canterbury Cathedral has been designed and installed by the Private Communications Division of Standard Telephones and Cables Limited.

Thirty-nine high quality loudspeakers are disposed at carefully chosen points throughout the building. They can be connected to a selection of 25 microphone circuits via specially designed amplifying and delaying equipment. The loudspeakers can be manually and automatically switched to ensure that maximum intelligibility is achieved as the pattern of events during a service changes.

The whole amplification system uses transistor equipment, all of the latest modular design and contained in a suite of two enclosed racks. Some of the modules are to special design to meet the needs of the audio delay system but all units can be easily withdrawn from the rack in the event of servicing being necessary.

The signal to be delayed is recorded on a band of magnetically loaded neoprene stretched around a precision machined drum rotating at 80rev/min. The reproduced signal is picked off by a series of playback heads appropriately spaced to correspond to the required time delay and all heads are in contact with the recording medium. A trace of silicone oil is used as a lubricant and upwards of 10 000 hours of use is expected before any wear is likely to be detected.

A system of automatic switching, entirely relay controlled from the selection of a microphone circuit, is applied to the loudspeakers to ensure that the correct group is brought into circuit with the appropriate time delays. Microphone circuit selection and certain (manual) loudspeaker selection together with overall volume adjustment can be carried out from two control panels located in the Cathedral furnishings in the Nave and Choir. From either position it is possible to relay to the other and in the equipment room beneath the tower steps the whole service may be tape recorded. Also from this position, previously recorded messages can be relayed over the entire loudspeaker network, a facility that will prove invaluable when wishing to give information to the many thousands of visitors that come to the Cathedral.

Muirhead facsimile transmission equipment is now being used by the New York State police to help fight crime.

This is the first comprehensive facsimile communication system to be installed by any United States police force, employing automatic transmitting machines for both photographs and documents.

The system speeds communications between New York City police and the State's Division of Identification at Albany. When a suspect is fingerprinted, identification can now be made, and a copy of his criminal record returned, within 2 to 3 hours instead of the previous 7 to 10 days.

The contract was awarded to Muirhead Instruments Inc., Mountainside, New Jersey, after 10 months of competitive evaluation, against strong American competition. Standard Muirhead equipment was modified by the Beckenham, Kent factory to meet the special requirements of the New York State Identification and Intelligence System.

NEW BOOKS

Principles of Linear Circuits

By E. A. Faulkner. 116 pp. Demy 8vo. Chapman and Hall Ltd. 1966. Price 25s.

THE first three chapters introduce briefly the basic concepts of electrical circuit theory, and although the treatment is lacking in depth it does provide the reader with a clear set of rules for analysing simple circuits. In chapter four the amplifier is introduced as a four terminal network and the frequency response and bandwidth are stated in terms of the voltage transfer function. The h parameter equivalent circuit of the linear amplifier is discussed and the resultant gain, input impedance and the output admittance quoted. A similar treatment of the Y parameters yield transfer, input and output admittances. The transistor is introduced as a circuit element in the common emitter configuration and biasing arrangements are discussed. Chapter five provides a superficial treatment of negative feedback, and various simple transfer functions are given together with their Nyquist plots. The effect of negative feedback on the gain and bandwidth of amplifiers, together with the effect on the input and output impedances is discussed and examples of transistor amplifier circuits with voltage and current feedback are given.

This book has apparently been written primarily for students in applied physics and to this end it no doubt serves a useful purpose. It is doubtful, however, whether it could be recommended to students in any branch of electrical engineering, as these students would, after one year of an undergraduate course, have covered far more circuit theory than is presented in this book. The major criticism of the book is its lack of bibliography, as this may be annoying to readers who wish to pursue further any of the material presented.

R. W. WHITEHEAD

Transistor Circuit, Design and Analysis

By E. Wolfendale. 292 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 70s.

IN recent years the subject of semiconductor circuit design has attracted a number of authors. Various approaches have been used but this is probably the only text that makes such extensive use of the hybrid Π equivalent circuit. Although a chapter is devoted to equivalent circuits and a number are described, apart from some analysis using the equivalent T network, Z, Y and h parameters are not mentioned elsewhere in the book.

One wonders how many practising engineers would use the hybrid Π equivalent circuit for the design of small signal, low frequency amplifiers. On the other hand this equivalent circuit is very useful for high frequency and pulse circuit analysis and the author gives a comprehensive treatment of these topics. It is worth noting here that the most recent revision of the Electronics Syllabus of the London External Degree has a specific reference to the hybrid Π equivalent circuit.

The preface states that the more advanced mathematical methods have deliberately been omitted for the benefit of older engineers and younger undergraduates. This is a point in favour of the book but the author tends to underrate the ability of the potential reader. There is little need for the detailed arithmetic, shown for example in Chapter One, and the conversion of complex numbers from the $a + jb$ to the $R\angle\theta$ form would greatly facilitate the solution of equations. In similar vane the author has made heavy weather of the series-parallel conversion of an *RC* network, since this may be derived in a few lines by initially stating the admittance instead of the impedance.

It is very easy to criticize a text and mask its true value. There is a lot of useful information on a variety of circuits in this volume and to the engineer who is already conversant with, or is willing to devote the time to understand the implications of the hybrid Π equivalent circuit, this book is a definite asset.

G. J. PRIDHAM

Electronic Engineering

By C. L. Alley and K. W. Atwood. Med. 8vo. John Wiley & Sons Inc. 1966. Price £4 14s.

THIS is a second, and somewhat revised, edition of the book originally published in 1962. Like the original edition, it attempts to give a unified treatment of valves and transistors bringing in, as a further step towards universality, complex frequency so that the complete response of the circuit can be obtained. This follows the growing tendency, especially in American textbooks, towards a unified treatment, for which this reviewer has little sympathy, especially when the natural processes of learning, logic and convenience are sacrificed on this particular altar. For example this approach leads the authors of the book to deal with grounded grid triode before the conventional circuit has been introduced simply because the grounded base transistor circuit appears more fundamental than the grounded emitter one. This sort of thing happens time and time again: on page 182 the authors are led to say: "Perhaps the most interesting behaviour of the triode tube is the current gain." They then go on to prove from the valve equivalent circuit, and by reference to such a strictly transistor term as α, that in the triode the current gain is infinite. What an enormous hammer to crack this nut! And not even the right hammer!

The book has its virtues. It covers a lot of ground even if the presentation is at times over elaborate. After a rather inadequate, for this decade, treatment of electron ballistics and semiconductor theory the book goes on to diodes, transistors and triodes, amplifiers untuned and tuned with and without feedback, small and large signal and to simple oscillators.

It then dives into Amplitude and Frequency Modulation and Demodulation and rushing through a very brief treatment of pulse circuits ends by packing in power supplies into its 730 pages of text, problems, appendices and index.

Perhaps the best way to characterize this book is to quote from page 133 where the authors calculate that $\alpha' =$ ·9839912 (α' is defined as $\alpha - (1 - \alpha\, r_b/r_c$ a new one to this reader). (*Sapientis sat.*

M. H. N. POTOK

Sequential-Circuit Synthesis: State Assignment Aspects

By D. R. Haring. 348 pp. Med. 8vo. M.I.T. Press. 1966.

THE problem of 'state assignment' is not particularly new, having been first recognized by D. A. Huffman in 1954. There have been many publications on the subject, mostly in the United States, and Dr. Haring lists sixty-six such references. However, the subject matter, and even the title, of Dr. Haring's book may be sufficiently abstruse to justify the unusual procedure of requiring definitions in this review.

A sequential circuit may be defined as

a switching circuit in which the present output is a function of both present and past inputs, the circuit thus possesses a memory. A sequential machine (or finite automaton) is a sequential circuit which is described by abstract symbols rather than switching variables. The state assignment for a sequential machine is the coding of the abstract symbols used in the sequential-machine definition of an information-processing system into the switching variables used in the sequential-circuit realization.

The choice of state assignment can have considerable effect on the properties of the sequential circuit including cost and reliability. It is therefore essential to find the state assignment that best satisfies some criterion of goodness under given constraints and for a given circuit technology. There is normally no one unique solution to this problem and an exhaustive search through all combinations is impossible even with digital computers since even a small sequential machine with only sixteen states has in the order of 10^{12} different possible state assigments.

Dr. Haring's declared principal intents, therefore, are to provide some insight into the general properties of state assignment and to develop some practical algorithms for selecting state assignments which satisfy useful engineering criteria.

His book thus fills a gap in the literature of the subject and represents an extension of his doctoral thesis submitted to the Department of Engineering at the Massachusetts Institute of Technology in 1962. It is directed at graduates with an interest in logic design or some aspects of communication and control. A background knowledge of modern algebra, and switching circuit theory is also assumed.

R. A. BONES

Phaselock Techniques

By F. M. Gardner. 182 pp. Med. 8vo. J. Wiley & Sons Ltd. 1966. Price 68s.

THE reception of radio signals from space vehicles or satellites presents one of the most challenging problems of modern communications. The signals are normally of low power and embedded in noise. In addition, since the velocity of the transmitter varies quite considerably with respect to the receiver, the frequency of the received signal varies because of Doppler shifts. With conventional reception techniques, the bandwidth would need to be unnecessarily large to accommodate these frequency changes, and the wide bandwidth would admit a large amount of noise.

The way in which this is overcome is to employ a phaselocked receiver, the local oscillator of which is locked by a negative feedback loop to the average frequency of the signal. The bandwidth can then be made very small, and the problems of noise are lessened.

The design of phaselocked loops, however, presents formidable difficulties involving both communication and control theory. This book consists mainly of an exposition of the mathematics of phaselocked loops.

After an introductory chapter, the basic transfer functions of second- and third-order loops are derived and root locus plots of the feedback system are examined. The third chapter is devoted to a consideration of the noise performance. The treatment is restricted to a linear analysis. In chapter 4 the problems of tracking, and 'acquisition' that is the initial 'lock-in' of the loop, are discussed. In chapter 5 the actual loop components are described, limiters, phase sensitive detectors, and voltage controlled oscillators. The optimization of loop performance is then considered in chapter 6.

In chapter 7, phaselocked receivers are introduced. The topics of importance are the effect of intermediate frequency amplifiers and automatic gain control. Other applications of phaselock are considered, and in a final chapter methods for testing a loop are given.

A useful set of appendices review the basic mathematics, formulae, and nomenclature, and a design example is given.

The subject is clearly presented and easy to understand. This book is a good introduction to phaselock techniques, and should be a useful reference on the theory of their design.

G. D. BERGMAN

Electronic Sensing Devices

By A. F. Giles. 158 pp. Med. 8vo. George Newnes Ltd. 1966. Price 45s.

This book reviews many of the known physical-electronic and chemical-electronic effects and shows how these effects are applied to the solution of sensing problems, in the control of industrial processes and systems.

The relation of these effects to solids, liquids and gases are discussed together with capacitance and magnetic induction effects.

The treatment is non-mathematical and assumes some knowledge of electronics.

Laplace Transforms for Electronic Engineers

By J. G. Holbrook. 347 pp. Demy 8vo. Pergamon Press. 1966. Price 65s.

The new subject matter included in the second edition of this book first published in 1959 deals with network analysis, electronic filters, specialized applications of the Laplace Transform and the synthesis of transfer function.

Plastic Insulating Materials

By I. M. Mayofis. 320 pp. Demy 8vo Iliffe Books Ltd. 1966. Price 55s.

The original version of this book was published in the U.S.S.R. in 1963 under the title 'Osnovy khimii dielektrikov' and has now been translated into English. It deals with the plastic insulating materials which are of specific use in the electrical and electronic industries.

A detailed explanation is given of the reactions, structure of polymers and physiochemical properties.

Directory of Electronic Circuits

By Matthew Mandl. 226 pp. Med. 8vo. Prentice/Hall International. 1967. Price 60s.

Some 150 of the more common electronic circuits are completely analysed and illustrated and including circuit component values, equations and formulae.

Emphasis is given to solid state circuits but valve circuits have been included for comparison purposes and in instances where they have specific application.

Fourth State of Matter

By B. Grycz. 120 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 15s.

Negative Absolute Temperatures

By M. Marvan. 70 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 7s. 6d.

These are the third and fourth in the series of 13 monographs dealing with advanced physics subjects, each written by a Czech physicist and originally published in Czechoslovakia.

The third of this series serves as an introduction to plasma physics, with applications in controlled thermonuclear reactions, high-temperature manufacturing techniques and space travel.

The fourth examines the whole concept of temperature and negative absolute temperatures, stating the new fundamentals of thermodynamics.

Radio Relay Systems

By Helmut Carl. 231 pp. Demy 8vo. Macdonald & Co. (Publishers) Ltd. 1966. Price 65s.

Based on the author's experience with Standard Elektrik Lorenz AG, Stuttgart, and his service with the Consultative Committees of the International Telecommunications Union, this book gives an overall picture of the principles and techniques of system planning and design, and describes the application of the engineering rules to be followed.

American Microelectronics Data 1966-1967

Edited by G. W. A. Dummer and J. Mackenzie Robertson. 1515 pp. Demy 4to. Pergamon Press. 1966. Price £13 10s.

This edition gives comprehensive detailed data on American micro-electronic systems, units and capabilities, as well as reliability testing and standards.

The main emphasis is on integrated circuits of the thin film and semiconductor types, but details of hybrid circuits are also included.

Also included is information on the specialized interconnexion and assembly techniques required for microelectronic equipment, together with a six-language glossary of microelectronic terms (English, French, German, Italian, Swedish and Spanish).

Introduction to Vector Analysis for Radio and Electronic Engineers

By W. D. Day. 260 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 42s.

Starting from basic definitions and notations, this book deals with the concepts of scalar and vector products of two vectors, triple products, differentiation, line and surface integrals.

Other subject matters include the differential equations of electron motion in uniform magnetic and electric fields.

Each chapter contains worked examples and exercises with answers.

Teleurope

Demy 4to. 1966. Teleurope, Holzhofallee 38, P.O.B. 320, Darmstadt, Germany

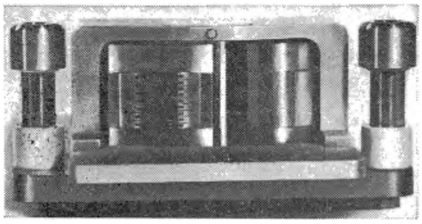
The 11th edition of this guide is divided into three parts and contains information on manufacturers in the 21 countries of the O.E.C.D. (Organization of Economic Cooperation and Development) engaged in export.

Part A gives an alphabetical list of manufacturers and full addresses in the various O.E.C.D. countries. Part B consists of a classified trade section and Part C is a list of telegraphic addresses.

NEW EQUIPMENT

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

(Voir page 127 pour la traduction en français; Deutsche Übersetzung Seite 134)



TAPE TESTING EQUIPMENT

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Illustrated above)

Gresham Lion Electronics Ltd has developed a new type of tape testing assembly which will test the full width of magnetic tape for 'drop-outs' and other defects.

Gresham is offering a wide choice of tracking arrangements and the version most suitable for computer work provides a full width record head for half inch tape followed by a dual replay head stack with seven and nine tracks (type P.S. 79). Top or bottom edge tape guidance is provided by spring loaded ceramic guides and head blocks are protected by an enclosed cover.

Electrical and mechanical specifications can be designed to suit specific applications and the specification of the seven and nine track replay head assembly for half inch tape is given below:

Record Head

Track width	0·505 + 0·001in
Gap length	0·0005in (nominal)
Resistance	25Ω
Inductance	7mH
Saturation current	33mA p-p

Replay Head	*7 track*	*9 track*
Track width	0·030±0·001in	0·040±0·001in
Gap length	0·00025in	0·00025in
Resistance	11·4Ω	12·5Ω
Inductance	5·6mH	5·3mH

EE 03 751 *for further details*

POTENTIOMETER SLAVE UNIT

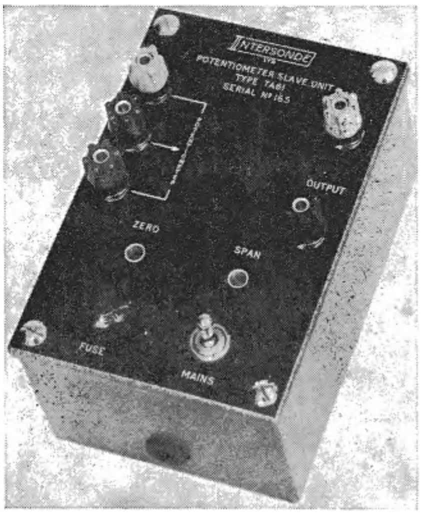
Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustrated above right)

The potentiometer slave unit type TA61 provides a moderately priced and simple means of energizing current operated loads from the output of potentiometer type transducers.

With this unit it now becomes possible to energize robust large-scale meters, pen recorders or galvanometer recorders from displacement, pressure and other transducers based on potentiometric sensing elements. The non-linearity error normally introduced when a current operated load is energized directly from the output of potentiometric transducers has been eliminated by basing the TA61 on a unique circuit design.

Although primarily intended for operation in conjunction with Intersonde tranducers the potentiometer slave unit may be used with any potentiometric transducer having a terminal resistance of between 500Ω and 10kΩ. A particularly useful feature of the unit, especially in industrial applications, is that it may be located many hundreds of yards distant from the transducer. The load connected to the unit may have a resistance of between 0 and 1 000Ω and may also be remotely located if required.



The TA61 is normally used in conjunction with transducers measuring static or slowly varying parameters but is equally suitable for dynamic applications. The frequency response of the units is flat from 0 to 10kHz and an accuracy of ±0·2 per cent of full range output is claimed over the temperature range 0 to 60°C.

EE 03 752 *for further details*

S.S.B. TRANSMITTER RECEIVER

Redifon Ltd, Broomhill Road, London, S.W.18

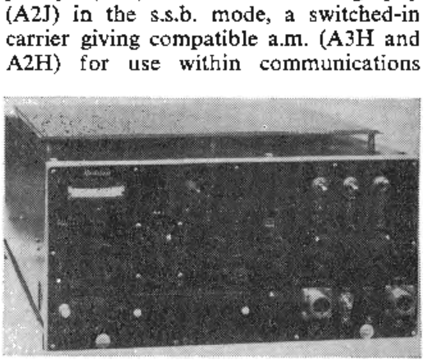
(Illustrated below)

Redifon Ltd Communications Division has developed a new 100W p.e.p. h.f., s.s.b. transmitter-receiver, type GR.410T, specifically to meet the needs of military and other security forces for maximum frequency flexibility at an economic price. It provides a half-way stage between existing a.m., d.s.b. equipment using v.f.o. frequency control and future s.s.b. equipment with full frequency synthesis, and is particularly intended for organizations that are considering the phasing-in of s.s.b.

The G.R. 410T economically combines the simplicity and stability of crystal control with the flexibility of front panel facilities for rapidly changing the crystals and re-tuning to new frequencies. Manual controls are provided for tuning the oscillator output and r.f. stages and for p.a. loading, while a built-in meter monitors drive level, p.a. anode current, forward aerial current, and supply voltages. The frequencies can be changed in a few minutes by operators in the field without maintenance personnel or workshop realignment.

Up to four switch-selected crystals can be fitted in the transmitter-receiver at any one time, and an operator can carry additional crystals with him for whatever other frequencies may be required, thus permitting rapid re-deployment to areas where different operating frequencies are in use. At the same time, the transmitter-receiver can accept an input from an external source, such as a frequency synthesizer.

The GR.410T covers the 2 to 16MHz frequency range, with switch selection of u.s.b. or l.s.b. and operates on telephony (A3J) and m.c.w. telegraphy (A2J) in the s.s.b. mode, a switched-in carrier giving compatible a.m. (A3H and A2H) for use within communications



networks not yet equipped for s.s.b. It is also capable of operation on c.w. (A1). Automatic send/receive switching can be selected by a switch on the front panel, and automatic level control prevents overmodulation while ensuring adequate modulation at low speech levels.

A Redifon manually tuned a.t.u. is available to match the transmitter-receiver into a vehicle-mounted whip or a simple wire aerial. The GR.410T can also be operated directly into a 75Ω dipole or into a 600Ω aerial via a wideband matching transformer.

Power units are available for 100 to 125, 200 to 250V, 50 to 60c/s mains or for operation from 12 or 24V batteries, enabling the GR.410T to be used in mobile, fixed or transportable roles.

The s.s.b. generator/demodulator, i.f. and audio units are constructed as plug-in modules and are secured by quick-release fasteners to permit rapid removal for servicing. The transmitter-receiver uses transistors throughout; valves are used only in the driver and p.a. stages.

Accessories include a high-impact handset, a telegraph key, and an external loudspeaker suitable for dashboard mounting. To permit the use of voice-operated send/receive switching while the loudspeaker is in use, the receiver incorporates a special circuit which prevents automatic switching to the 'send' condition when the loudspeaker output is picked up by the microphone.

EE 03 753 *for further details*

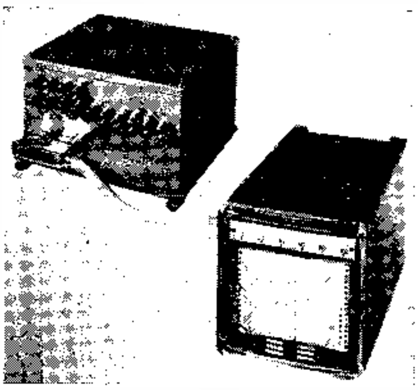
MULTI-POINT RECORDERS

Distributed by: Radiatron, 7 Sheen Park, Richmond, Surrey

(Illustrated below)

Radiatron announces three further additions to the range of multi-point Jacquet miniature potentiometric recorders. All recorders have a paper width of 100mm.

A single line recorder of the new module design is available in single and two part versions; the latter for applications where depth behind the panel is limited to 12in. The recorder has an accuracy of 0·5 per cent of full scale deflexion, and is suitable for measurements from 5mV to 100V and 0·5μA to 100mA as well as for all types of thermocouple and resistance thermometer applications. Paper speed is adjustable from 10 to 3600mm/h.



There is a two-part double line version available which has two completely independent simultaneous measuring systems.

The multi-point recorder, having 2 to 6 channels on a time sharing principle finds many uses in control panel and laboratory applications. Complete interchangeability of channel allocations between voltage current and temperature measurement is provided, and range cards are available for a wide selection of all ranges. A.C. voltage measurements are possible on all models from 20Hz to 20kHz.

EE 03 754 *for further details*

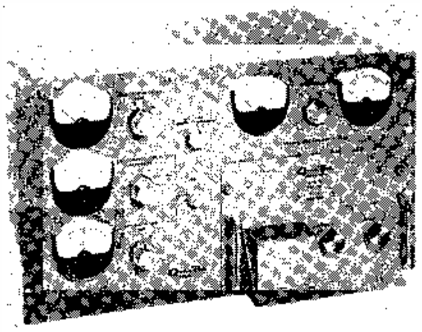
F.E.T. TEST SET

Distributed by: Livingston Laboratories Ltd, Livingston House, Greycaine Road, North Watford, Hertfordshire

(Illustrated below)

A transistor noise test set designed to check the reliability of field effect and bipolar transistors by measuring the noise figure at radio frequencies, is now available from Livingston Laboratories Ltd.

The new Quan-Tech model 340 contains a power supply and measuring circuits plus plug-in r.f. modules designed to increase the instrument's measuring capabilities.



Each module has a transistor test socket, base resistance switch and r.f. circuits. Two standard modules are available at 1MHz and 10·7MHz, although other test frequencies can be supplied.

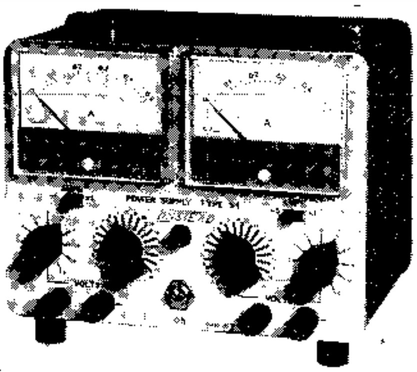
The noise figure is read directly in decibels on a large front panel meter for any of the selected base resistance values—from 50Ω to 5kΩ. Collector current and voltage is variable from 10μA to 30mA and 1V to 60V respectively.

EE 03 755 *for further details*

DUAL POWER SUPPLY

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Illustrated above right)



The dual power supply, type S1, contains two identical stabilized power units giving 0 to 20V, 0 to 0·5A each, which may be connected in series or parallel. The instrument is controlled entirely by silicon devices, giving reliable performance in a small size. Voltage is selected by 20 position switches with continuous control between steps. There is a separate ammeter for each unit which may be switched for 0 to 0·5A or 0 to 100mA f.s.d. Overload protection is incorporated so that the units may be overloaded or short-circuited for an indefinite period without damage or excessive heat rise. On removing the overload the correct voltage is automatically restored.

EE 03 756 *for further details*

DIRECT READING STRAIN BRIDGE

Tecquipment Ltd, Clinton House, 2A Sherwood Rise, Nottingham

(Illustrated on page 122)

This instrument, which was designed at the University of Cambridge, is essentially a Wheatstone bridge in which the two measuring arms consist of two switched analogue transformers in series with a slide wire, by means of which balance is obtained.

The external arms of the bridge consist of any pair of the eleven pairs of strain gauges which can be connected. This number can be extended to 66 pairs by means of a switched extension box which plugs into the bridge unit. All gauges, whether in active use or not, receive power from the 1700Hz oscillator so that there is no waiting to warm up.

A rotary switch selects any gauge channel required, and the balance is indicated on a 5in flat-faced cathode-ray tube. The decade measuring switches are first adjusted to give a horizontal trace on the screen; then, on application of a strain, the trace deflects and is brought back to the horizontal position by manipulation of the measuring switches. The differences between the reading obtained at balance and after re-balancing gives the value of the strain in micro-inches per inch for a gauge factor of two.

The circuit initiating the display consists of a very high adjustable gain balance detecting amplifier, the output of which is used to deflect the trace on the tube in a vertical direction, horizontal deflexion being achieved by direct connexion to the oscillator.

Periodic dynamic strains appear on the tube as two isosceles triangles joined at their apices. If the two sloping sides of the triangles are adjusted by the decade switches to be horizontal in turn, then the difference between the two dial readings is the strain amplitude.

The accuracy of the bridge, which is permanent, is within 3 micro-inches per inch, and is unaffected by mains variations. The range of the bridge is from −10 000 to +22 000 micro-inches per inch; it is robust and almost impossible to damage.

EE 03 757 *for further details*

PROGRAMMING BOARDS

D.E.B. Electronics Ltd, New Barnes Hill, Cottonmill Lane, St. Albans, Hertfordshire

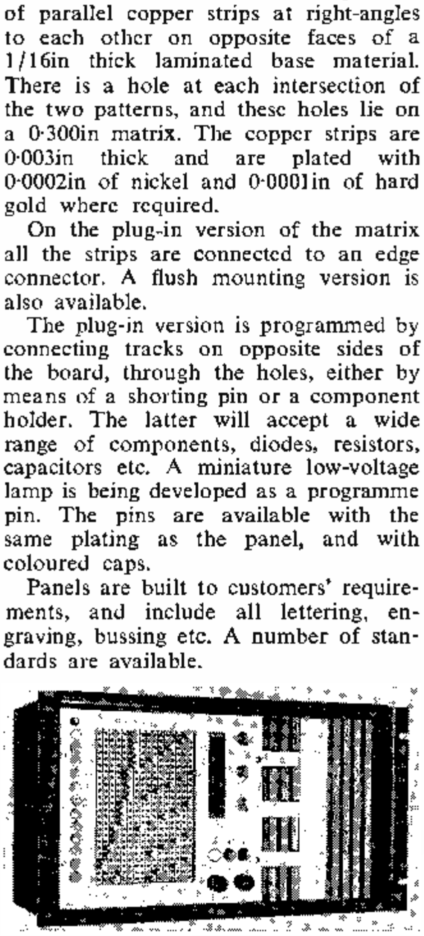
(Illustrated below)

A new and inexpensive programming system has been developed by D.E.B. Electronics. It consists of two patterns of parallel copper strips at right-angles to each other on opposite faces of a 1/16in thick laminated base material. There is a hole at each intersection of the two patterns, and these holes lie on a 0·300in matrix. The copper strips are 0·003in thick and are plated with 0·0002in of nickel and 0·0001in of hard gold where required.

On the plug-in version of the matrix all the strips are connected to an edge connector. A flush mounting version is also available.

The plug-in version is programmed by connecting tracks on opposite sides of the board, through the holes, either by means of a shorting pin or a component holder. The latter will accept a wide range of components, diodes, resistors, capacitors etc. A miniature low-voltage lamp is being developed as a programme pin. The pins are available with the same plating as the panel, and with coloured caps.

Panels are built to customers' requirements, and include all lettering, engraving, bussing etc. A number of standards are available.



EE 03 758 *for further details*

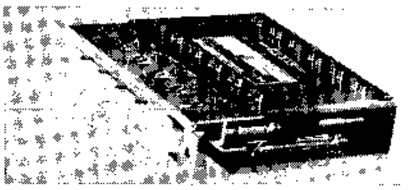
DRY REED PUSH-BUTTON AND KEY SWITCHES

Distributed by: Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustrated below)

The Rotireed type RG key switch is of the standard GPO or Kellog pattern, except that the contacts are totally enclosed miniature dry-reed units mounted horizontally on vertical printed circuit cards. They are actuated by small permanent magnets mounted on an extension of the actuating key. The switch has a central neutral position with up and down make positions, each of which may carry two or four normally-open contacts. Since they close in sequence depending on speed and direction of manual operation, they can be used as late- or early-opening (or closing) contacts. The small front panel (¾ by 2¼in) can be engraved on-off, or 1-0-2, or with any other brief indication. The miniature dry-reed contacts employed switch a maximum contact load of 6W, with a maximum level of 0·3A at 250V, 50Hz, or 26V d.c.

The push-button switch type Rotireed RT, illustrated, is available in banks of 1, 4, 6 and 10 keys as standard. Stations with up to 25 keys are available to special order. Up to four normally-open or normally-closed contacts in any combination may be fitted. Keys may be self-locking when depressed and released by further pressure, self-locking and released by pressure on another key, or



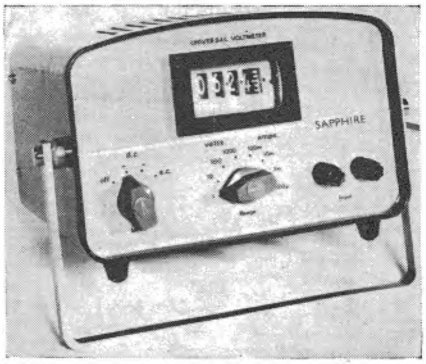
non-locking. Electromagnetic locking or release is also available. Standard key colours are white or black. Keys may have numerals in the opposite colours if desired. Other colours are available to special order. Internally illuminated keys can also be provided. When four sets of contacts per push-button are used, binary-coded outputs are possible. Standard contacts are employed, capable of switching 15W with a maximum level of 1A at 250V. The front panel of the ten-button unit measures 8½ by ⅞in. All RT units are 5¾in deep overall.

EE 03 759 *for further details*

DIGITAL VOLTMETER

Sapphire Research and Electronics Ltd, Sapphire Works, Ferndale, Glam.

(Illustrated above right)



Using novel semiconductor circuits and incorporating an a.c. servo system, the Sapphire Research and Electronics model 250, reads d.c. voltages from 1000V and a.c. voltages from 500V r.m.s., and in addition it has d.c. and a.c. current ranges.

It can resolve voltage differences as small as 200μV. The a.c. frequency range is 10Hz to 60kHz. The standard of accuracy is 0·1 per cent f.s. on the d.c. voltage ranges; and 0·25 per cent on the a.c. voltage ranges. The input impedance of the instrument is high, being of the order 1000MΩ on the 1V range, and 2·2MΩ on the other voltage ranges. The new instrument has a very stable set zero calibration performance, and requires little or no operator attention. The read-out is both clear and bright under all conditions of ambient lighting, while the decimal point position is automatic on all ranges. Over-voltage protection is also provided on all ranges.

The instrument is intended for laboratory and industrial use, and provides high performance with excellent quality, portability and very low cost.

EE 03 760 *for further details*

MULTI-CHANNEL RECORDER

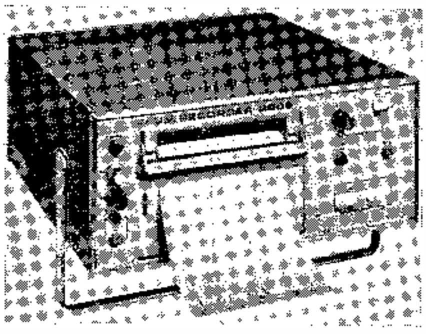
S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Illustrated below)

S.E. Laboratories (Engineering) Ltd has introduced a low cost, low weight oscillograph SE.3006. Strict attention to value, engineering and simple all modular plug-in construction has made possible a trace recorder which combines the advantages and flexibility normally associated with very much higher priced instruments.

The oscillograph is supplied either as a 6 or 12 channel instrument, using standard English galvanometers, writing on 6in wide sensitized paper. It is a portable unit only 7in high, weighing 42lb, and is suitable for table or rack mounting.

An important feature is its low power consumption of 250VA while still employing the mercury vapour ultraviolet lamp, with its high writing speed capa-



bilities. A 1500 : 1 range of paper speeds, chosen from 5mm/min to 1250mm/sec, with an 8 step push-button selection, is instantly selectable while running. Full width timing lines at 0·01, 0·1 and 1sec and 0·1 and 1min intervals as well as 2mm grid lines are printed on the record. Individual trace and grid intensity controls are provided as well as remote control facilities.

Paper loading is easily effected by simply dropping a roll of paper into the cassette, threading being automatic.

EE 03 761 *for further details*

TRUE R.M.S. INSTRUMENTS

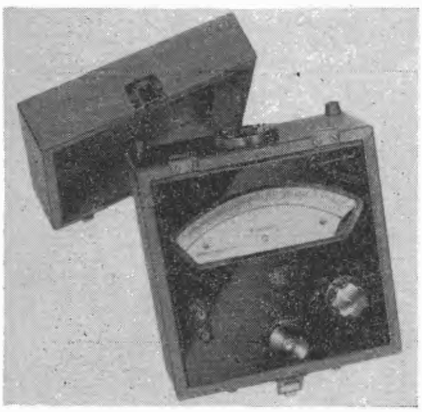
Distributed by: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustrated below)

Greibach 'Transquare' true r.m.s. instruments are designed for high accuracy measurements on complex waveforms and are particularly suitable for work with silicon controlled rectifiers (thyristors), such as the precision measurement of the power input to s.c.r.-controlled heat rigs. Conventional electronic true r.m.s. instruments are capable of accuracy only in the 2 to 3 per cent region, as opposed to the 0·5 per cent accuracy of the 'Transquare' meter.

The true r.m.s. voltmeter range extends from 30mV to 1000V, full scale. Current measuring capability extends up to 200A, self-contained and 1000A with external current transformer or shunt (for a.c./d.c. measurements). Maximum sensitivity has hitherto been 0·1mA, full scale, but Greibach has now developed a true r.m.s. microammeter with a low full scale range of 50μA.

Typical of these new sensitive models the type 500T005-8 has eight ranges from 0·05mA to 100mA. The instrument will achieve an accuracy of 0·5 per cent of full scale for irregular waveforms with crest factors in excess of 4—also square waves up to several kHz. Exceptional stability with repeatability of better than 0·1 per cent is obtained since no amplifiers, power supplies, or other active components are used. It is capable of withstanding 100 times over-current at the more sensitive current ranges and will respond to input frequencies of d.c. and 5Hz to 500kHz with an accuracy of $\pm$0·5 per cent of full scale at the most sensitive ranges.



EE 03 762 *for further details*

D.C. BRIDGE AND AMPLIFIER

Vibro-Meter SA., Haletop Civic Centre, Wythenshawe, Manchester, 22

(Illustrated below)

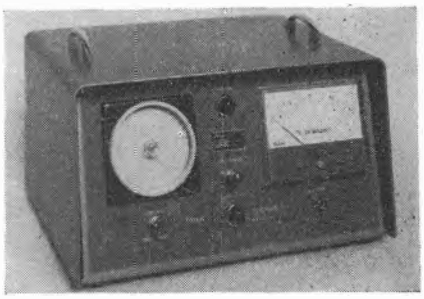
Vibro-Meter SA., the Swiss instrument company, has recently introduced a fully transistorized d.c. power supply complete with balancing and calibration facilities. The unit SPB-1/A, is designed for use with strain gauges of 60 to 600Ω resistance and load cells, accelerometers and pressure transducers of the strain gauge type. A matching wideband voltage amplifier (DCA-1/A) is also available for signal amplification if required.

The SPB-1/A power supply utilizes a fully floating bridge circuit suitable for either full or half bridge systems and has a minimum insulation to earth of 10^9. Measurement is made by the deflexion or compensation method and adjustment of the bridge balance is claimed to be exactly reproducible.

The instrument includes a balancing network which permits compensation of zero unbalance in the transducer and compensation for tare weight up to 300 per cent of the nominal value. Final measurement is achieved by the compensation method, a built-in calibration device is provided and positive or negative signals proportional to microstrain can be selected.

The associated amplifier, Vibro-Meter type DCA-1/A has a zero drift figure of only 3μV°C. The instrument, which is fully transistorized, is basically a direct coupled high stability d.c. amplifier with differential input and variable gain from 3 to 1000. Facilities for zero balance and common mode rejection are provided. Outputs of $\pm$10V, $\pm$2mA, and $\pm$10V, $\pm$100mA, are available.

EE 03 763 *for further details*



AUTOMATIC DEWPOINT RECORDER

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Illustrated above)

The automatic recording of the dewpoint of air or gases is now possible by means of the new Shaw Automatic dewpoint recorder. This new instrument is completely solid state and the recorder gives daily, weekly or monthly charts. Absolute readings of dewpoint are given by means of a mirror, which is cooled by means of the Peltier effect, the formation of dew is sensed by a photocell, and the mirror temperature is shown on the 5in meter dial and also recorded.

The automatic dewpoint recorder has positive indication that the true dewpoint is being indicated and this takes the form of a lamp which lights once a second, indicating that dew and no dew conditions exist on the mirror.

Standard ranges are -60°C to -20°C and -20°C to $+20$°C dewpoint, although any range is available to special order. The recorder has no controls to adjust and no external services, other than a mains supply, are needed.

A model for meteorological purposes or for use in ambient conditions has also been developed.

EE 03 764 *for further details*

DIGITAL COUNTER

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated on page 124)

This new electronic counter has the wide frequency range and front panel accessory plug-in capability of more expensive counters but because certain refinements are omitted, its cost is considerably lower. The basic counter has a maximum counting rate of 50MHz and it can measure frequencies accurately and with 1Hz resolution to as high as 12·4GHz with available HP plug-in heterodyne frequency convertors. Although it is an economy model, the new counter has a full complement of gate times from 1μsec to 1sec.

The new counter (HP Model 5246L) accepts all plug-ins designed for the HP model 5245L 50MHz counter. Besides measurements to 12·4GHz using HP convertor plug-ins, the new economy model can measure frequencies directly up to 350MHz with a prescaler plug-in, operate with a sensitivity of 1mV with a pre-amplifier plug-in (basic counter sensitivity is 100mV), normalize readings with a preset counter plug-in, or

measure time intervals from 1μsec to 10^8sec and make digital voltage measurements to 1000V with two other plug-ins.

The new counter has a six-digit readout. Optionally, 7^{th} and 8^{th} digits may be added for higher resolution at the time of purchase or at any later time. Basic accuracy of the instrument is determined by the time-base oscillator which has an ageing rate of less than ± 2 parts in 10^7 per month. An optional high stability time-base is available with an ageing rate of less than ± 3 parts in 10^9 per 24 hours and with an excellent short term stability (less than ± 2 parts in 10^{10} with a 1sec averaging time). The new counter may also be operated from an external frequency standard, and can measure frequency ratios.

Input impedance of the dual f.e.t. amplifier input is 1MΩ shunted by 25pF for all signal levels up to 1V independent of frequency to 50MHz. With the constant high input impedance, a low v.s.w.r. is assured when the input is connected to a terminated line and it is possible to use an oscilloscope type probe when voltage attenuation or a 10MΩ 10pF input is desired.

Other optional features available with the new counter include b.c.d. output in 1-2-2-4 positive '1' code or 1-2-4-8 code with positive or negative '1' state.

EE 03 765 *for further details*

ULTRASONIC HARDNESS TESTER

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated right)

A new instrument for the precision micro-hardness testing of materials, called the Sonodur micro-hardness tester type 1851, is now made by Dawe Instruments Ltd. The unit has a range from 20 to 70 Rockwell C, with an accuracy of ±1 Rockwell C.

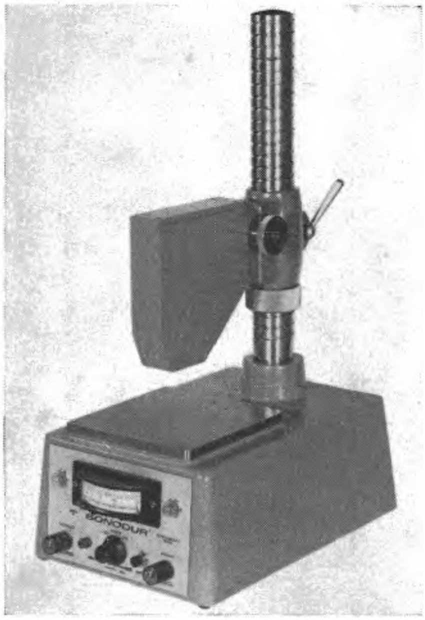
In operation, the component to be tested is placed on the table of the base. The column-mounted indenter probe, tipped by a Vickers diamond is allowed to descend on the component under gravity. The static load applied is only 150g. The probe is excited at its natural resonant frequency by an ultrasonic generator housed in the base. Vibrational frequency of the probe changes in proportion to the depth of penetration of its tip into the component. Change in frequency is measured electronically and displayed on a meter, also housed in the base. The meter scale is calibrated directly in degrees hardness.

This new model differs from the type 1850A Sonodur introduced last year in that the probe is coloumn-mounted instead of being hand-held. Greater uniformity of loading ensures greater accuracy in read-out. The type 1850 continues to be available, however, partly for testing in the field and partly for work on workpieces too large to be accommodated by the type 1851.

Since readings are taken while the probe is in contact with the work, errors due to elastic recovery are eliminated. Each reading takes than 10 seconds and requires no skill or training. Depth of penetration in steel is less than 0·0002in, even on fully annealed material, and the area is less than 0·1 per cent of the area of endentation made by a Rockwell C tester.

The type 1851 Sonodur eliminates the necessity for optical and mechanical measuring devices. Because of the electrical read-out, any portion of the scale may easily be expanded to give increased ease of reading when used for production testing, thus reducing the possibility of reading errors.

The instrument is supplied as stan-



dard graduated for measuring the hardness of steel (Young's modulus 29-30 million), but it can be set up for testing all other metals by simply calibrating it against two samples of known hardness. An additional scale graduated in arbitrary units marked from 0 to 100, is supplied for comparison-testing on other materials.

EE 03 766 *for further details*

GUNN EFFECT OSCILLATOR

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

The first commercially available X-band (7 to 12GHz) Gunn effect oscillator is announced by Mullard Ltd.

Developed at the Wembley Laboratories of Associated Semiconductor Manufacturers Ltd and the Mullard Research Laboratories, the device gives a continuous output of 5mW and is powered from a simple 6V d.c. supply. It measures only 5mm in length × 3mm in diameter.

Because of its extremely small size and simplicity of operation, it is an attractive alternative to the conventional klystron valve or transistor/varactor type of local oscillator used in the mixer stages of microwave communications links and radar systems. It can also be used as the direct source of microwave energy in low-power radar systems (for example, speed checking equipment) and bench-top microwave demonstration systems for schools and technical colleges.

In use the device, development type 106CXY, is coupled into a simple microwave cavity, the resonant frequency of which determines the output frequency. The cavity may be fixed frequency or made to tune over the 7 to 12GHz range.

Previous Gunn effect oscillators have been produced by creating a suitable field between the two faces of a thin slice of gallium arsenide (GaAs)—the thinner the slice, the higher the operating frequency. Until now the slices have been made by a grinding and lapping technique which limited the maximum operating frequency to about 4GHz.

To overcome this problem Mullard has developed a technique of epitaxial deposition which allows a very thin layer of GaAs to be formed on a thicker slice of the same material. A device made in this way has its active element at the surface in a high-resistivity layer on a low-resistivity substrate, the action taking place through the layer.

In addition to increasing the operating frequency of the device, this technique offers greater control of resistance for a given resistivity layer and higher power dissipation.

EE 03 767 *for further details*

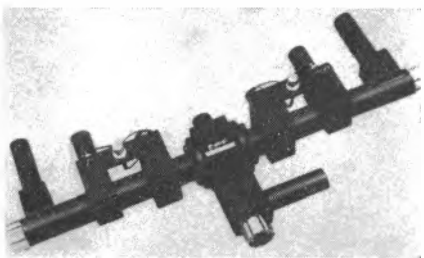
MICROWAVE SWITCH

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

(Illustrated below)

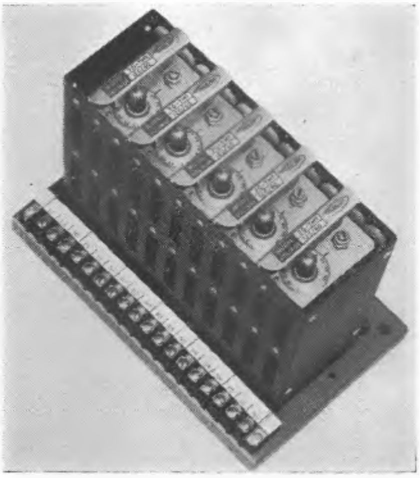
Microwave Associates announce the development of a new coaxial s.p.d.t. diode switch with a 60dB isolation, a v.s.w.r. of under 1·07 and an insertion loss of less than 0·5dB over a 20 per cent bandwidth in the 1 to 2GHz frequency range.

For bandwidths of up to 50 per cent this switch, the MA8302-2L4S, offers more than 40dB isolation, v.s.w.r. less than 1·20 and an insertion loss of under 0·75dB.



Designed primarily for microwave automatic test equipment in conjunction with phase bridges, the switch is also ideally suited to 'hot-switching' applications where two transmitters are switched into a common output, or for alternate switching between two antennas.

EE 03 768 *for further details*



CURRENT MONITORS

Scama Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(Illustrated above)

A new current monitor for control or alarm applications is announced by Scama Ltd, manufacturers of 'Panalarm' annunciators. The series 70 'Trip-Matic' monitor can be operated either as a separate unit for control cubicle or rear panel mounting, or an integral part of a model 75 Panalarm annunciator to provide direct annunciation from analogue inputs.

In a typical application the 'Trip-Matic' monitors a d.c. signal which is proportional to the level of the operating parameter. As the value of the parameter rises so does the d.c. signal. When it just exceeds a preset adjustable reference a comparator-amplifier has a sharp change in output which positively signals the annunciator or other controlled device.

Exceptional accuracy is claimed for the new signal monitor, with a very quick response. Designed to operate over a broad range of d.c. signals it can also be adapted to handle many a.c. applications. It is completely transistorized with silicon semiconductors throughout. All inputs are isolated from each other and from outputs. The differential between trip and reset can be fixed or adjustable, with either a high or low trip operation to suit conditions of use.

EE 03 769 *for further details*

DIFFERENTIAL VOLTMETER

Distributed by: General Test Instruments Ltd, Gloucester Trading Estate, Hucclecote, Gloucester

(Illustrated above right)

The new Dialamatic voltmeter manufactured by Wavetek is a manually operated unit that approaches the speed of automatic voltmeters but offers the accuracy, rugged reliability, and price of conventional differential models.

The key to this performance is the Transfermatic switch. This device mechanically couples decade switches so the 'carries' are transferred to the next higher decade. Upon approaching a major transfer point, such as 2·999, one click of the right-hand knob automatically moves the read-out to 3·000. The procedure also works in reverse.

To get the same result with a conventional differential voltmeter necessitates resetting each knob individually.

The Dialamatic also features a fast non-saturating null amplifier which allows the operator to dial the correct balancing without changing null sensitivity.

In addition to speed of operation, the Dialamatic has been designed for quick and easy maintenance. Each circuit is mounted on a modular plug-in card which pivots out of the case should maintenance be necessary. All silicon semiconductors have been used to ensure reliability under extreme environmental conditions.

The Dialamatic is available in two basic models. The model 201 measures d.c. voltages from 0 to 1 000V in four ranges, and the model 202 features both a.c. and d.c. with measurement ranging from 1mV to 1 000V. Each model is available in battery-powered versions with built-in recharger.

EE 03 770 *for further details*

PRECISION VARIABLE RESISTORS

Steatite Insulations Ltd, Hagley House, Hagley Road, Birmingham 16

This variable precision resistor, known as the VariZistor combines the function of a precision resistor with the adjustability of a trimmer potentiometer in a package that provides infinite resolution. Extremely accurate resistance levels are said to be easily achieved, thus tailoring the resistance to suit the specific requirements of the circuit.

The device consists of a deposited metal thin-film resistance element, bonded to a ceramic substrate and inserted into a rectangular-shaped housing. A multi-contact wiper is attached to the lead and insulated on one side. The lead runs through the body, allowing the clip to make contact with the resistance element in order to provide circuit continuity. The entire body slides within fixed limits, moving the wiper across the resistance element. An approximate 20 per cent variation in resistance above and below the nominal value is available from 0·5Ω to 1MΩ adjustable to 0·01 per cent.

To the circuit designer the device provides an opportunity to experiment with circuit components without the usual problems inherent with fixed resistors. In sensitive circuit applications, minute resistance changes can be made at the final inspection, after soldering the leads, to compensate for component tolerances and stray resistance.

VariZistors are designed to meet or exceed all requirements of applicable high reliability specifications. Leads may be either welded or soldered in any circuit. Accurate resistance values may be fixed by soldering the leads at the body junction. Endcaps may be soldered at the body junction to ensure a hermetic enclosure. Models are available $\frac{1}{4}$, $\frac{1}{2}$ and 1W ratings, and special temperature coefficients may be ordered as standard to as low as $5 \times 10^{-6}/°C$.

EE 03 771 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.
Date: 1 February.
Lecture: Sub-millimetre Wave Techniques.
By: H. A. Gebbie.
Date: 2 February.
Discussion: The Assessment of the Economics of Automation Projects.
Opened by: D. W. Gillings and M. F. Binney.
(Joint meeting with the Automatic Control Group of the I.Mech.E. held at 1 Birdcage Walk, London, S.W.1, at 6 p.m.)
Date: 6 February.
Discussion: The Place of Mathematics in the Training of Electrical Engineers at University.
Opened by: G. S. Brayshaw.
Date: 7 February.
Lecture: The Automated Launching Sequence System for the First Stage of the ELDO 1 Satellite Launching Vehicle.
By: T. Cast and P. Waples.
(Joint meeting with the Automatic Control Group of the I.Mech.E.)
Date: 8 February.
Lecture: Audio Reproduction in the Home. Fifth Annual Lecture of the Electronics Division.
By: G. F. Dutton.
Time: 6 p.m.
Date: 9 February.
Discussion: The Use of Transducers in Vibration Measurement.
Opened by: P. J. Hurst.
(Joint meeting with the Automatic Control Group of the I.Mech.E.)
Date: 13 February.
Discussion: The Impact of Integrated Circuits on the Role of the Circuit Designer.
Opened by: A. G. I. Cressell and R. Towell.
Date: 14 February.
Lecture: Transition Characteristics of Jump Phenomena in Non-Linear Resonant Circuits.
By: J. C. West, B. V. Jayawant and D. P. Rea.
Date: 20 February.
Colloquium: Precision Resistance Standards.
Time: 9.30 a.m.
Date: 20 February.
Colloquium: Message Switching.
Time: 9.30 a.m.
Date: 20-21 February.
Symposium: Area Control of Road Traffic.
(Joint meeting with the I.C.E. and I.Mech.E. at the Institution of Civil Engineers, 1-7 Great George Street, London, S.W.1.)
Date: 23 February.
Discussion: The Manley-Rowe Relations.
Opened by: J. Brown.
Date: 27 February.
Lecture: Storage Systems for Telephone Switching.
By: J. R. Pollard.

London Graduate and Student Section

Date: 6 February.
Lecture: Some Problems in the Design and Use of Power Cables.
By: B. J. Hardy.

THE INSTITUTION OF ELECTRICAL AND RADIO ENGINEERS

Radar Group

Held at: Institution of Electronic and Radio Engineers, 8-9 Bedofrd Square, London, W.C.1.
Date: 15 February.
Time: 6 p.m.
Lecture: The Remote Control of Lighthouses and Beacons.
By: A. C. MacKellar and M. J. Dilworth.

Television Group

Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: A Dual Standard Colour Television Receiver.
By: P. L. Mothersole, D. S. Hobbs and D. J. King.

Southern Section

Held at: Basingstoke Technical College.
Date: 9 February.
Time: 7.30 p.m.
Lecture: Introduction to Digital Computers.
By: E. G. Anderson.
Held at: Bournemouth College of Technology.
Date: 22 February.
Time: 7 p.m.
Lecture: Development of Satellite Communications.
By: J. K. S. Jowett.
Held at: Brighton College of Technology.
Date: 14 February.
Time: 6.30 p.m.
Lecture: Flight Simulation.
By: R. A. Marvin.
Held at: Technical College, Newport, Isle of Wight.
Date: 3 February.
Time: 7 p.m.
Lecture: M.O.S. Transistors.
By: G. G. Bloodworth.

South Western Section

Held at: University of Bristol, Small Lecture Theatre, University Walk, Clifton, Bristol.
Date: 6 February.
Time: 7 p.m.
Lecture: Colour Television.
By: H. V. Sims.
(Joint meeting with Western Centre of the I.E.E. (Electronics Section).)
Held at: The College, Swindon.
Date: 15 February.
Time: 6.15 p.m.
Lecture: Television Recording.
By: P. Leggat.
(Joint meeting with the Western Centre of the I.E.E. (Electronics Section).)
Held at: South Devon Technical College, Torquay.
Date: 21 February.
Time: 7 p.m.
Lecture: Automatic Landing Systems.
By: R. A. Bailey and J. Meadows.
(Joint meeting with the Western Centre of the I.E.E. (Electronics Section).)

South Wales Section

Held at: Welsh College of Advanced Technology, Cardiff.
Date: 8 February.
Time: 6.30 p.m.
Lecture: Transmission Measuring Equipment for Telecommunication Systems.
By: H. M. Evans.

East Anglian Section

Held at: University Engineering Department, Trumpington Street, Cambridge.
Date: 2 February.
Time: 8 p.m.
Lecture: A Simplified Approach to the Use of Transistors in Video, Pulse and I.F. Circuits.
By: E. Davies.
Held at: Chelmsford Technical High School.
Date: 16 February.
Time: 6.30 p.m.
Lecture: High Frequency Guided Waves in Application to Railway Signalling and Control.
By H. M. Barlow.
(Joint meeting with the I.E.E. Chelmsford District.)
Held at: Southend College of Technology.
Date: 28 February.
Time: 7 p.m.
Lecture: Adaptive Astable/Monostable Circuits in Class D Amplifiers.
By: D. C. Smith.

Scottish Section

Held at: University of Strathclyde, Glasgow, C.1.
Date: 13 February.
Time: 6 p.m.
Lecture: Computer Aided Study of Character Recognition.
By: J. A. Weaver.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Date: 14 February.
Time: 6 p.m.
Lecture: Computer Aided Study of Character Recognition.
By: J. A. Weaver.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Date: 23 February.
Time: 6 p.m.
Lecture: Some Modern Advances in Scintillation Scanners and Cameras.
By: W. G. Walker.
(Joint I.E.E./I.E.R.E. Medical Electronics Meeting.)

Yorkshire Section

Held at: Bradford Institute of Technology.
Date: 9 February.
Time: 7 p.m.
Lecture: Digital Logic.
By: F. Houldsworth.

Merseyside Section

Held at: Liverpool College of Technology.
Date: 22 February.
Time: 7 p.m.
Lecture: G.P.O. Towers.
By: S. G. Young.

East Midland Section

Held at: Loughborough College of Technology.
Date: 14 February.
Time: 6.30 p.m.
Lecture: Stereophonic Sound Reproduction.
By: J. Moir.
(Joint meeting with the I.E.E. (East Midland Electronics Control Section).)

South Midland Section

Held at: Abbey Ballroom, Malvern.
Date: 14 February.
Time: 7 p.m.
Annual General Meeting followed by:
Lecture: Speech Synthesis by R.R.E.A.C. Digital Computer.
By: P. M. Woodward.

West Midland Section

Held at: Stafford College of Further Education, Tenterbanks, Stafford.
Date: 14 February.
Time: 7.15 p.m.
Lecture: Railway Control and Signalling.
By: J. H. Fews.

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Date: 8 February.
Time: 6 p.m.
Lecture: The Stereoscan Electron Microscope.
By: I. H. Gordon.

Thames Valley Section

Held at: Lecture Theatre, Slough College, Slough.
Date: 21 February.
Time: 7.30 p.m.
Lecture: Pulse Code Modulation.
By: J. R. Jarvis.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Formal Meeting

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 14 February.
Time: 5 p.m.
Lecture: The Introduction of Crossbar Exchanges into the P.O. Network.
By: J. Spinks, W. F. J. Hall and P. H. Skinner.

Informal Meeting

Held at: The Conference Room, Fleet Building, 70 Farringdon Street, London, E.C.4.
Date: 28 February.
Time: 5 p.m.
Lecture: Telecommunications in the Midland Region of British Railways.
By: C. B. Drury and J. A. Warburton.

THE ROYAL TELEVISION SOCIETY

Held at: The I.T.A., 70 Brompton Road, London, S.W.3.
Time: 7 p.m.
Date: 7 February.
Lecture and Demonstration: Slow Motion Videodisc Recording.
By: K. Machein.
Date: 10 February.
Lecture: P.A.L. Studio Operation—A First Look at the Problems.
By: G. B. Townsend.
Date: 28 February.
40th Anniversary Banquet at the Goldsmiths Hall, Foster Lane, London, E.C.2.

NOUVELLES Réalisations

Traduction des pages 120 à 125

Une description basée sur des renseignements fournis par les fabricants de nouveaux composants, accessoires et instruments d'essai

ÉQUIPEMENT DE CONTRÔLE DE BANDE

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Illustration à la page 120)

La société Gresham Lion Electronics Ltd a réalisé un nouveau type de contrôleur de bande magnétique pouvant contrôler les moindres défauts sur la largeur totale.

La société Gresham offre un choix étendu de dispositifs de repérage et la version convenant le mieux pour les travaux de calculs constitue une tête d'enregistrement de largeur totale pour une bande de 1,25 cm suivie d'un dispositif de réenregistrement double avec 7 et 9 pistes (type P.S. 79). Le guidage de la bande sur les bords est assuré par des "guides" en céramique à ressorts et les blocs de tête sont protégés par une housse.

Les spécifications électriques et mécaniques peuvent être adaptées aux applications particulières et la spécification de l'assemblage de tête à réenregistrement à 7 et 9 pistes pour bande de 1,25 cm est indiquée ci-après.

Tête d'enregistrement

Largeur de piste	0,505 ÷ 0,001 pouce
Longueur d'entrefer	0,0005 pouce (nominal)
Résistance	25Ω
Inductance	7mH
Courant de saturation	33mA

Tête de réenregistrement	*7 pistes*	*9 pistes*
Largeur de piste	0,76 mm ±0,0025 mm	1,02 mm ±0,0025 mm
Largeur d'entrefer	6,35 μm	6,35 μm
Résistance	11,4 Ω	12,5 Ω
Inductance	5,6 mH	5,3 mH

EE 03 751 *pour plus amples renseignements*

ÉLÉMENT ASSERVI DE POTENTIOMÈTRE

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustration à la page 120)

L'élément asservi de potentiomètre type TA61 constitue un moyen simple et à prix modéré d'amorcer des charges de courant à partir de la sortie de transducteurs de type potentiométrique.

Grâce à cet élément il est maintenant possible d'amorcer de grands instruments de mesure, des enregistreurs à plume ou des enregistreurs galvanométriques à partir de transducteurs de déplacement et de pression basés sur des éléments sensible potentiométriques. L'erreur de non-linéarité normalement introduite lorsqu'une change de courant est amorcée directement à partir de la sortie de transducteurs potentiométriques a été éliminée en basant le TA61 sur un dessin de circuit unique.

Bien que principalement prévu pour le fonctionnement en liaison avec les transducteurs intersonde, l'élément asservi de potentiomètre peut être utilisé avec n'importe quel transducteur potentiométrique ayant une resistance terminale entre 500 Ω et 10 Ωk. Cet élément est particulièrement utile surtout dans les applications industrielles, en raison du fait qu'il peut être logé à plusieurs centaines de mètres du transducteur. La charge reliée à l'élément peut avoir une résistance de 0 à 1000 Ω et elle peut être également logée à distance si nécessaire.

Le TA61 est normalement utilisé avec des transducteurs mesurant des paramètres statiques ou à variation lente mais il est également utilisable pour les applications dynamiques. La réponse de fréquence des éléments est linéaire de 0 à 10 kHz et une précision de ±0,2% de la sortie sur la totalité de la gamme peut être obtenue dans la gamme de température de 0 à 60°C.

EE 03 752 *pour plus amples renseignements*

RÉCEPTEUR-ÉMETTEUR À BANDE LATÉRALE UNIQUE

Redifon Ltd, Broomhill Road, London, S.W.18

(Illustration à la page 120)

La Division des Communications de la société Redifon Ltd a mis au point un nouveau récepteur émetteur de 100 W à haute fréquence et à bande latérale unique, type GR.410T, pour répondre à la demande des forces militaires et de sécurité d'un appareil d'une souplesse de fréquence maxima à un prix économique. Ce récepteur émetteur constitue une étape intermédiaire entre le matériel actuel à modulation d'amplitude et à double bande latérale utilisant une commande de fréquence par oscillateur à fréquence variable et le matériel futur à bande latérale unique avec synthèse de fréquence totale et il est particulièrement indiqué pour les organisations qui envisagent la mise en phase de la bande latérale unique.

Le G.R. 410T allie la simplicité et la stabilité du contrôle piézoélectrique à la souplesse des commandes du panneau frontal pour le changement rapide des cristaux et le nouvel accord aux nouvelles fréquences. Des commandes manuelles sont prévues pour accorder la sortie de l'oscillateur, des étages HF et de la charge de l'amplificateur de puissance, cependant qu'un instrument de mesure incorporé contrôle le niveau d'entraînement, le courant anodique d'amplification de puissance, le courant d'antenne avant et les tensions d'alimentation. Les fréquences peuvent être changées en quelques minutes par des opérateurs sur place sans personnel d'entretien ou ajustage d'atelier.

L'émetteur récepteur peut être muni à n'importe quel moment d'un maximum de quatre cristaux de commutation et un opérateur peut porter des cristaux supplémentaires avec lui pour n'importe quelle autre fréquence pouvant être requise, ce qui permet un redéploiement rapide dans des régions où d'autres fréquences de fonctionnement sont utilisées. D'autre part, l'émetteur récepteur peut recevoir une entrée d'une source externe tel qu'un synthétiseur de fréquence.

Le GR.410T couvre la gamme de fréquence de 2 à 16 MHz avec sélection par commutateur de bande latérale universel ou de bande latérale longue et mode à bande latérale unique, une porteuse commutée donnant une modulation d'amplitude compatible (A3H et A2H) pour l'emploi dans des réseaux de communication qui ne sont pas encore équipés pour la bande latérale unique. Il peut également être utilisé sur ondes progressives. La commutation automatique de transmission/émission peut être effectuée au moyen d'un commutateur sur le panneau frontal, et la commande de niveau automatique empêche la surmodulation tout en assurant une modulation adéquate au niveau de parole réduit.

Un appareil de réglage Redifon à accord manuel est prévu pour ajuster l'émetteur transmetteur à une antenne montée sur véhicule ou une simple antenne métallique. Le GR.410T peut être utilisé également dans un dipôle de 75Ω ou dans une antenne de 600Ω à travers un transformateur d'équilibrage à large bande.

Des blocs d'alimentation sont prévus pour 100 à 125, 200 à 250 V 50 à 60 Hz secteur ou pour le fonctionnement sur batterie de 12 ou 24 V permettant d'utiliser le GR.410T dans des rôles mobiles, fixes ou transportables.

Le générateur-démodulateur à bande latérale unique, les éléments acoustiques et à basse fréquence sont construits sous forme de module à fiche et ils sont fixés par des agrafes à détente rapide pour permettre leur enlèvement rapide pour l'entretien. L'émetteur récepteur utilise exclusivement des transistors; les tubes ne sont utilisés que dans les étages d'entraînement et d'amplification de puissance.

Les accessoires comprennent un combiné à impact élevé, une clé télégraphique et un haut-parleur externe pouvent être monté sur tableau de bord. Pour permettre l'emploi de la commutation d'émission/réception entraînée par la voix tandis que le haut parleur est en usage, le récepteur comporte un circuit spécial qui empêche la commutation automatique à l'émission tandis que la sortie du haut-parleur est captée par le microphone.

EE 03 753 *pour plus amples renseignements*

ENREGISTREUR MULTIPLE À PLUSIEURS TRACES

Distributeurs: Radiatron, 7 Sheen Park, Richmond, Surrey

(Illustration à la page 121)

La société Radiatron vient d'ajouter trois nouveaux modèles à sa gamme d'enregistreurs potentiométriques miniature Jacquet multiples à plusieurs traces. Tous ces enregistreurs ont une largeur de papier de 100 mm.

Un enregistreur à ligne unique du nouveau dessin à module peut être obtenu en version simple et à deux parties; cette dernière est pour les applications où la profondeur derrière le panneau est limitée à 30 cm. Cet enregistreur a une précision de 0,5 % pour une déviation sur la totalité de l'échelle et il convient pour les mesures de 5 mV à 100 V et de 0,5 μA à 100 mA ainsi que pour toutes les applications de thermocouples et de thermomètre à résistance. La vitesse d'avance du papier est réglable de 10 à 3600 mm/h.

Il existe également un modèle à double ligne en deux parties comportant deux systèmes de mesure simultanée entièrement indépendants.

L'enregistreur à plusieurs traces, à 2 à 6 voies utilisées suivant le principe de la division du temps, trouve de nombreux usages dans les applications de panneaux de contrôle et de laboratoire. L'interchangeabilité complète des canaux pour la mesure de tension, du courant et de la température peut être effectuée et des cartes de gamme sont prévues pour un choix étendu de gammes. La mesure de tension alternative peut être effectuée sur tous les modèles de 20 Hz à 20 kHz.

EE 03 754 *pour plus amples renseignements*

CONTRÔLEUR D'EFFET DE CHAMP

Distributeurs: Livingston Laboratories Ltd. Livingston House, Greycaine Road, North Watford, Hertfordshire

(Illustration à la page 121)

Un contrôleur de bruit à transistors, conçu pour contrôleur l'effet de champ et les transistors bipolaires en mesurant l'indice de bruit aux hyperfréquences, peut être obtenu maintenant de la société Livingston Laboratories Ltd.

Le nouveau contrôleur Quan-Tech, modèle 340, contient un bloc d'alimentation et des circuits de mesure ainsi que des modules HF à fiches prévus pour augmenter le pouvoir de mesure de l'instrument.

Chacun des modules comporte une douille de contrôle de transistors, un commutateur de résistance de base et des circuits HF. Deux modules standard sont prévus à 1 MHz et 10,7 MHz, bien que d'autres fréquences d'essai puissent être fournies.

L'indice de bruit se lit directement en décibels sur un grand instrument de mesure à panneau frontal pour n'importe laquelle des valeurs de résistance de base choises, de 50Ω à 5 kΩ. Le courant au collecteur et la tension varient de 10 μA à 30 mA et de 1 V à 60 V respectivement.

EE 03 755 *pour plus amples renseignements*

BLOC D'ALIMENTATION DOUBLE

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Illustration à la page 121)

Le bloc d'alimentation double, type S1, contient deux éléments identiques de puissance stabilisée donnant 0 à 20 V et 0 à 0,5 A chacun. Ces éléments peuvent être reliés en série ou en parallèle. L'instrument est entièrement commandé par dispositifs au silicium, combinant un fonctionnement sûr avec des dimensions réduites. La tension voulue est obtenue par 20 commutateurs de direction avec commande continue entre plots. L'appareil comporte un ampèremètre à part pour chacun des éléments et pouvant être commuté de 0 à 0,5 A ou de 0 à 100 mA de déviation sur la totalité de l'échelle. Un disjoncteur à maxima est incorporé à l'appareil de sorte que les éléments peuvent être surchargés ou court-circuités pendant une période de temps indéfinie sans risque ou échauffement excessifs. Lorsque la surcharge est supprimée, la tension voulue est automatiquement rétablie.

EE 03 756 *pour plus amples renseignements*

EXTENSOMÈTRE À LECTURE DIRECTE

Tecquipment Ltd. Clinton House, 2A Sherwood Rise, Nottingham

(Illustration à la page 122)

Cet instrument, conçu à l'université de Cambridge, est essentiellement un pont de Wheatstone dans lequel les deux bras de mesure consistent en deux transformateurs analogiques commutés en série avec un curseur au moyen duquel l'équilibre est réalisé.

Les bras extérieurs du pont consistent en une paire des onze paires d'extensomètres pouvant être reliés. Ce chiffre peut être porté à 66 paires au moyen d'une boîte d'extension commutée qui peut être branchée dans l'élément à pont. Tous les extensomètres, qu'ils soient actifs ou non, reçoivent une puissance de l'oscillateur de 1700 Hz, de sorte qu'il n'y a aucune période d'attente de chauffe.

Un commutateur rotatif choisit le canal de mesure voulu et l'équilibre est indiqué sur un tube cathodique à face plate de 12,5 cm. Les commutateurs de mesure à décade sont d'abord réglés pour produire une trace horizontale sur l'écran; puis en appliquant une contrainte, la trace dévie et elle est ramenée à la position horizontale par manipulation des commutateurs de mesure. Les différences entre l'indication obtenue à l'équilibre et après ré-équilibrage donnent la valeur de la contrainte en micropouce par pouce pour un facteur de mesure de deux.

Le circuit qui déclenche l'affichage se compose d'un amplificateur de détection à équilibre de gain d'une très haute adjustabilité dont la sortie est utilisée pour faire dévier la trace sur le tube dans une direction verticale, la déviation horizontale étant réalisée par connexion directe à l'oscillateur.

Des contraintes dynamiques périodiques apparaissent sur le tube sous forme de deux triangles isocèles rejoints à leur sommet. Si les deux côtés en pente des triangles sont réglés par les commutateurs à décade de manière à être horizontaux à tour de rôle, alors la différence entre les deux lectures de cadran constitue l'amplitude de contrainte.

La précision du pont qui est permanente est d'environ 3 micropouces par pouce et elle est insensible aux variations de secteur. La gamme du pont va de −10 000 à +22 000 micro-pouce par pouce; il est robuste et presque impossible à endommager.

EE 03 757 *pour plus amples renseignements*

PLAQUETTES DE PROGRAMMATION

D.E.B. Electronics Ltd, New Barnes Hill, Cottonmill Lane, St. Albans, Hertfordshire

(Illustration à la page 122)

Un nouveau système de programmation peu coûteux a été réalisé par la société D.E.B. Electronics Ltd. Il se compose de deux types de bande de cuivre parallèles à angle droit, l'un par rapport à l'autre, sur les faces oposées d'un matériau de base laminé d'une épaisseur de 0,15 cm. Il y a une ouverture à chacune des intersections des deux réseaux de bande de cuivre et ces trous se trouvent sur une matrice de 0,300 pouce. Les bandes de cuivre sont d'une épaisseur de 0,300 pouce et elles sont doublées de nickel d'une épaisseur de

0,0002 pouce et d'or dur d'une épaisseur de 0,0001 pouce.

Sur la version à fiches de la matrice toutes les bandes sont reliées à un connecteur de tranche. Il existe également une version à montage affleurant.

La programmation de la version à fiches s'effectue en reliant des pistes sur les côtés opposés de la plaquette à travers les trous, soit au moyen d'une broche soit à l'aide d'un porte-composant. Ce dernier peut recevoir une gamme étendue de composants, de diodes, de résistances, de condensateurs, etc. Une lampe à faible tension miniature est en cours de mise au point pour servir de broche de programme. Les broches sont livrables avec le même revêtement que le panneau et avec des capsules en couleur.

Les panneaux sont construits suivant la spécification du client et comprennent tout le lettrage, la gravure, etc. Un nombre d'étalons est prévu.

EE 03 758 *pour plus amples renseignements*

BOUTON-POUSSOIR À LAME VIBRANTE ET COMMUTATEUR DE BASE

Distributeurs: Kynmore Engineering Co. Ltd. 19 Buckingham Street, London, W.C.2

(Illustration à la page 122)

Le commutateur de base Rotireed type RG est du modèle postal standard ou du modèle Kellog, sauf que les contacts sont des éléments à lame vibrante miniature entièrement enfermés et montés horizontalement sur des plaquettes de circuit imprimé vertical. Ils sont mis en action par de petits aimants permanents montés sur une extension de la clé d'entraînement. Le commutateur a une position neutre centrale ainsi que des positions de rupture dont chacune peut comporter deux ou quatre contacts normalement ouverts. Etant donné qu'ils se ferment en séquence suivant la vitesse et la direction du fonctionnement manuel, ils peuvent être utilisés comme contacts d'ouverture ou de fermeture. Le petit panneau frontal peut être marqué "arrêt-marche" ou 1-0-2 ou porter n'importe quelle indication brève. Les contacts à lame vibrante miniature utilisés peuvent commuter une charge de contact maxima de 6 W, avec un niveau maximum de 0,3 A à 250 V, 50 Hz ou 26 V c.c.

Le commutateur à bouton-poussoir type Rotireed RT que l'on voit dans notre gravure peut être fourni dans des bandes de 1, 4, 6 et 10 clés standard. Des éléments comportant jusqu'à 25 clés peuvent être fournis sur commande spéciale. On peut monter un maximum de quatre contacts normalement ouverts ou normalement fermés dans n'importe quelle combinaison. Les clés peuvent être à auto-verrouillage lorsqu'elles sont pressées et déverrouillées par une nouvelle pression. Le verrouillage électromagnétique est également prévu. Les couleurs de clé standard sont le blanc ou le noir. Les clés peuvent porter des chiffres d'une couleur différente sur demande. D'autres couleurs sont également prévues sur demande spéciale. Des clés éclairées par l'intérieur peuvent être fournies. Lorsque quatre jeux de contacts par bouton-poussoir sont utilisés, des sorties à code binaire peuvent être effectuées. Des contacts standard sont utilisés, capables de commuter 15 W avec un niveau maximum de 1 A à 250 V. Le panneau frontal de l'élément à 10 boutons mesure 21,59 cm × 2,22 cm. Tous les éléments RT ont une profondeur totale de 14,60 cm.

EE 03 759 *pour plus amples renseignements*

VOLTMÈTRE NUMÉRIQUE

Sapphire Research and Electronics Ltd, Sapphire Works, Ferndale, Glamorganshire

(Illustration à la page 122)

Le voltmètre, modèle 250, de la société Sapphire Research and Electronics Ltd utilise des circuits semi-conducteurs d'un nouveau modèle et comprend un système à asservissement à tension alternative. Il indique les tensions continues à partir de 1000 V et les tensions alternatives à partir de 500 V efficaces et, de plus, il comprend des gammes de tension continue et de tension alternative.

Il peut résoudre des différences de tensions d'un minimum de 200 μV. La gamme de fréquences alternatives s'étend de 10 Hz à 60 kHz. L'étalon de fréquence est de 0,1 % de déviation totale sur les gammes de tension continue et de 0,25 % sur les gammes de tension alternative. L'impédance d'entrée de l'instrument est élevée, étant de l'ordre de 1000 MΩ sur la gamme de 1V et de 2,2 MΩ sur les autres gammes de tension. Le nouvel instrument a une performance d'étalonnage de zéro fixe très stable et n'exige que très peu de surveillance. La lecture est claire dans toutes les conditions d'éclairage ambiant et la position de la décimale est automatique sur toutes les gammes. La protection contre les surtensions est également prévue sur toutes les gammes.

L'instrument a été conçu à l'usage des laboratoires et de l'industrie et assure un rendement élevé. C'est un appareil portatif d'excellente qualité et à très bas prix.

EE 03 760 *pour plus amples renseignements*

ENREGISTREUR MULTIVOIES

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Illustration à la page 122)

La société S.E. Laboratories (Engineering) Ltd a réalisé un oscillographe, modèle SE.3006, à bas prix et d'un poids réduit. Une attention très stricte à la valeur, à la conception et une construction entièrement modulaire à fiche ont permis de mettre au point un enregistreur de trace qui comporte les avantages et la souplesse qui sont normalement le propre d'instruments d'un prix beaucoup plus élevé.

L'oscillographe est fourni soit comme instrument à six voies soit comme instrument à douze voies, utilisant des galvanomètres anglais classiques et enregistrant sur papier sensible de 15 cm de large. Il s'agit d'un appareil portatif ne mesurant que 17,78 cm de hauteur, pesant une vingtaine de kilos et pouvant être monté sur table ou banc d'essai.

Une caractéristique importante de l'enregistreur est sa faible consommation électrique de 250 VA, tout en employant la lampe ultraviolette à vapeur de mercure à grande vitesse d'écriture. On peut choisir instantanément pendant le fonctionnement une gamme de 1500:1 de vitesse de papier, allant de 5mm/min à 1250 mm/sec, par boutonpoussoir à huit plots. Des lignes de minutage à largeur totale de 0,01, 0,1 et 1 sec et 0,1 et 1 min d'intervalle ainsi que des lignes de grille de 2 mm sont imprimées sur l'enregistrement. Des commandes de trace individuelles et d'intensité de grille sont prévues ainsi que la commande à distance.

La charge de papier est effectuée facilement en insérant un rouleau de papier dans la cassette, le filetage étant automatique.

EE 03 761 *pour plus amples renseignements*

INSTRUMENTS À VALEUR EFFICACE RÉELLE

Distributeurs: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustration à la page 123)

Les instruments à valeur réelle "Transquare" Greibach ont été conçus pour la mesure de haute précision sur les formes d'ondes complexes et ils sont particulièrement indiqués pour les opérations avec les redresseurs pilotés au silicium (thyristors), dont en particulier la mesure de précision de l'entrée de puissance dans le matériel thermique à redresseur piloté au silicium. Les instruments à valeur efficace électroniques conventionnels ne sont capables de précision que dans la région de 2 à 3 %, par rapport à la précision de 0,5 % de l'instrument Transquare.

La gamme de voltmètres à valeur efficace réelle s'étend de 30 mV à 1000 V, sur la totalité de l'échelle.

La capacité de mesure du courant va jusqu'à 200 A autonomes et 1000 A avec transformateur de courant externe ou shunt (pour la mesure de courant continu ou de courant alternatif). La sensibilité maxima a été jusqu'ici de 0,1 mA sur la totalité de l'échelle mais la société Greibach a maintenant mis au point un microampèremètre à valeur efficace réelle avec une faible gamme sur la totalité de l'échelle de 50 μA.

Le type 500T005-8, qui est caratéristique de ces nouveaux modèles sensibles, comporte huit gammes de 0,05 mA à

100 mA. Il peut atteindre une précision de 0,5 % sur la totalité de l'échelle pour des formes d'onde irrégulières avec facteurs de crête dépassant 4, ainsi que des formes d'onde carrées jusqu'à plusieurs kHz. Une stabilité exceptionnelle avec une répétabilité supérieure à 0,1 % est obtenue car ces instruments ne comportent ni amplificateurs, ni alimentations, ni d'autres composants actifs. Ils sont capables de résister 100 fois au courant aux gammes de courant les plus sensibles et ils répondent aux fréquences d'entrée du courant continu et de 5 Hz à 500 kHz avec une précision de ±0,5 % sur la totalité de l'échelle aux gammes les plus sensibles.

EE 03 762 *pour plus amples renseignements*

PONT ET AMPLIFICATEUR À TENSION CONTINUE

Vibro-Meter SA., Haletop Civic Centre, Wythenshawe, Manchester, 22

(Illustration à la page 123)

La société d'instruments suisse, Vibro-Meter SA, vient d'introduire un bloc d'alimentation en continu entièrement transistorisé et complet avec dispositifs d'équilibrage et d'étalonnage. L'appareil SPB-1/A a été conçu pour l'emploi avec des extensomètres d'une résistance de 60 à 600Ω et des cellules de charge, des accéléromètres et des transducteurs de puissance du type extensométrique. Un amplificateur de tension à large bande équilibrée (DCA-1/A) est également prévu pour l'amplification des signaux si nécessaire.

Le bloc d'alimentation SPB-1/A utilise un circuit en pont entièrement flottant convenant tant pour les systèmes à pont complets que pour les systèmes à demi-pont et ayant un isolement minimum à la masse de $10^9\Omega$. La mesure s'effectue par la méthode de compensation ou d'effacement et le réglage de l'équilibre du pont peut être reproduit de manière exacte.

L'instrument comprend un réseau d'équilibrage qui permet d'effecteur la compensation du déséquilibre zéro dans le transducteur et la compensation du poids de la tare jusqu'à 300% de la valeur nominale. La mesure finale est réalisée par la méthode de compensation; un dispositif d'étalonnage incorporé est prévu et des signaux négatifs ou positifs proportionnels à la microtension peuvent être obtenus.

L'amplificateur connexe, le Vibro-Meter, type DCA-1/A, a une valeur de dérive nulle qui n'est que de 3 μV/°C. Cet instrument qui est entièrement transistorisé est fondamentalement un amplificateur de courant continu de haute stabilité à couplage direct avec entrée différentielle, et gain variable de 3 à 1000. Il comporte des commandes d'équilibre zéro et de rejet de mode commun. Des sorties de ±10 V, ±2 mA et ±10 V ±100 mA peuvent être obtenues.

EE 03 763 *pour plus amples renseignements*

ENREGISTREUR AUTOMATIQUE DU POINT DE ROSÉE

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Illustration à la page 123)

L'enregistrement automatique du point de rosée de l'air ou des gaz est actuellement possible grâce au nouvel enregistreur automatique Shaw de point de rosée. Ce nouvel instrument est entièrement constitué de corps solides et l'enregistreur fournit des diagrammes quotidiens, hebdomadaires ou mensuels. Les indications absolues de point de rosée sont données par un miroir, refroidi par l'effet Peletier, la formation de rosée étant perçue par une cellule photoélectrique et la température du miroir est indiquée sur le cadran de l'instrument de mesure de 12,5 cm et enregistrée également.

L'enregistreur automatique de point rosée fournit une indication positive du point de rosée réel au moyen d'un voyant lumineux s'éclairant toutes les secondes et indiquant la présence ou l'absence de rosée sur le miroir.

Les gammes normales vont de −60° C à −20° C et de −20° C à +20° C de point de rosée, bien que n'importe quelle gamme puisse être prévue sur commande spéciale. L'enregistreur ne comporte pas de commande de réglage et aucun service externe en dehors de l'alimentation secteur n'est nécessaire.

Un modèle pour les besoins météorologiques ou pour l'emploi dans des conditions de température ambiante a été également mis au point.

EE 03 764 *pour plus amples renseignements*

COMPTEUR NUMÉRIQUE

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustration à la page 124)

Ce nouveau compteur électronique a la même gamme de fréquence étendue et la même capacité d'accessoires à fiche sur le panneau frontal que des compteurs plus coûteux mais, en raison de l'omission de certains raffinements, son prix est considérablements moins élevé. Le compteur de base a un taux de comptage maximum de 50 MHz et il peut mesurer des fréquences avec précision et avec une résolution de 1 Hz à 12,4 GHz avec des transformateurs de fréquence Hétérodyne à fiche HP. Bien qu'il s'agisse d'un modèle économique, le nouveau compteur a une gamme complète de temps de porte de 1 μsec à 1 sec.

Le nouveau compteur (modèle HP 5246L) reçoit toutes les fiches conçues pour le compteur modèle HP 5245L de 50 MHz. En dehors des mesures de fréquence allant jusqu'à 12,4 GHz à l'aide de fiches transformatrices, le nouveau modèle économique peut mesurer des fréquences directement jusqu'à 350 MHz avec une fiche spéciale. Il peut, en outre, être utilisé avec une sensibilité de 1 mV avec un dispositif à fiches préamplificateur (sensibilité du compteur de base: 100 mV), normaliser des lectures avec un dispositif à fiches compteur préréglé ou mesurer les intervalles de temps de 1 μsec à 10^6 sec et effectuer des mesures de tension numérique jusqu'à 1000 V avec deux autres dispositifs à fiches.

Le nouveau compteur offre une lecture à six chiffres. En variante, un septième et un huitième chiffre peuvent être ajoutés pour fournir une résolution plus élevée au moment de l'achat ou par la suite. La précision de base de l'instrument est déterminée par l'oscillateur de la base de temps dont le taux de vieillissement est inférieur à ±2 parties dans 10^7 par mois. Une base de temps à haute stabilité peut être fournie en variante; son taux de vieillissement est inférieur à ±3 parties dans 10^9 par 24 heures et elle a une stabilité excellente à court terme (inférieure à ±2 parties dans 10^{10} avec un temps de moyenne de 1 sec). Le nouveau compteur peut être utilisé également à partir d'un étalon de fréquence extérieur et il peut mesurer les rapports de fréquence.

L'impédance d'entrée de l'amplificateur est de 1 MΩ, shuntée par 25 pF pour tous les niveaux de signaux jusqu'à 1 V, indépendamment de la fréquence jusqu'à 50 MHz. Grâce à l'impédance d'entrée constamment élevée, on est assuré d'un faible rapport d'amplitude de tension lorsque l'entrée est relié à une ligne à terminaison et on peut utiliser une sonde de type oscilloscopique lorsqu'on désire une atténuation de tension ou une entrée de 10 MΩ, 10 pF.

Le nouveau compteur comporte également, comme facilité facultative, une sortie pour code "1" positif de 1-2-2-4 ou pour code 1-2-4-8 avec état "1" positif ou négatif.

EE 03 765 *pour plus amples renseignements*

MACHINE D'ESSAI DE DURETÉ À ULTRASONS

Dawe Instruments Ltd, Western Avenue. Acton, London, W.3

(Illustration à la page 124)

Un nouvel instrument pour l'essai de précision de microdureté des matériaux. appelé le contrôleur de microdureté Sonodur, type 1851, est maintenant réalisé par la société Dawe Instruments Ltd. Cet appareil a une gamme de 20 à 70 Rockwell C et une précision de ±1 Rockwell C.

Durant l'utilisation, le composant devant être contrôlé est placé sur la table de la base. La sonde montée sur colonne, à l'extrémité de laquelle se trouve un diamant Vickers, est abaissée sur le composant sous gravité. La charge statique appliquée n'est que de 150 g. La sonde est excitée à sa fréquence résonante naturelle par un générateur à ultrasons logé dans la base. La fréquence de vibration de la sonde change en fonction de la profondeur de pénétration de sa pointe dans le composant.

Les changements de fréquence sont mesurés électroniquement et affichés sur un instrument de mesure qui est également logé dans la base. L'échelle de mesure est étalonnée directement en degrés de dureté.

Le nouveau modèle diffère du Sonodur type 185A présenté l'année dernière par le fait que la sonde est montée sur colonne au lieu d'être tenue à la main. Une plus grande uniformité de charge assure une plus grande précision de lecture. Le type 1850 continue cependant a être livrable en partie pour les essais sur place et en partie pour les travaux sur les spécimens trop grands pour être examinés par le type 1851.

Etant donné que les lectures sont effectuées pendant que la sonde est en contact avec le spécimen, les erreurs dûes au rétablissement élastique sont éliminées. Chaque lecture prend moins de 10 secondes et n'exige aucun talent ou aucune formation préalable. La profondeur de pénétration dans l'acier est inférieure à 0,0002 pouce, même sur les matériaux entièrement recuits et la zone d'indentation est inférieure de 0,1 % à la zone d'indentation produite par un contrôleur Rockwell C.

Le Sonodur type 1851 obvie à la nécessité d'employer des instruments de mesure optiques ou mécaniques. En raison de la lecture électrique, n'importe quelle portion de l'échelle peut être facilement étendue pour faciliter la lecture lorsque l'appareil est utilisé pour les essais de production, ce qui réduit la possibilité de lecture d'erreurs.

L'instrument est fourni comme appareil standard étalonné pour mesurer la dureté de l'acier (module de Young de 29-30 millions) mais il peut être réglé pour contrôler tous les autres métaux par simple étalonnage en fonction de deux échantillons de dureté connus. Une échelle supplémentaire, étalonnée en unités arbitraires marquée de 0 à 100, est fournie pour les essais de comparaison avec d'autres matériaux.

EE 03 766 *pour plus amples renseignements*

OSCILLATEUR À EFFET GUNN

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

La société Mullard Ltd vient de présenter le premier oscillateur à effet Gunn pour la bande X (7 à 12 GHz).

Mis au point par les laboratoires de Wembley de la société Associated Semiconductor Manufacturers Ltd et par les laboratoires de recherches Mullard, le nouveau dispositif fournit une sortie continue de 5 mW et il est alimenté par un simple courant continu de 6 V. Il ne mesure que 5 mm de longeur × 3 mm de diamètre.

En raison de son encombrement extrêmement réduit et de sa simplicité de fonctionnement, il constitue une variante intéressante de l'oscillateur local du type à transistor/varactor ou du modèle classique à lampes klystron utilisé dans les étages mélangeurs des faisceaux de communication microondes et les systèmes de radar. Il peut être utilisé également comme source directe d'énergie microondes dans les systèmes de radar de faible puissance (par exemple le matériel de contrôle de vitesse) ainsi que dans les systèmes de démonstration de microondes pour les écoles et collèges techniques.

Durant l'emploi, le type 106CXY, est accouplé à une simple cavité microondes, dont la fréquence résonante détermine la fréquence de sortie. La cavité peut être une fréquence fixe ou accordée à la gamme de 7 à 12 GHz.

Les oscillateurs à effet Gunn précédents avaient été produits en créant un champ approprié entre les deux faces d'un mince tranche d'arsenure de gallium (GaAs): plus la tranche était mince plus la fréquence de fonctionnement était élevée. Jusqu'à présent, ces tranches avaient été réalisées par la méthode de meulage et de polissage qui limitait la fréquence de fonctionnement maxima à environ 4 GHz.

Pour surmonter ce problème, la société Mullard a mis au point une technique de dépôt épitaxial qui permet à une couche très mince d'arsenure de gallium de se former sur une tranche plus épaisse du même matériau. Tout dispositif produit de cette manière a son élément actif à la surface dans une couche à haute résistivité sur une couche subjacente à faible résistivité, l'effet se produisant à travers la couche.

Non seulement cette technique augmente la fréquence de fonctionnement du dispositif, mais elle assure également une plus grande commande de résistance pour une couche de résistivité donnée ainsi que pour une dissipation de puissance plus élevée.

EE 03 767 *pour plus amples renseignements*

COMMUTATEUR MICROONDES

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

(Illustration à la page 124)

La société Microwave Associates vient d'annoncer la mise au point d'un nouveau commutateur à diode coaxial avec isolement de 60 dB, un taux d'onde stationnaires inférieur à 0,07 et une perte par insertion inférieure à 0.5 dB dans une largeur de bande de 20 % dans la gamme de fréquence de 1 à 2 GHz.

Pour les largeurs de bande allant jusqu'à 50 %, ce commutateur, le MA8302-2L4S, offre un isolement supérieur à 40 dB, un taux d'onde stationnaires inférieur à 1,20 et une perte par insertion inférieure à 0,75 dB.

Conçu principalement pour le matériel de contrôle automatique à microondes, en liaison avec les ponts de phases, le nouveau commutateur est également indiqué pour les applications de commutations "à chaud" lorsque deux transmetteurs sont branchés sur une sortie commune ou pour la commutation alternante entre deux antennes.

EE 03 768 *pour plus amples renseignements*

CONTRÔLEURS DE COURANT

Scama Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(Illustration à la page 125)

La société Scama Ltd, constructeurs des annonceurs "Panalarm," présente un nouveau contrôleur de courant, le "Trip-Matic," série 70, qui peut être utilisé soit comme élément à part pour cabine de contrôle soit pour le montage sur panneau arrière soit encore pour faire partie intégrante d'un annonceur "Panalarm", modèle 75, relié directement aux entrées analogiques.

Dans une application typique, le Trip-Matic contrôle un signal de courant continu proportionnel au niveau du paramètre de fonctionnement. Le signal de courant continu monte à mesure que monte la valeur du paramètre. Quand ce signal dépasse une référence réglable préajustée, un comparateur-amplificateur accuse un changement marqué de sortie qui indique positivement l'annonceur ou d'autres dispositifs commandés.

Le nouveau contrôleur de signaux serait d'une précision exceptionnelle et aurait une réponse extrêmement rapide. Conçu pour fonctionner dans une bande étendue de signaux de courant continu, il peut également être adapté pour contrôler de nombreuses applications à courant alternatif. Il est entièrement transistorisé et muni de semi-conducteurs au silicium. Toutes les entrées sont isolées les unes des autres ainsi que des sorties. Le différentiel entre le déclenchement et le réenclenchement peut être fixé ou être réglable, soit avec un fonctionnement de déclenchement faible ou élevé pour répondre aux conditions d'utilisation.

EE 03 769 *pour plus amples renseignements*

VOLTMÈTRE DIFFÉRENTIEL

Distributeurs: General Test Instruments Ltd, Gloucester Trading Estate, Hucclecote, Gloucester

(Illustration à la page 125)

Le nouveau voltmètre Dialamatic fabriqué par Wavetek est un appareil manuel dont la vitesse approche celle des voltmètres automatiques mais qui offre la précision, la robustesse et le prix de modèles différentiels classiques.

La clé de cette performance est le commutateur Transfermatic. Ce dispositif effectue le couplage mécanique de commutateurs de décades et les reports sont transférés à la décade plus élevée suivante. Au moment d'approcher un point de transfert important, comme 2,999, une simple pression du bouton de droite déplace automatiquement la lecture à 3,000. Ce processus s'effectue également dans le sens inverse.

Pour obtenir le même résultat avec un voltmètre différentiel conventionnel, il faut réenclencher chaque bouton individuellement.

Le Dialamatic comporte également un amplificateur de zéro rapide ne se saturant pas qui permet à l'opérateur de mettre en oeuvre l'équilibrage correct sans changer la sensibilité à zéro.

En plus de la vitesse de fonctionnement, le Dialamatic a été conçu pour l'entretien facile et rapide. Chacun des circuits est monté sur un carton modulaire à fiches qui pivote hors du coffret au cas ou l'entretien serait nécessaire. Tous les semi-conducteurs au silicium ont été utilisés pour assurer la fiabilité dans les conditions d'environnement extrêmes.

Le Dialamatic est prévu en deux modèles de base. Le modèle 201 mesure les tensions continues de 0 à 1000 V dans quatre gammes et le modèle 202 est à courant alternatif et à courant continu et sa gamme de mesure s'étend de 1 mV à 1000 V. Tous les modèles peuvent être obtenus en version à batterie avec rechargeur incorporé.

EE 03 770 *pour plus amples renseignements*

RÉSISTANCES VARIABLES DE PRÉCISION

Steatite Insulations Ltd, Hagley House, Hagley Road, Birmingham 16

Cette résistance variable de précision, appellée Varizistor, allie les fonctions d'une résistance de précision à l'ajustabilité d'un potentiomètre trimmer dans un ensemble qui fournit une résolution infinie. Des niveaux de résistance extrêmement précis peuvent être atteints facilement, adaptant ainsi la résistance aux besoins spécifiques du circuit. Le dispositif se compose d'un élément de résistance à film mince métallique, aggloméré à une couche sous-jacente en céramique et introduit dans un logement de forme rectangulaire. Un curseur à plusieurs contacts est fixé au câble et isolé sur un des côtés. Le conducteur traverse le corps, permettant ainsi à la pince d'entrer en contact avec l'élément de résistance de manière à fournir une continuité de circuit. Le corps entier glisse à l'intérieur de limites fixes, déplaçant le curseur à travers l'élément de résistance. Une variation de résistance d'environ 20 % au-dessus et au-dessous de la valeur nominale peut être obtenue entre 0,5 Ω à 1 MΩ, réglable jusqu'à 0,01 %. Pour le réalisateur de circuits, le dispositif fournit l'occasion d'expérimenter avec des composants de circuit sans les problèmes habituels inhérents aux résistances fixes. Dans les applications de circuits sensibles, des changements de résistance très réduits peuvent être effectués au cours du contrôle final après soudage des conducteurs pour compenser les tolérances de composants et la résistance parasite.

Les Varizistors ont été conçus pour répondre ou dépasser toutes les conditions de spécification de haute fiabilité. Les conducteurs peuvent être soudés à n'importe quel circuit. Des valeurs de résistance précises peuvent être fixées en soudant les conducteurs à la jonction du corps. Les capsules d'extrémité peuvent être soudées à la jonction du corps pour assurer une fermeture hermétique. Les modèles prévus sont d'une puissance nominale de $\frac{1}{4}$, $\frac{1}{2}$ et 1 W et des coefficients spéciaux de température peuvent être commandés jusqu'à des valeurs minima de $5 \times 10^{-6}/°$ C.

EE 03 771 *pour plus amples renseignements*

Résumés des Principaux Articles

Un amplificateur de puissance V.H.F. constitué de corps solides par T. Spencer

Résumé de l'article aux pages 72 à 77

Il s'agit ici d'une technique de réalisation pour les amplificateurs de puissance de la Classe C, basée sur les données standard publiées par les fabricants de composants. L'analyse du fonctionnement des transistors de la Classe C est identique à celle utilisée pour les tubes et les courbes de réalisation normalisées. Par l'emploi d'un processus pas à pas, les calculs de mise au point se comparent favorablement aux résultats pratiques.

Plusieurs circuits multiplicateurs et d'amplificateurs de puissance sont décrits, de même que les circuits d'entraînement d'oscillateurs pilotés par quartz et à marche libre. Les facteurs qui influent sur l'efficacité et la puissance de sortie sont analysés.

La simulation du cône de confusion par S. Dalal

Résumé de l'article aux pages 78 à 81

La réalisation de la simulation de la navigation par radioguidage pour un simulateur de vol pose souvent aux ingénieurs le problème de la simulation du cône de confusion qui se trouve au sommet d'une radio-balise. Le système décrit dans cet article est relativement peu coûteux, entièrement transistorisé et il constitue un moyen intégré de simuler un cône de confusion réaliste.

Un détecteur de loi quadratique pour la mesure de puissance par J. F. Golding

Résumé de l'article aux pages 82 à 86

Le montage de circuit de détection décrit dans cet article a été conçu à l'origine pour produire une déviation oscilloscopique proportionnelle à la sortie de puissance moyenne d'un récepteur. Le principe peut cependant être appliqué dans d'autres manières y compris l'emploi du circuit comme doubleur de fréquence apériodique.

Un magnétophone à modulation de fréquence simple par W. Tempest et M. E. Bryan

Résumé de l'article aux pages 87 à 89

Cet article décrit un appareil de modulation de fréquence transistorisé pouvant être utilisé avec n'importe quel magnétophone domestique ou semi-professionnel pour enregistrer et réenregistrer des signaux dans une gamme de fréquence allant du courant continu à 1250Hz. Ce système a un rapport signal/bruit de 42dB et sa stabilité est telle que dans une période de 20 minutes un signal de courant alternatif reproduit par le système ne change pas de plus de 0,1dB et un niveau de courant continu dérivera de manière typique d'environ 5mV. La gamme d'entrée de signaux et de ±5V et la sortie est inférieure d'environ 20dB à ±0,5V.

Le dessin d'alimentations de photomultiplicateurs par R. Barlow et J. Watson

Résumé de l'article aux pages 90 à 96

Les exigences de puissance des photomultiplicateurs sont étudiées et il est montré que les circuits connexes et le mode de fonctionnement de ces dispositifs peuvent parfois permettre des spécifications moins rigides pour les alimentations. Le dessin de transformateurs inverseurs à haute tension et de leur circuit stabilisateur est considéré et les avantages pouvant être gagnés de l'emploi d'étages comparateurs à transistors à effet de champ sont présentés.

Dessin simplifié de la paire à réaction par J. C. S. Richards

Résumé de l'article aux pages 97 à 99

Le dessin d'un étage amplificateur de courant alternatif à deux transistors avec réaction en dérivation est réduit à l'application de quelques formules très simples couvrant la polarisation de tension continue, le gain et la largeur de bande. Les approximations et les hypothèses qui en découlent sont valables pour la plupart des transistors planaires au silicium de faible puissance; le rendement obtenu est suffisament près de la valeur optimum pour de nombreux usages.

Economie et fiabilité dans le dessin d'instruments nucléaires par B. Shepherd

Résumé de l'article aux pages 100 à 105

L'auteur décrit une série d'instruments réalisés de façon sûre mais peu coûteuse. De nouveaux circuits ont été employés dans le dessin d'une plaquette à décades rapide avec indication visuelle, d'un élément à haute tension sans transformateur stable et de discriminateurs de hauteur d'impulsion à diode tunnel.

Un instrument numérique pour mesurer le décalage de moteurs asynchrones par V. Böhm, V. Popescu et T. Teitel

Résumé de l'article aux pages 106 à 109

Cet article décrit un instrument numérique entièrement transistorisé qui fournit une lecture directe sous forme décimale du décalage du moteur à induction par couplage du pick-up photoélectrique à un disque monté sur l'arbre du moteur. Grâce à l'emploi exclusif des mèthodes de comptage d'impulsions, cette technique est entièrement exempte des erreurs qu'on rencontre dans les autres méthodes pour la mesure de ce paramètre.

Multiplication à haute fréquence par l'emploi de diodes logarithmiques par R. P. Ingram

Résumé de l'article aux pages 110 à 112

Une méthode de multiplication aux fréquences élevées à l'aide de la modulation à anneau en pont de diodes logarithmiques est décrite par l'auteur. La théorie indique qu'à condition de remplir certaines conditions d'entrainement et de sortie on obtient une très bonne approximation de la multiplication réelle. Ces multiplicateurs peuvent être construits pour fonctionner jusqu'à 100kHz avec peu de difficulté. Avec un choix approprié de diodes des résultats assez bons peuvent être obtenus aux fréquences plus élevées. Des dispositifs de ce genre peuvent être forts utiles pour l'analyse de corrélation et pour l'emploi dans les systèmes de balayage.

Une nouvelle source de référence de tension par T. P. Dawson et G. C. Taylor

Résumé de l'article à la page 113

Un circuit stabilisateur de tension à montage en pont utilisant des dispositifs semi-conducteurs a été mis au point pour produire des rapports de stabilisation dépassant 700, comparés à des rapports d'environ 25 de circuits stabilisateurs classiques utilisant deux diodes zéner.

Une montre actionnée par la voix par K. J. Kapota

Résumé de l'article à la page 114

Le composant en question est une montre électrique qui peut être mise en marche ou arrêtée au moyen d'un microphone de gorge et qui comporte un circuit électronique permettant de mesurer le temps mis par une personne à parler. Ce dispositif a été prévu pour l'emploi dans les domaines de la psychologie et de la psychiâtrie.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 120 bis 125

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Magnetbandtester

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Abbildung Seite 120)

Gresham Lion Electronics Ltd hat einen neuen Magnetbandtester entwickelt, der die volle Breite des Bandes auf Beschichtungslöcher und andere Fehler prüft.

Gresham bietet eine breite Auswahl von Spurenanordnungen an, und die für EDV am besten geeignete Ausführung für 12,7-mm-Magnetband hat einen Aufnahmekopf für die volle Breite und einen Doppelwiedergabekopf mit sieben und neun Spuren (Typ P.S.79). Die obere und untere Kante des Magnetbandes läuft in gefederten keramischen Bandführungen; die Kopfanordnung wird durch eine Haube geschützt.

Elektrische und mechanische Kenndaten können spezifischen Anwendungen angepasst werden. Das Pflichtenblatt für Wiedergabeköpfe mit sieben und neun Spuren für 12,7-mm-Magnetband ist wie folgt:

Aufnahmekopf

Spurbreite, 12,83 mm ±0,0025 mm
Spaltbreite, 12,7 µm (nominell)
Widerstand, 25 Ω
Induktivität, 7 mH
Sättigungsstrom, 33 mA_{ss}

Wiedergabeköpfe	*7 Spuren*	*9 Spuren*
Spurbreite	0,76 mm ±0,0025 mm	1,02 mm ±0,0025 mm
Spaltbreite	6,35 µm	6,35 µm
Widerstand	11,4 Ω	12,5 Ω
Induktivität	5,6 mH	5,3 mH

EE 03 751 *für weitere Einzelheiten*

Potentiometerzusatz

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Abbildung Seite 120)

Der Potentiometerzusatz TA61 ist ein preisgünstiges und einfaches Gerät, das die Erregung strombetriebener Lasten durch das Ausgangssignal von Potentiometergebern ermöglicht.

Bei Einsatz dieses Gerätes lassen sich robuste Grossskaleninstrumente, Direkt- und Lichtpunktschreiber von Weg-, Druck- und anderen Gebern, die mit Potentiometerfühlern arbeiten, treiben. Der normalerweise beim direkten Treiben von strombetriebenen Lasten durch den Ausgang von Potentiometergebern auftretende nichtlineare Fehler wird durch eine einzigartige Schaltung, auf der der TA61 beruht, beseitigt.

Der in erster Linie für Einsatz mit Intersonde-Gebern entwickelte Zusatz ist jedoch mit jedem beliebigen Potentiometergeber, dessen Abschlusswiderstand zwischen 500 Ω und 10 kΩ liegt, verwendbar. Eine vor allem im industriellen Einsatz sehr nützliche Eigenschaft des Zusatzes ist, dass man ihn viele hundert Meter vom Geber entfernt positionieren kann. Die an das Gerät angeschlossene Last muss zwischen 0 und 1000 Ω Widerstand haben und kann, wenn erforderlich, auch entfernt liegen.

Üblicherweise wird der TA61 mit Gebern eingesetzt, die statische oder sich langsam ändernde Parameter messen, ist aber genauso für dynamische Anwendung geeignet. Der Frequenzgang ist zwischen 0 und 10 kHz geradlinig, und die Genauigkeit soll über den Temperaturbereich 0...60° C bei ±0,2 Prozent des Ausgangswertes liegen.

EE 03 752 *für weitere Einzelheiten*

ESB-Sende-Empfangsgerät

Redifon Ltd, Broomhill Road, London, S.W.18

(Abbildung Seite 120)

Der Bereich Nachrichtenverkehr der Redifon Ltd hat ein neues HF-ESB-Sendeempfangsgerät mit 100 W Oberstrichleistung entwickelt, das vor allem den Anforderungen der Streit- und Schutzkräfte bezüglich maximaler Frequenzvielseitigkeit bei günstigem Preis genügt. Es liegt ungefähr halbwegs zwischen den bestehenden ZAM-Ausrüstungen mit durchstimmbarem Oszillator und den ESB-Geräten der Zukunft mit voller Frequenzsynthese und ist hauptsächlich für Organisationen bestimmt, die Einführung von ESB-Betrieb in Betracht ziehen.

Einfachheit und Konstanz der Quarzsteuerung sind im G.R.410T auf wirtschaftliche Weise mit der Vielseitigkeit der Frontplatteneinrichtungen zum schnellen Auswechseln der Quarze und Umstellung auf neue Frequenzen verbunden. Für die Abstimmung des Oszillatorausgangs, der HF-Stufen und Belastung des Endverstärkers gibt es handbetätigte Bedienorgane. Ein eingebautes Messgerät überwacht Treibpegel, Anodenstrom des Endverstärkers, Vorwärtsantennenstrom und Speisespannungen. Frequenzen können ohne Servicepersonal und Nachabgleichung in der Werkstatt in wenigen Minuten an Ort und Stelle geändert werden.

Das Sende-Empfangsgerät kann jederzeit bis zu vier, durch Schalter wählbare Quarze aufnehmen, und die Bedienung kann weitere Quarze für andere Frequenzen mit sich führen. Dadurch kann das Gerät schnell an andere Standorte verlegt werden, in denen andere Frequenzen benutzt werden. Ausserdem kann das Sende-Empfangsgerät Eingangssignale von externen Quellen wie z.B. einem Frequenzsynthesegerät aufnehmen.

Der G.R.410T überstreicht den Bereich 2...16 MHz mit Schalterwahl des oberen und unteren Seitenbandes und arbeitet mit Telefonie (A3J) und modulierter Trägertelegrafie (A2J) in der ESB-Arbeitsweise; ein eingeschalteter Träger gibt kompatible AM (A3H und A2H) für Verkehr mit Netzen, die noch nicht ESB eingeführt haben. Ausserdem kann das Gerät im Al-Betrieb arbeiten. Durch einen Frontplattenschalter kann man auf automatische Sende-Empfangsumschaltung umstellen. Automatische Pegel-regelung verhindert Übermodulieren und gewährleistet bei niedrigem Sprechpegel ausreichende Modulation.

Ein von Hand bedientes Redifon-Antennenabstimmgerät zum Anpassen des Sende-Empfangsgerätes an die Peitschenantenne eines Fahrzeugs oder eine einfache Drahtantenne ist auch lieferbar. Modell GR.410T kann auch direkt mit einem 75-Ω-Dipol oder über einen Breitband-Anpassübertrager mit einer 600-Ω-Antenne betrieben werden.

Stromversorgungen sind für Anschluss an 100...125 V~, 200...250 V~ 50 oder 60 Hz, sowie für Betrieb an 12-V- oder 24-V-Batterien lieferbar, so dass das GR410T sich als bewegliches, festes oder transportables Gerät einsetzen lässt.

Der ESB-Generator-Demodulator, NF- und Tonfrequenzteile sind als Einsteckmoduln ausgeführt und mit Schnellverschlüssen befestigt, was schnellen Ausbau für Service ermöglicht. Das Gerät ist transistoriert; Röhren finden nur in den Treiber- und Endstufen Verwendung.

Als Zubehör gibt es u.a. einen schlagfesten Handapparat, eine Morsetaste und einen für Instrumentenbretteinbau geeigneten externen Lautsprecher. Sprachgesteuerte Sende-Empfangumschaltung bei Betrieb mit Lautsprecher wird durch eine Spezialschaltung im Empfänger ermöglicht, die automatisches Umschalten auf "Senden" verhindert, wenn die Lautsprecherabstrahlung vom Mikrofon aufgefangen wird.

EE 03 753 *für weitere Einzelheiten*

Vielfachmessstellenschreiber

Vertrieb: Radiatron, 7 Sheen Park, Richmond, Surrey

(Abbildung Seite 121)

Radiatron gibt bekannt, dass die Auswahl der Jacquet-Miniaturpotentiometer für mehrere Messstellen durch drei Modelle ergänzt wurde, die für 100 mm breite Streifen ausgelegt sind.

Ein Einkurvenschreiber nach dem Baukastenprinzip ist in einem oder zwei Gehäusen lieferbar; letztere Ausführung ist für Anwendungen bestimmt, in denen die Einbautiefe auf 30,5 cm beschränkt ist. Der Schreiber hat eine Messunsicherheit von 0,5 Prozent des Vollausschlags und ist sowohl für Messungen von 5 mV...100 V und 0,5 μA...100 mA, als auch für Einsatz mit Thermoelementen und Widerstandsthermometern aller Art geeignet. Der Streifenvorschub ist zwischen 10 und 3600 mm/h einstellbar.

Ausserdem gibt es einen 2-Kurvenschreiber in der Zwei-Gehäuse-Ausführung mit zwei gleichzeitig, aber unabhängig voneinander arbeitenden Messsystemen.

Der Vielfachmessstellenschreiber hat 2 bis 6 Kanäle und arbeitet nach dem Zeitmultiplexverfahren; er hat viele Anwendungsmöglichkeiten auf Schalttafeln und im Labor. Die Kanäle sind ohne Begrenzung für Spannungs-, Strom- und Temperaturmessungen zuteilbar. Eine breite Auswahl von Bereichkarten ist lieferbar. Wechselspannungsmessungen von 20 Hz...20 kHz können mit allen Modellen durchgeführt werden.

EE 03 754 *für weitere Einzelheiten*

FET-Tester

Vertrieb: Livingston Laboratories Ltd, Livingston House, Greycaine Road, North Watford, Hertfordshire

(Abbildung Seite 121)

Livingston Laboratories Ltd kann jetzt einen Transistor-Rauschtester liefern, der die Zuverlässigkeit von FETs und bipolaren Transistoren durch Messen ihrer Rauschzahl bei Hochfrequenzen prüft.

Das neue Quan-Tech-Modell 340 enthält eine Stromversorgung und Messschaltungen, deren Messfähigkeit durch HF-Einsteckmoduln zu erweitern ist.

Jeder Modul enthält eine Testfassung für Transistor, einen Basiswiderstandschalter und HF-Schaltungen. Standardmoduln sind für 1 MHz und 10,7 MHz lieferbar; sie können aber auch für andere Testfrequenzen bestellt werden.

Für jeden der einstellbaren Basiswiderstände von 50 Ω bis zu 5 kΩ ist die Rauschzahl auf dem Grossskaleninstrument in der Frontplatte direkt in dB ablesbar. Kollektorstrom und -spannung sind zwischen 10 μA und 30 mA bzw. 1 V und 60 V regelbar.

EE 03 755 *für weitere Einzelheiten*

Doppelspeisegerät

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Abbildung Seite 121)

Das Doppelspeisegerät S besteht aus zwei identischen Konstantstromquellen, die je 0...20 V und 0...0.5 A abgeben und in Serie oder parallel geschaltet werden können. Das Gerät wird durchweg mittels Siliziumelementen geregelt, was Zuverlässigkeit und kleine Abmessungen gewährleistet. Die Spannung wird mit einem 20stufigen Schalter und kontinuierlicher Regelung zwischen Stufen eingestellt. Jede Stromquelle ist mit einem getrennten, von 0...0,5 A auf 0...100 mA umschaltbaren Strommesser ausgerüstet. Der eingebaute Überlastungsschutz gestattet Überlastung oder Kurzschluss für unbegrenzte Zeit ohne Beschädigung oder übermässige Erwärmung. Nach Entfernung der Überlastung stellt sich die korrekte Spannung automatisch wieder ein.

EE 03 756 *für weitere Einzelheiten*

Direktanzeigende Dehnungsmessbrücke

Tecquipment Ltd, Clinton House, 2A Sherwood Rise, Nottingham

(Abbildung Seite 122)

Dieses an der Universität Cambridge entwickelte Instrument ist im Grunde genommen eine Wheatstonsche Brücke, in der die beiden Messzweige aus zwei umschaltbaren Analogtransformatoren in Serie mit einem Schleifdraht für den Abgleich bestehen.

Die externen Brückenzweige bestehen aus einem beliebigen der elf Dehnstreifenpaare, die angeschlossen werden können. Diese Anzahl lässt sich durch einen schaltbaren Erweiterungs-Steckzusatz auf 66 Paare erhöhen. Alle Messstreifen werden—unabhängig davon, ob sie ausgeschaltet sind oder nicht—durch einen 1700-Hz-Oszillator erregt, so dass man Stabilisierung nicht abzuwarten braucht.

Ein Drehschalter stellt den gewünschten Messstreifenkanal ein, und der Abgleich wird auf einer 13-cm-Oszillografenröhre angezeigt. Die Dekadenmessschalter werden erst so eingeregelt, dass auf dem Schirm eine horizontale Spur entsteht; bei Beanspruchung der Messstreifen wird die Spur abgelenkt und durch Betätigung der Messschalter in den Horizontalverlauf zurückgebracht. Die Differenz zwischen den bei Abgleich und Wiederabgleich abgelesenen Werten gibt die Dehnung in Mikrodehnung pro Zoll für einen Empfindlichkeitsfaktor von 2.

Die die Schirmanzeige steuernde Schaltung besteht aus einem Brückengleichrichter mit sehr hoher, regelbarer Verstärkung, dessen Ausgang die Spur in vertikaler Richtung ablenkt; die horizontale Ablenkung erfolgt direkt durch den Oszillator.

Periodische dynamische Dehnungen erscheinen auf dem Bildschirm als zwei gleichschenklige Dreiecke, die sich an ihren Scheitelpunkten berühren. Wenn die beiden schrägen Seiten nacheinander mit Hilfe der Dekadenschalter horizontal gemacht werden, ergibt die Differenz der Skaleneinstellung die Dehnungsamplitude.

Die Messunsicherheit der Brücke ist unveränderlich innerhalb 3 Mikrozoll je Zoll und unabhängig von Netzschwankungen. Der Messumfang ist von −10 000...+22 000 Mikrozoll je Zoll. Die Messbrücke ist so robust, dass es fast unmöglich ist, sie zu beschädigen.

EE 03 757 *für weitere Einzelheiten*

Programmierplatten

D.E.B. Electronics Ltd, New Barnes Hill, Cottonmill Lane, St. Albans, Hertfordshire

(Abbildung Seite 122)

D.E.B. Electronics hat ein neues, preiswertes Programmiersystem entwickelt. Es besteht aus parallelen Kupferstreifen, die rechtwinklig zueinander auf den entgegengesetzten Seiten eines 1,6 mm dicken Schichtstoffträgers angeordnet sind. In jedem Kreuzungspunkt der Streifen gibt es ein Loch, und diese Löcher sind im Abstand von 7,62 mm angeordnet. Die Kupferstreifen sind

0,076 mm dick und mit 0,005 mm Kupfer, auf Wunsch auch noch mit 0,0025 mm Hartgold plattiert.

Eine Einsteckausführung wird mit Federleiste geliefert, jedoch ist das System auch für Tafeleinbau zu haben.

Die Einsteckausführung wird durch Verbinden von Bahnen auf entgegengesetzten Seiten mit Hilfe von Kurzschlussstiften oder einer Bauelementefassung in den Löchern programmiert. Die Fassung kann die verschiedensten Bauelemente, wie z.B. Dioden, Widerstände, Kondensatoren usw., aufnehmen. Eine als Programmierstift ausgebildete Niederspannungs-Miniaturlampe ist in der Entwicklung. Stifte werden mit derselben Plattierung wie die Bahnen und mit farbigen Kappen geliefert.

Die Platten werden Kundenwünschen entsprechend gefertigt und mit Beschriftung, Gravierung und Sammelschienen geliefert.

EE 03 758 *für weitere Einzelheiten*

Reed-Drucktasten und -Kippschalter

Vertrieb: Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Abbildung Seite 122)

Mit Ausnahme der Kontakte, die als völlig geschlossene Miniatur-Reedkontakte horizontal auf vertikale Druckschaltplatten montiert sind, entspricht der Rotireed-Kippschalter RG dem Standardtyp der britischen Post oder Kellogschalter. Die Kontakte werden durch kleine Dauermagnete auf einer Verlängerung des Kipphebels betätigt. Der Schalter hat in der Mitte eine neutrale Position mit Schaltlagen nach oben und unten, in denen er jeweils mit zwei oder vier Ruhekontakten bestückt werden kann. Da sie nacheinander geschlossen werden, können sie—je nach Geschwindigkeit und Richtung der manuellen Betätigung—als spät oder früh öffnende (oder schliessende) Kontakte Anwendung finden. Die kleine Frontplatte (19 × 57 mm) kann entweder mit ein-aus oder 1-0-2 oder anderen Bezeichnungen graviert werden.

Der abgebildete Drucktastenschalter Rotireed RT ist in Schaltreihen von 1, 4, 6 und 10 Tasten als Standardausführung zu haben, im Sonderauftrag jedoch mit bis zu 25 Tasten. Diese können mit bis zu vier Arbeits- oder Ruhekontakten in jeder Kombination bestückt werden. Die Tasten sind entweder nach Druck selbstverriegelnd und durch weiteren Druck Freigegeben, mit Selbstverriegelung und Freigabe durch Auslöschtaste oder nichtverriegelnd erhältlich. Elektromagnetische Verriegelung und Freigabe gibt es ebenfalls. Die Tasten kommen in Standardfarben weiss oder schwarz und können mit Ziffern in der anderen Farbe beschriftet werden. Andere Farben oder Leuchttasten gibt es gegen Sonderauftrag. Wenn jede Taste mit vier Kontaktsätzen bestückt ist, wird ein binärkodierter Ausgang möglich. Die verwendeten Standardkontakte sind für Schalten von 15 W mit maximal 1 A bei 250 V bemessen. Die Frontplatte der 10-Tasten-Schalterreihe hat Abmessungen von 213 × 22,3 mm; alle RT-Modelle sind 146 mm tief.

EE 03 759 *für weitere Einzelheiten*

Digitalvoltmeter

Sapphire Research and Electronics Ltd, Sapphire Works, Ferndale, Glam.

(Abbildung Seite 122)

Das mit neuartigen Halbleiterschaltungen und einem Wechselstrom-Servosystem ausgerüstete Modell 250 der Sapphire Research and Electronics Ltd misst Gleichspannungen bis zu 1000 V und Wechselspannungen bis zu 500 V_{eff} sowie Gleich- und Wechselströme.

Es kann Spannungsunterschiede von nur 200 μV unterscheiden und hat einen Frequenzbereich von 10 Hz...60 kHz. Die Messunsicherheit ist für Gleichspannungsbereiche 0,1 Prozent des Endwertes, für Wechselspannungen 0,25 Prozent. Die Eingangsimpedanz des Instrumentes ist hoch: für den 1-V-Bereich 1 GΩ, für alle anderen Spannungsbereiche 2,2 MΩ. Die Nulleinstellung des neuen Gerätes ist sehr konstant und erfordert kaum Wartung seitens der Bedienung. Die Anzeige ist unter allen Beleuchtungsverhältnissen klar und hell, die Kommaposition für alle Bereiche automatisch. Überspannungsschutz für alle Bereiche ist vorgesehen.

Das für Einsatz in Labor und Industrie gedachte Instrument ist bei ausgezeichneter Qualität, Tragbarkeit und niedrigem Preis sehr leistungsfähig.

EE 03 760 *für weitere Einzelheiten*

Mehrkanalschreiber

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Abbildung Seite 122)

S.E. Laboratories (Engineering) Ltd hat einen preisgünstigen und leichten Direktschreiber SE3006 eingeführt. Rationalisierung durch Wertanalyse und einfache, modulare Einschubbauweise ergaben einen Kurvenschreiber, in dem Vorteile und Anpassungsmöglichkeiten kombiniert sind, die man nur von viel teureren Geräten erwartet.

Der Direktschreiber wird entweder für 6 oder 12 Kanäle geliefert, ist mit englischen Standardgalvanometern bestückt und schreibt auf 127 mm breitem lichtempfindlichen Papier. Das Gerät ist tragbar, 178 mm hoch, wiegt 19 kg und ist für den Labortisch und Gestelleinbau geeignet.

Trotz Bestückung mit einer UV-Quecksilberdampflampe, die hohe Schreibgeschwindigkeiten ermöglicht, hat das Gerät nur eine Leistungsaufnahme von 250 VA. Papiervorschübe lassen sich im Verhältnis 1500:1 zwischen 5 mm/min und 1250 mm/s durch 8stufige Drucktastenwahl selbst während des Laufens einstellen. Der Registrierstreifen ist mit über die ganze Breite laufenden Zeitmarkierungslinien in 0,01 s, 0,1 s, 1 s, 0,1 min und 1 min Abstand, sowie 2-mm-Rasterlinien bedruckt. Regler für die Spur- und Rasterlinienstärke sowie Fernsteuerungseinrichtungen sind vorhanden.

Papier wird einfach durch Laden der Kassette mit einer Papierrolle eingelegt, da die Einführung automatisch erfolgt.

EE 03 761 *für weitere Einzelheiten*

Echte Effektivwertmessgeräte

Vertrieb: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Abbildung Seite 123)

Die Greibach-Messgeräte "Transquare" sind für das genaue Messen echter Effektivwerte komplexer Wellenformen bestimmt und eignen sich besonders für Arbeiten mit Thyristoren, z.B. für Präzisionsmessung der Eingangsleistung thyristorgesteuerter Heizeinrichtungen. Herkömmliche elektronische Messgeräte für echte Effektivwerte haben Messunsicherheiten in der Grössenordnung von 2...3 %, während die "Transquare"-Messgeräte zur Güteklasse 0,5 gehören.

Die Bereiche für echte Effektivspannungswerte haben Endwerte von 30 mV bis zu 1 kV. Die Kapazität für Strommessungen erstreckt sich bis auf 200 A mit dem geschlossenen Gerät und bis 1 kA mit Stromwandler oder Nebenschluss (für Allstrommessungen). Bisher war die höchste Empfindlichkeit 0,1 mA Vollausschlag, jetzt hat Greibach aber einen Mikrostrommesser mit 50 μA Bereichendwert entwickelt.

Typisch für diese hochempfindlichen Modelle ist das 500T005-8 mit acht Bereichen von 0,05 mA bis zu 100 mA. Das Gerät hat eine Messunsicherheit von 0,5 Prozent des Vollausschlags für unregelmässige Wellenformen mit Scheitelfaktoren über 4 und für Rechteckwellen bis zu mehreren kHz. Da weder Verstärker, noch Stromversorgungen oder andere aktive Elemente Verwendung finden, sind ungewöhnliche Zuverlässigkeit und eine Wiederholbarkeit innerhalb 0,1 Prozent gewährleistet. Es kann in den empfindlichsten Strombereichen 100mal überlastet werden und spricht auf Eingangsfrequenzen von 0 und 5...500 kHz mit einer Genauigkeit von ±0,5 Prozent des Vollausschlags an.

EE 03 762 *für weitere Einzelheiten*

Gleichstrommessbrücke und Verstärker

Vibro-Meter SA., Haletop Civic Centre, Wythenshawe, Manchester, 22

(Abbildung Seite 123)

Die Schweizer Messgerätefirma Vibro-Meter SA hat vor kurzem ein volltransistoriertes Brückenspeisegerät mit Abgleich- und Eicheinrichtung herausgebracht. Das Modell SPB-1/A ist für Verwendung mit Dehnmessstreifen von 60...600 Ω Widerstand und Lastzellen, Beschleunigungsmessern und Druckgebern auf Dehnmessstreifenbasis ausgelegt. Auf Wunsch ist auch ein passender Breitband - Gleichspannungsverstärker DCA-1/A für Signalverstärkung lieferbar.

Das Speisegerät SPB-1/A arbeitet mit einer vollständig erdfreien Brückenschaltung, die für Voll- oder Halbbrükkensysteme geeignet ist, und hat eine Mindestisolation gegen Erde von 10^9 Ω. Messungen können nach dem Ausschlag- oder Kompensationsverfahren durchgeführt werden, und der Brückenabgleich soll genau wiederholbar sein.

Das Gerät enthält ein Abgleichnetzwerk, das Kompensation der Nullpunkttoleranz des Wandlers und der Tara bis zu 300 Prozent des Nennwertes ermöglicht. Die Messung wird im Kompensationsverfahren beendet. Eine eingebaute Eicheinrichtung kann positive oder negative Signale proportional zu Mikrodehnung abgeben.

Der zusätzliche Vibro-Meter-Verstärker DCA-1/A hat eine Nullpunktdrift von nur 3 μV/°C. Das Gerät ist volltransistoriert und im Grunde genommen ein direktgekoppelter, hochkonstanter Gleichspannungsverstärker mit Differenz-Eingang und zwischen 3 und 1000 regelbarer Verstärkung. Einrichtungen für Nullpunkt-Abgleich und Gleichtaktunterdrückung sind vorhanden. Lieferbar mit Ausgangsleistung von ±10 V, ±2 mA und ±10 V, ±100 mA.

EE 03 763 *für weitere Einzelheiten*

Automatischer Taupunktschreiber

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Abbildung Seite 123)

Der neue automatische Taupunktschreiber von Shaw gestattet jetzt automatisches Registrieren des Taupunktes von Luft und Gasen. Das neue Gerät ist völlig in Festkörpertechnik ausgeführt, und der Schreiber erstellt Tages-, Wochen- oder Monatskurvenblätter. Absolute Anzeige des Taupunktes wird mittels eines Spiegels gegeben, der durch den Peltier-Effekt gekühlt wird; die Taubildung wird mittels einer Fotozelle ermittelt und die Spiegeltemperatur auf einer 127-mm-Instrumentskala angezeigt wie auch registriert.

Der automatische Taupunktschreiber zeigt positiv an, dass der wahre Taupunkt gemessen wird, und zwar in Form einer Lampe, die einmal je Sekunde aufleuchtet, wodurch nachgewiesen wird, dass auf dem Spiegel Tau vorhanden oder nicht vorhanden ist.

Standard-Taupunktbereiche sind −60° C...−20° C und −20° C...+20° C; im Sonderauftrag sind aber beliebige Bereiche lieferbar. Der Schreiber hat keine nachzustellenden Bedienorgane und ausser der Netzleitung keine externen Hilfsgeräte.

Ein Modell für meteorologische Zwecke und für Verwendung bei Raumtemperatur wurde auch entwickelt.

EE 03 764 *für weitere Einzelheiten*

Digitalzähler

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Abbildung Seite 124)

Ein neuer elektronischer Zähler hat trotz seines niedrigen Preises den grossen Frequenzbereich und Frontplatten-Einschubmöglichkeiten teurer Zähler, da gewisse Verfeinerungen weggelassen wurden. In der Grundausführung hat das Gerät eine Höchstzählrate von 50 MHz und kann mit Hilfe der greifbaren HP-Überlagerungsumsetzer Frequenzen bis zu 12,4 GHz mit 1 Hz Auflösung messen. Der Zähler hat trotz der preisgünstigen Ausführung die volle Messzeitenauswahl von 1 μs bis zu 1 s.

Dieser neue HP-Zähler 5246L kann alle für den 50-MHz-Zähler 5245L entwickelten Einschübe aufnehmen. Ausser Messungen bis zu 12,4 GHz mit dem HP-Einschub-Umsetzer kann dieser neue preisgünstige Zähler mittels Frequenzteiler-Einschub Frequenzen bis zu 350 MHz direkt messen, bei 100 mV Empfindlichkeit des Grundmodells über einen Einschub-Vorverstärker mit 1 mV Empfindlichkeit arbeiten, Anzeigen mittels eines Einschub-Vorwahlzählers normieren, oder mit zwei anderen Einschüben Zeitintervalle von 1 μs...10^8 s messen oder digitale Spannungsmessungen durchführen.

Der neue Zähler hat eine 6stellige Anzeige, die entweder bei Bestellung oder durch Nachrüstung zur Erzielung besserer Auflösung auf 7 oder 8 Stellen erweitert werden kann. Die Grundunsicherheit des Gerätes wird durch den Zeitbasisoszillator bestimmt, dessen Alterungsrate unter $\pm 2 \times 10^{-7}$ pro Monat liegt. Auf Wunsch ist eine hochkonstante Zeitbasis mit einer Alterungsrate von unter $\pm 3 \times 10^{-9}$ in 24 Stunden und mit der ausgezeichneten Kurzzeitkonstanz von $\pm 2 \times 10^{-10}$ mit einer Mittelwertzeit von 1 s lieferbar. Der neue Zähler kann auch mit einem externen Frequenznormal arbeiten und Frequenzverhältnisse messen.

Die Impedanz des Doppel-FET-Verstärkereingangs ist 1 MΩ parallel zu 25 pF für alle Signalpegel bis zu 1 V und frequenzunabhängig bis zu 50 MHz. Mit konstanter hoher Eingangsimpedanz ist ein niedriges Stehwellenverhältnis gewährleistet, wenn der Eingang mit einer abgeschlossenen Leitung verbunden wird. Für Spannungsabschwächung oder einen Eingang mit 10 MΩ, 10 pF kann ein Messkopf des Oszillografentyps benutzt werden.

Gegen Zuschlag ist der neue Zähler auch mit binärkodiertem Dezimalausgang in positivem '1'-Code 1-2-2-4 oder positivem oder negativem '1'-Code 1-2-4-8 zu haben.

EE 03 765 *für weitere Einzelheiten*

Überschall-Härtemesser

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 124)

Ein neues Messgerät für die Präzisions - Mikrohärteprüfung von Materialien wird jetzt von Dawe Instruments Ltd unter der Bezeichnung Sonodur-Mikrohärtemesser 1851 hergestellt. Der Messumfang ist 20...70 Rockwell C mit einer Messunsicherheit von ±1 Rockwell C.

In der Prüfung wird der Prüfling auf den Tisch des Unterteils gelegt. Der säulengeführte, mit einem Vickers-Diamanten bestückte Eindringkopf gleitet dann unter seiner Schwerkraft auf den Prüfling herunter und wird mit einer statischen Last von nur 150 g eingedrückt. Ein im Unterteil eingebauter Überschallgenerator erregt dann den Eindringkopf mit seiner Eigenfrequenz, die sich mit der Eindringtiefe der Spitze in den Prüfling ändert. Die Frequenzänderung wird elektronisch gemessen und auf einem ebenfalls im Unterteil eingebauten Messgerät angezeigt. Die Skala des Messgerätes ist direkt in Härtegraden geeicht.

Das neue Modell unterscheidet sich insofern von dem im vorigen Jahr eingeführten Modell Sonodur 1850A, als der Eindringkopf nicht in der Hand gehalten wird, sondern auf einer Säule gleitet. Die gleichförmigere Belastung gewährleistet höhere Messgenauigkeit. Das Modell 1850 bleibt weiter im Fertigungsprogramm, z.T. für den Aussendienst, aber auch für Werkstücke, die für Modell 1851 zu gross sind.

Da die Anzeige erfolgt, während der Kopf mit dem Prüfling in Berührung ist, werden durch elastische Nachwirkung verursachte Fehler vermieden. Jede Prüfung dauert weniger als 10 Sekunden und erfordert weder Fachkenntnisse noch Schulung. Die Eindringtiefe ist selbst bei völlig weichgeglühtem Stahl geringer als 0,005 mm, und die Eindruckfläche ist nicht einmal 0,1 Prozent der Eindruckfläche im Rockwell-Tester.

Der Sonodur 1851 macht optische oder mechanische Messeinrichtungen überflüssig. Da es sich um eine elektrische Anzeige handelt, ist es möglich, jeden beliebigen Teil der Skala zu dehnen, um —vor allem bei Einsatz in der Fertigung —Ablesefehler zu vermeiden.

In der Standardausführung wird das Instrument für das Prüfen von Stahl (Elastizitätsmodul 29...30 Millionen) geeicht geliefert. Es kann jedoch für Prüfung anderer Metalle eingerichtet werden, wofür man das Instrument gegen zwei Proben bekannter Härte eicht. Für Vergleichsmessungen anderer Werkstoffe gibt es eine zusätzliche, willkürlich von 0...100 unterteilte Skala.

EE 03 766 *für weitere Einzelheiten*

Gunn-Effekt-Oszillator

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

Der erste kommerziell greifbare Gunn-Effekt-Oszillator für das X-band (7...12 GHz) wurde von Mullard Ltd angekündigt.

Der in den Laboratorien der Associated Semiconductor Manufacturers Ltd in Wembley und den Mullard-Forschungslaboratorien entwickelte Baustein hat Dauerstrichleistung von 5 mW und wird aus einer einfachen 6-V-Gleichstromversorgung gespeist. Er ist nur 5 mm lang und hat 3 mm Durchmesser.

Mit seinen äusserst kleinen Abmessungen und der einfachen Arbeitsweise ist er eine verlockende Alternative zum herkömmlichen Klystronröhren- oder Transistor-Varactor-Überlagerungsoszillator, der in den Mischstufen von Mikrowellen - Nachrichtenverbindungen und Radarsystemen Anwendung findet. Er kann auch als direkte Quelle von Mikrowellenenergie für Radarsysteme niedriger Leistung, z.B. für Geschwindigkeitskontrolle im Verkehr und Tisch-Demonstrationsmodelle für Schulen und technische Lehranstalten, eingesetzt werden.

In Betrieb wird der Baustein Entwicklungstyp 106CXY in einen einfachen Mikrowellenhohlraum gekoppelt, dessen Resonanzfrequenz die Ausgangsfrequenz bestimmt. Der Hohlraum kann entweder eine Festfrequenz haben oder von 7... 12 GHz durchstimmbar sein.

Bisherige Gunn - Effekt - Oszillatoren wurden durch Erzeugung eines geeigneten Feldes zwischen den beiden Seiten eines dünnen Scheibchens von Galliumarsenid (GaAs) erzeugt—je dünner die Scheibe, desto höher die Betriebsfrequenz. Bisher wurden diese Scheiben durch Schleifen und Läppen erzeugt, was die Betriebsfrequenz auf maximal 4 GHz beschränkt.

Zur Lösung dieses Problems hatte Mullard ein epitaxiales Abscheidungsverfahren entwickelt, das Wachsen einer sehr dünnen GaAs-Schicht auf einer dickeren Scheibe desselben Materials ermöglicht. Ein auf diese Weise hergestelltes Element ist auf der Oberfläche in einer Oberflächenschicht hohen spezifischen Widerstandes auf einem Substrat niedrigen spezifischen Widerstandes aktiv, und die Aktion geht in dieser Schicht vor sich.

Ausser der Erhöhung der Betriebsfrequenz gestattet das Verfahren bessere Kontrolle des Widerstandes für eine Schicht gegebenen spezifischen Widerstandes und höhere Belastbarkeit.

EE 03 767 *für weitere Einzelheiten*

Mikrowellenschalter

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

(Abbildung Seite 124)

Microwave Associates gibt die Entwicklung eines neuen koaxialen, einpoligen Diodenumschalters mit 60 dB Trennung, einem Stehwellenverhältnis von unter 1,07 und einer Einfügungsdämpfung von weniger als 0,5 dB über eine 20-Prozent-Bandbreite im Frequenzbereich 1...2 GHz bekannt.

Für Bandbreiten bis zu 50 Prozent bietet dieser Schalter—Typ MA8032-2L4S—mehr als 40 dB Trennung, ein Stehwellenverhältnis unter 1,20 und eine Einfügungsdämpfung von weniger als 0,75 dB.

Der in erster Linie für automatische Mikrowellen-Testausrüstungen in Verbindung mit Phasen-Messbrücken entwikkelte Schalter ist auch für Anwendungen ideal, bei denen unter Strom geschaltet wird, z.B. wo zwei Sender an einem gemeinsamen Ausgang liegen oder beim Umschalten zwischen zwei Antennen.

EE 03 768 *für weitere Einzelheiten*

Stromüberwachung

Scama Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(Abbildung Seite 125)

Ein neues Stromüberwachungsgerät für Steuer- oder Alarmzwecke wird von Scama Ltd, Hersteller des Panalarm-Systems, angekündigt. Der "Trip-Matic"-Monitor Baureihe 70 kann entweder einzeln in Kontrollschränken oder Rückwandmontage Verwendung finden, oder als integraler Teil des Panalarm-Systems Modell 75, um von analogen Eingängen direkte Alarmsignale zu geben.

In einer typischen Anwendung überwacht der Trip-Matic ein dem Betriebsparameter proportionales Gleichstromsignal. Der Gleichstromsignalpegel steigt mit wachsenden Parameterwerten. Wenn das Signal den vorgewählten, regelbaren Bezugswert gerade überschreitet, tritt eine scharfe Änderung im Ausgang eines Vergleicherverstärkers auf, die ein Alarmsignal oder andere Kontrolleinrichtungen erregt.

Der neue Signalmonitor soll ungewöhnlich genau sein und äusserst schnell ansprechen. Er ist für Gleichstromsignale ausgelegt, kann jedoch vielen Wechselstromanwendugen angepasst werden. Er ist volltransistoriert und durchweg mit Silizium-Halbleitern bestückt. Alle Eingänge sind voneinander und von den Ausgängen isoliert. Die Differenz zwischen Auslösung und Rückstellung kann entweder fest oder regelbar sein, und die Auslösung kann entsprechend den Anwendungsbedingungen entweder bei hohem oder niedrigem Pegel erfolgen.

EE 03 769 *für weitere Einzelheiten*

Differenzvoltmeter

Vertrieb: General Test Instruments Ltd, Gloucester Trading Estate, Hucclecote, Gloucester

(Abbildung Seite 125)

Das von Wavetek hergestellte neue Voltmeter Dialamatic erreicht trotz Handbedienung fast die Geschwindigkeit automatischer Voltmeter, bietet jedoch die Messgenauigkeit, robuste Zuverlässigkeit und Preis herkömmlicher Differenzmodelle.

Das Herz dieser Leistung ist der Transfermatic-Schalter, der die Dekadenschalter mechanisch so kuppelt, dass der "Übertrag" automatisch auf die nächsthöhere Dekade übertragen wird. Wenn sich die Anzeige einem Hauptübertragungspunkt, z.B. 2,9999, nähert, wird sie durch Schalten des rechten Bedienungsknopfes automatisch auf 3,000 geändert. Dieser Vorgang ist auch in umgekehrter Richtung möglich.

Mit einem herkömmlichen Differenzvoltmeter müsste jeder Knopf einzeln bedient werden.

Das Dialamatic-Modell ist ausserdem mit einem schnellen nichtsättigenden Nullabgleichverstärker ausgerüstet, so dass der Benutzer ohne Änderung der Nullempfindlichkeit den richtigen Abgleich einstellen kann.

Das neue Voltmeter wurde nicht nur für schnelle Arbeitsweise, sondern auch für schnelle und einfache Wartung entwickelt. Jede Schaltung ist auf einer modularen Steckkarte untergebracht, die für gegebenenfalls erforderlichen Service aus dem Gehäuse herausgeschwenkt wird. Das Gerät ist für hohe Zuverlässigkeit unter extremen Umgebungsbedingungen ausschliesslich mit Silizium-Halbleitern bestückt.

Das Voltmeter Dialamatic kommt in zwei Grundausführungen: Modell 201 misst Gleichspannungen von 0...1000 V in vier Bereichen, Modell 202 Gleich- und Wechselspannungen von 1 mV... 1000V. Beide Modelle sind für Batteriebetrieb mit eingebautem Lader lieferbar.

EE 03 770 *für weitere Einzelheiten*

Veränderliche Präzisionswiderstände

Steatite Insulations Ltd, Hagley House, Hagley Road, Birmingham 16

Die als VariZistor angebotenen, neuen Präzisionswiderstände verbinden die Funktion eines Präzisionswiderstandes mit der Einstellbarkeit eines Trimmerpotentiometers in einem Bauelement, das unendliche Auflösungsmöglichkeit gibt. Äusserst genaue Widerstandswerte sind leicht erreichbar, so dass der Widerstand auf die Anforderungen der Schaltung zugeschnitten werden kann.

Das Bauelement besteht aus einem auf ein Keramiksubstrat abgeschiedenen Dünnfilm-Widerstandselement in einem rechteckigen Gehäuse. Ein Vielfachkontaktschleifer ist an dem Anschlussdraht befestigt und an einer Seite isoliert. Der Draht läuft durch den Körper und gestattet Kontakt zwischen der Klammer und dem Widerstandselement, wodurch der Stromschluss der Schaltung hergestellt wird. Der ganze Körper gleitet innerhalb fester Grenzen, wodurch der Schleifer über das Widerstandselement geführt wird. Etwa 20 Prozent Widerstandsänderung über und unter dem Nennwert ist für Nennwerte von 0,5 Ω bis zu 1 MΩ mit 0,01 Prozent Einstellunsicherheit möglich.

Dem Schaltungsingenieur wird hier die Möglichkeit geboten, mit Schaltungselementen ohne Auftreten der Festwiderständen eigenen Probleme zu experimentieren. In empfindlichen Schaltungen können in der Endprüfung nach Einlöten der Anschlüsse sehr feine Widerstandsänderungen vorgenommen werden, um für Elementtoleranzen und Streuwiderstand zu kompensieren.

VariZistors sind so konstruiert, dass sie den Anforderungen aller hohe Zuverlässigkeit vorschreibenden Pflichtenblätter genügen. Anschlüsse können in jede Schaltung gelötet oder geschweisst werden. Genaue Widerstandswerte lassen sich durch Verlöten des Anschlussdrahtes mit dem Körper fest einstellen. Man kann Endkappen anlöten, um den Körper hermetisch abzudichten. VariZistors sind für Nennlasten von ¼, ½ und 1 W lieferbar, und Spezialtemperaturkoeffizienten gibt es als Standard bis zu $5 \times 10^{-6}/°C$ herunter.

EE 03 771 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Entwurf eines VHF-Festkörper-Endverstärkers von T. Spencer

Zusammenfassung des Beitrages auf Seite 72-77

Eine auf von Element-Herstellern veröffentlichten Standarddaten beruhende Entwurfstechnik für C-Endverstärker wird gegeben. Die Analyse für in Klasse C betriebene Transistoren ist mit der für Röhren angewendeten identisch, und normierte Entwurfskurven werden behandelt. Es wird gezeigt, dass der Vergleich von Entwurfsberechnungen nach dem schrittweisen Verfahren mit den praktischen Ergebnissen gut übereinstimmt.

Verschiedene Endverstärker- und Multipliziersschaltungen werden zusammen mit solchen für freilaufende und kristallgesteuerte Oszillatortreiber erläutert. Die Wirkungsgrad und Ausgangsleistung beeinflussenden Faktoren werden besprochen.

Nachbildung eines Konfusionskegels von S. Dalal

Zusammenfassung des Beitrages auf Seite 78-81

Beim Entwurf der Funknavigationshilfennachbildung für einen Flugsimulator müssen sich Ingenieure oft mit dem Problem der Nachbildung des über einem Funkfeuer bestehenden Konfusionskegels befassen. Das beschriebene System ist ein verhältnismässig billiges, volltransistoriertes und integriertes System zur Nachbildung eines realistischen Konfusionskegels.

Ein quadratischer Detektor für Leistungsmessungen von J. F. Golding

Zusammenfassung des Beitrages auf Seite 82-86

In diesem Beitrag wird eine Detektorschaltungsanordnung beschrieben, die ursprünglich für eine der mittleren Ausgangsleistung eines Empfängers proportionale Oszillografenablenkung entwickelt wurde. Das Prinzip hat jedoch auch andere Anwendungsmöglichkeiten, einschliesslich Benutzung der Schaltung als aperiodischer Frequenzverdoppler.

Ein einfaches frequenzmoduliertes Magnetbandgerätsystem von W. Tempest und M. E. Bryan

Zusammenfassung des Beitrages auf Seite 87-89

Ein einfaches, transistoriertes Frequenzmodulationsgerät, das mit jedem guten Heim- oder halbprofessionellen Tonbandgerät zur Aufzeichnung und Wiedergabe von Signalen im Frequenzbereich 0 . . . 1250 Hz Anwendung finden kann, wird beschrieben. Das Gerät hat einen Störabstand von 42 dB und ist so konstant, dass ein vom System wiedergegebenes Wechselstromsignal in 20 Minuten nicht mehr als 0,1 dB, ein Gleichstrompegel nicht mehr als etwa 5 mV wandern wird. Der Eingangssignalbereich ist ±5 V, der Ausgang bei ±0,5 V etwa 20 dB niedriger.

Einstellungen zum Entwurf von Stromversorgungen für Photovervielfacher von R. Barlow und J. Watson

Zusammenfassung des Beitrages auf Seite 90-96

Der Leistungsbedarf von Photovervielfachern wird besprochen und gezeigt, dass die zugehörigen Schaltungen und die Arbeitsweise der Photovervielfacher manchmal Lockerung des Pflichtenblattes für die diesbezüglichen Stromversorgungen gestatten. Der Entwurf von Hochspannungs-Wechselrichtertransformatoren und der zugehörigen Konstanthalterschaltungen wird in Betracht gezogen und mit Anwendung einer FET-Vergleichsstufe erreichbare Vorteile aufgezeigt.

Vereinfachter Entwurf des Gegenkopplungspaares von J. C. S. Richards

Zusammenfassung des Beitrages auf Seite 97-99

Der Entwurf einer Wechselstromverstärkerstufe mit zwei Transistoren und Parallelgegenkopplung wird auf die Anwendung einiger sehr einfacher Formeln reduziert, die sich auf Gleichstromvorspannung, Verstärkung und Bandbreite beziehen. Die erforderlichen Annäherungen und Annahmen sind für die meisten Silizium-Planartransistoren für niedrige Leistung gültig; die erzielte Leistung ist für viele Zwecke dem Optimum nahe genug.

Wirtschaftlichkeit und Zuverlässigkeit im Entwurf kerntechnischer Instrumente von B. Shepherd

Zusammenfassung des Beitrages auf Seite 100-105

In der Entwicklung der in diesem Beitrag beschriebenen Instrumente wurde zur Erzeugung zuverlässiger und preisgünstiger Messgeräte Wertanalyse angewendet. Neue Schaltungen fanden im Entwurf schneller Dekadenkarten mit Sichtanzeige, eines konstanten Hochspannungstransformators sowie von Tunneldioden-Impulshöhendiskriminatoren Anwendung.

Ein Digitalinstrument zum Messen der Schlüpfung von Asynchronmotoren von V. Böhm, V. Popescu und T. Teitel

Zusammenfassung des Beitrages auf Seite 106-109

Der Beitrag beschreibt ein volltransistoriertes Digitalinstrument, das die Schlüpfung von Induktionsmotoren direkt in Dezimalform anzeigt, was durch Kopplung eines fotoelektrischen Abtasters mit einer auf der Motorenwelle aufgebrachten Scheibe erreicht wird. Das auf der Impulszähltechnik beruhende Verfahren ist von den anderen Messverfahren für diesen Parameter eigenen Fehlern völlig frei.

Hochfrequenz-Multiplikation mit logarithmischen Dioden von R. P. Ingram

Zusammenfassung des Beitrages auf Seite 110-112

Eine Methode zum Multiplizieren bei hohen Frequenzen mit einem Brücken-Ringmodulator aus logarithmischen Dioden wird beschrieben. Aus der Theorie lässt sich schliessen, dass unter Voraussetzung gewisser Aussteuerungs- und Ausgangszustände eine sehr gute Annäherung an die echten Multiplikationsergebnisse möglich ist. Solche Multiplizierschaltungen können ohne grosse Schwierigkeiten für Betrieb bis zu 100 kHz konstruiert werden, und bei geeigneter Auswahl der Dioden lassen sich auch bei höheren Frequenzen ziemlich gute Resultate erreichen. Schaltungen dieser Art könnten in der Korrelationsanalyse und in Abtastsystemen nützlich sein.

Eine neuartige Bezugsspannungsquelle von T. P. Dawson und G. C. Taylor

Zusammenfassung des Beitrages auf Seite 113

Eine mit Halbleitern bestückte Spannungsstabilisierungsschaltung in Brückenanordnung gibt einen Stabilisierungsfaktor von über 700. In üblichen Konstanthalterschaltungen mit zwei Zenerdioden ist der Faktor nur etwa 25.

Eine sprachgesteuerte Uhr von K. J. Kapota

Zusammenfassung des Beitrages auf Seite 114

Eine elektrische Uhr wird mit Hilfe eines Kehlkopfmikrofons und elektronischer Schaltungen ein- und abgeschaltet, so dass gemessen werden kann, wie lange die Person spricht. Die Einrichtung ist für Anwendung in der Psychologie und Psychiatrie bestimmt.