# Wireless World

Helical v.h.f. aerial

Audio oscillator

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### Electronics, Television, Radio, Audio

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September 1971  17½p

Our **cover photograph** is of part of the vibrations and sound section of Evoluon, the permanent exhibition at Philips, Eindhoven. In this abstract presentation sounds are converted into electronic pulses, transmitted and reconverted into sound. *Photographer Paul Brierley.*

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## IN OUR NEXT ISSUE

How a modified f.m. tuner used in conjunction with a simple oscilloscope and a home-made aerial will receive weather pictures from satellites.

A review of television receiver techniques.

Making a turntable and pickup arm.

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# Wireless World

## The Plight of the Microcircuit Industry

During the past month there have been some dramatic moves made in the British microcircuit industry. First came the announcement that the G.E.C. proposed closing the Marconi-Elliott Microelectronics factories at Witham, Essex, (which was purpose built in 1968) and at Glenrothes, Fife. Within days of this proposal Motorola held a party in East Kilbride, Lanarkshire, to celebrate the start of work on the building of a new microcircuit factory! Then came a press release from Mackintosh Consultants Company, of Glenrothes, outlining the results of a survey of the British microelectronics industry they had undertaken on behalf of the Department of Trade and Industry and the National Research Development Corporation.

This report, which is confidential, although abridged copies have been made available to the companies who participated in the survey, expresses the views of the consultants and not necessarily those of the companies concerned nor the sponsors of the study. However, the brief details given in the press release must have sent a shudder down the spine of some British companies. The view is expressed that because of the dominance of European markets by American manufacturers, no single national market in Europe (and this applies equally to Britain) is capable of supporting even one major i.c. company and, moreover, no company can succeed in this industry without access to markets which are both large and innovative. When one looks at the production figures of the big five microcircuit companies in the States and compares them with the total output of all the indigenous European companies (including Philips) one finds that the aggregate is not half of any *one* of the major American i.c. companies. In face of competition from such giants what prospect is there for a *British* microcircuit industry or even a joint European enterprise. In spite of this, it has been announced by the Italian finance organization I.R.I. that it has taken over SGS and ATES and, reading between the lines it would appear that a national electronic components company—both passive and active—may emerge.

To get back to the British scene, there were, of course, the inevitable questions in the House of Commons and the letters in the Press condemning the proposed G.E.C. closures as, to use the words of Brian Harrison, Conservative M.P., it would be little short of a national tragedy if Government action was not taken to prevent the expertise associated with microcircuit production being lost to Britain. Similar sentiments were expressed in a letter in *The Times* from the general secretary of the Electrical, Electronic and Telecommunications Union who concluded with the words 'the Government must act now or the future of the British electronics industry must surely be at risk'.

Only a few days before these chilly winds blew through the industry Mr. John Davies, the Secretary for State, stated, in the course of a debate on the conditions in another industry, that the Government would not finance 'lame ducks'. It will, however, be recalled that grants to the tune of £5M were made by the National Research Development Corporation to three British-owned i.c. manufacturers (one was Marconi-Elliott) only three years ago. This money is recoverable by levy on the sales of i.cs.

What then are the long-term prospects for this country's microelectronics industry? In the present climate of internationalism is it reasonable to think in terms of national companies? American companies have for some time been setting up factories in Europe to be in E.F.T.A. and E.E.C. In the face of such competition and in view of our plans to join the Common Market would it not be in our interest and in the interest of our Continental partners to set up a strong joint European i.c. company to compete in the world market for mass-produced microcircuits? This need not mean the end of i.c. research and production in this country; there would, we believe, still be room for one or two British manufacturers of specialized microcircuits.

# Sweep-frequency Audio Oscillator

## Two-decade linear sweep using b.f.o. technique

*by R. J. Ward, B.A.*

For many linear circuits the characteristic of prime practical importance is their frequency-amplitude response. This class of circuits includes tuned circuits, equalizers, filters and selective amplifiers. The response is commonly measured by connecting a variable-frequency oscillator to the input of the circuit and a suitable output measuring meter. Measurements are then made at as many fixed frequencies as necessary and plotted to obtain an overall picture of the performance. Though simple and convenient this becomes tedious when many response graphs have to be measured, and is too slow when demonstrating the properties of such circuits to a class of students.

In such situations it is useful to display the graph of gain or loss against frequency directly and quickly. This article describes an instrument which used with an oscilloscope or X-Y plotter enables such a direct plot to be made.

The components needed to build this oscillator cost at least £20 which is no doubt more than the cost of the much simpler audio sweep generator recently described by F. H. Trist.* The main difference in performance is the sweep linearity—0.2% in this design over a 100:1 frequency range as opposed to 15% over a 10:1 range for the simpler design. This figure of 0.2% allows direct accurate plotting along the frequency axis and the use of calibrated controls for varying the sweep range.

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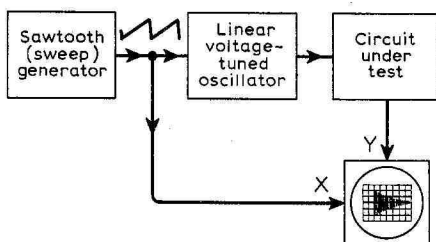* Audio Sweep Generator by F. H. Trist, *Wireless World*, vol. 77 July 1971 pp. 335-8.



*Fig. 1. In the swept frequency technique for obtaining the amplitude-frequency response of networks quickly, a sawtooth waveform controls a voltage-to-frequency converter and simultaneously sweeps the oscilloscope beam in the X-direction.*
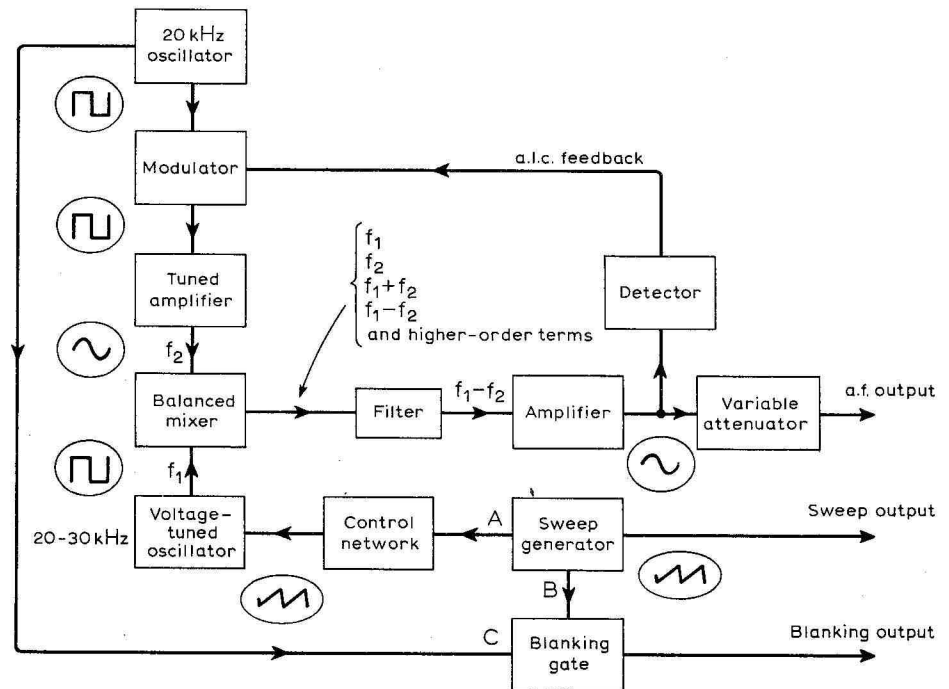


*Fig. 2. Using the beat frequency technique two decades can be covered without switching. As well as the $f_1 - f_2$ signal others are generated which must be filtered out.*

The present design is single range and therefore limited in total coverage. Other features such as output level and sweep-rate controls are not fundamental to the technique but come from aiming for maximum flexibility within one frequency range.

Automatic level control and waveform purity are similar in both designs except that F. H. Trist's generator should be completely free of spurious signals not harmonically related to the fundamental.

The technique used is illustrated in Fig. 1. The sawtooth voltage simultaneously tunes the oscillator and sweeps the beam across the screen, so that the sideways deflection is proportional to frequency—a 'frequency base'. The Y-deflection is proportional to the output amplitude, so assuming the oscillator output level remains constant a plot of gain against frequency is obtained.

In this instrument two decades of frequency are covered in one band using a beat-frequency technique—Fig. 2. Oscillations at frequencies $f_1$ and $f_2$ are fed into

a mixer which produces many frequencies at its output, principally the original ones together with the sum $(f_1 + f_2)$ and difference $(f_1 - f_2)$ frequencies. The difference frequency signal is selected by a filter. If the original frequencies are well above the difference frequency then the latter can be easily filtered from the mixer output, but if they are too high a small fractional change in either $f_1$ or $f_2$ will result in a large fractional change in $f_1 - f_2$, so frequency stability will be poor.

Frequency $f_2$ is fixed at 20kHz and $f_1$ variable from 20 to 30kHz giving an output at 0 to 10kHz. In practice the output waveform deteriorates when $f_1$ and $f_2$ are very nearly equal so the usable output range is 100Hz to 10kHz.

### Swept generator

The overall structure of the complete sweep oscillator is shown in Fig. 2. It is convenient to start its description with the sweep generator. Audio-frequency systems

## Prototype specification

**Frequency range:** 100Hz to 10kHz in one range, accurate to ±100Hz mid-scale or ±20Hz by calibrating potentiometer; ±20Hz at range ends.

**Amplitude:** adjustable up to 3V r.m.s. (open-circuit) in six ranges and accurate to ±5%. Level to 3% over range. Output impedance is 600Ω ±2%, except on highest ranges (±20% for 3-V range and −2%+ 7% for 1-V range).

**Spurious outputs** second harmonic 0.5% or −45dB. All other spurious signals at least 45dB below fundamental at 1kHz.

**Sweep times:** 50ms to 20s in four ranges.

**Sweep modes:** 'full'–100Hz-10kHz; 'wide'– 100Hz to frequency set on dial; 'symmetrical'–sweep widths of 30, 100, 300Hz, 1 or 3kHz ±3%, centre frequency set on dial; 'external'–sensitivity about 800Hz/V, $Z_{in}$=50kΩ.

**Other outputs:** sweep output +5V and −5V from 1kΩ. Blanking −20V pk-pk 20-kHz square wave for bright-up from less than 6kΩ.
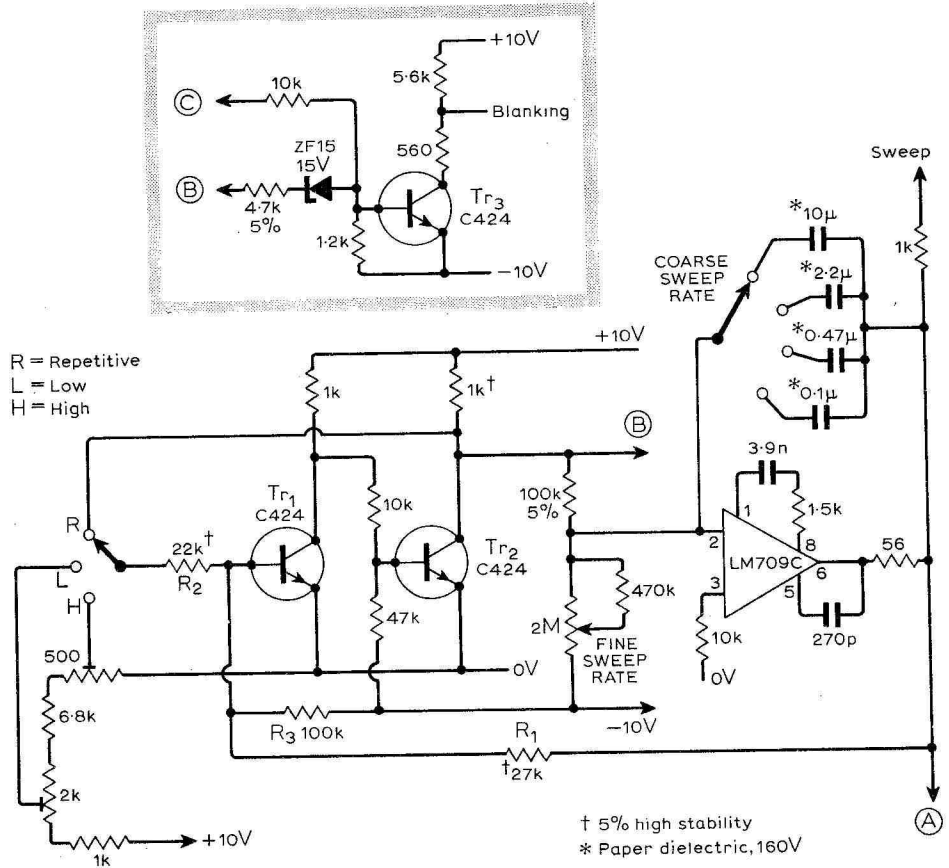


Fig. 3. Sweep generator circuit comprising bistable multivibrator and integrator achieves sweep times as low as 20s. Beam-blanking waveforms are shown in Fig. 4. In this and subsequent circuits the LM709C integrated circuits require +10V on lead 7 and −10V on lead 4. Resistors are $\frac{1}{4}$-watt, 10% tolerance unless shown otherwise.

can easily have resonances with a width of tens of Hz or less with corresponding response times longer than one tenth of a second; very slow sweep rates are needed if such detail is not to be lost. The circuit used to achieve sweep rates down to 20s per sweep is shown in Fig. 3.

Transistors $Tr_1$ and $Tr_2$ are coupled together as a bistable multivibrator with $R_1$, $R_2$ and $R_3$ chosen so that when the potential at point A is greater than +5V it switches over to the condition with $Tr_1$ saturated. When the potential at point A is more negative than −5V the circuit switches over to the condition with $Tr_2$ saturated. The operational amplifier is used as an integrating amplifier, the associated capacitor being selected by the coarse sweep-rate control.

When the R-L-H switch is in position R, the sweep starts with $Tr_2$ saturated and point A at −5V. Point B is then nearly at zero potential and current flows from the input of the integrator through the fine sweep-rate control to the negative supply rail, raising the potential at the output of the integrator. When this reaches +5V the circuit switches over leaving B at nearly the positive supply potential; current flows into the integrator and the potential at the output returns comparatively quickly to −5V. The cycle then repeats.

With the switch in position L the sweep generator takes up the stable non-oscillatory condition with A at near −5V, the precise value being set by the 2-kΩ potentiometer. In this condition $Tr_1$ and $Tr_2$ behave as a high-gain amplifier feeding any difference between the desired and actual potential at A to the input of the integrator in such a sense as to bring the potential at A back to the value desired. When the switch is turned to H, $Tr_1$ cuts off and $Tr_2$ saturates. The output of the integrator changes in precisely the same

manner as in position R and so does the output frequency. When the potential at A reaches +5V this time it stops changing because the circuit has reached a stable condition with the precise output potential set by the 500-Ω potentiometer. Similarly when the switch is turned back to L the voltage at A flies back to −5V. By manipulating the switch in this way a manual sweep is obtained which is useful for X-Y plotters. The ability to hold the potential of



Fig. 4. Beam blanking is achieved with an h.f. waveform because input to the c.r.t. grid is often a.c.-coupled. Waveform from $Tr_3$ is at (c), and (d) is at c.r.t. grid.

A at either end of the sweep range is needed to calibrate correctly.

The potential at B (Fig. 3) is such that a positive blanking pulse of nearly 10V is available during flyback. If this pulse were inverted and amplified to give a negative pulse this would be satisfactory for bright-up on oscilloscopes at the faster sweep rates. Unfortunately, the Z-modulation input of oscilloscopes is commonly a.c.-coupled to the grid of the c.r.t. so that this blanking is ineffective at slow sweep rates. To circumvent this a gated high-frequency oscillation is used.

Referring to Fig. 4, a 20-kHz square wave from the fixed-frequency oscillator is applied to $Tr_3$ together with the positive blanking pulse (b). Transistor $Tr_3$ and its associated components now behave as a NOR gate with output (c) which after a.c. coupling to the grid of the c.r.t. becomes (d). If the brightness control is adjusted so that the beam is just cut off with no Z-modulation applied, as shown, it will remain cut off during flyback but pulse on some 20,000 times per second during the sweep. Subjectively these dots merge together on the screen to give a normal display, except for some moiré fringing when displaying the envelope of a waveform whose frequency is close to a low-order sub-harmonic of 20kHz; this is not normally a nuisance.

### Voltage-tuned oscillator
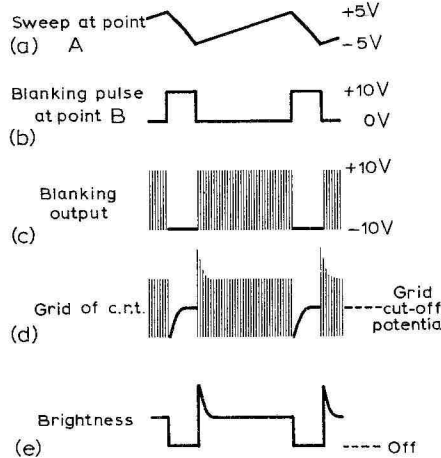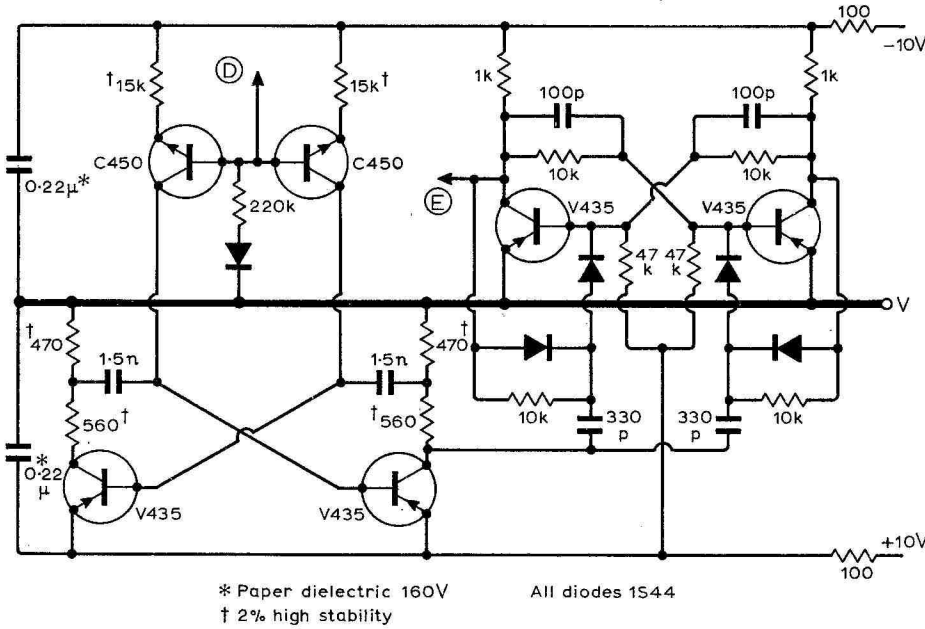
The voltage from the sweep generator

*Fig. 5. Astable multivibrator—with constant-current sources controlled by input voltage at D—acts as the voltage-tuned oscillator covering 40 to 60kHz. Bistable multivibrator provides anti-phase outputs with 1:1 mark:space ratio at E and F at 20 to 30kHz.*
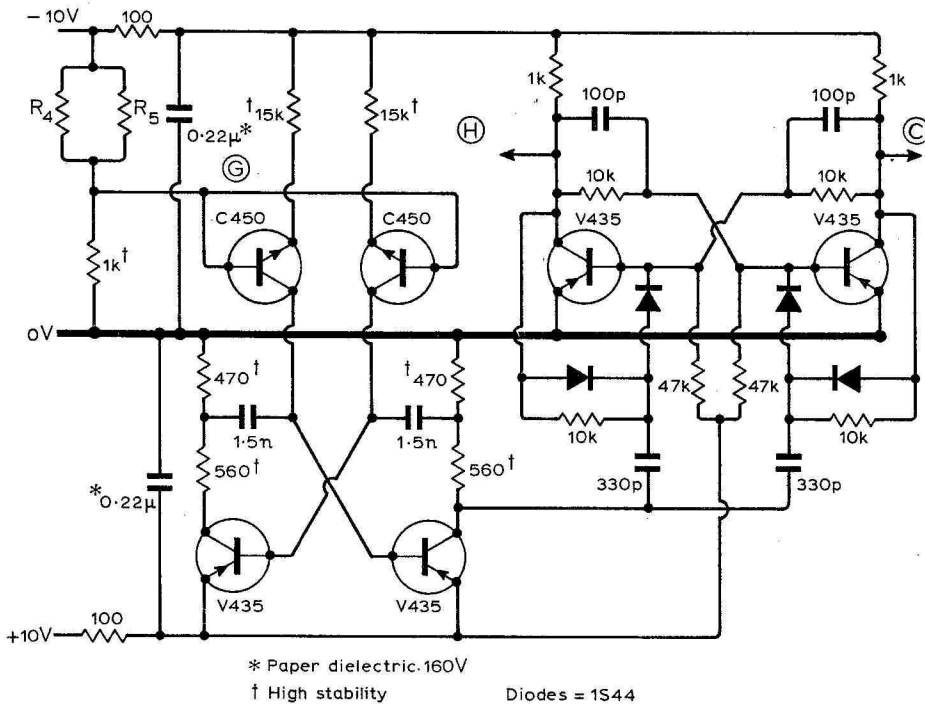


*Fig. 6. In the hope that frequency drifts in the fixed and variable-frequency oscillators will cancel, this fixed frequency oscillator is identical to Fig. 5.*

after passing through the control network (see later) is converted to the frequency of an oscillation by a voltage-tuned oscillator. The circuit is shown in Fig. 5 where the basic oscillator is followed by a conventional divide-by-two bistable multivibrator to provide two square waves, at E and F, in anti-phase and with unity mark/space ratio. The basic oscillator is an astable multivibrator with a pair of constant-current sources controlled by the input voltage at D.

The astable multivibrator tunes over the band 40 to 60kHz. This is divided by two and mixed with 20kHz to produce the 0 to 10kHz audio output. In practice the relation between input potential and output frequency gives a maximum deviation from a straight line of 70Hz–0.7% of full range at worst. This is reduced to 0.2% by adding a linearizing diode and resistor across the output.

## Fixed-frequency oscillator

The design of the fixed-frequency oscillator is identical with that of the voltage-tuned oscillator in the hope that frequency drifts in the two oscillators will be similar and to a large extent cancel. The nominal frequency at the point H, Fig. 6, is 20kHz. To achieve this within a tolerance of 200Hz (1%) the input voltage at G is obtained from a potentiometer across the supply lines. To find the values for $R_4$ and $R_5$, first connect a 10-k$\Omega$ potentiometer across the $\pm$10-V supply lines, and measure the required voltage. Calculate $R_4$ to give a voltage 10% lower than this, using a preferred value, and try values of $R_5$ around ten times $R_4$ until the frequency is as close as possible to 20kHz. The square wave for the blanking oscillator is obtained from the collector of $Tr_4$.

## Conversion to sine-wave

The 20-kHz output from the fixed-frequency oscillator at H drives $Tr_5$ as a

*(Below) Fig. 7. Amplitude of the fixed-frequency output is controlled by the shunt a.l.c. transistor and drive. Tuned filter reduces second harmonic to 0.05% of fundamental before applying signal to balanced mixer.*
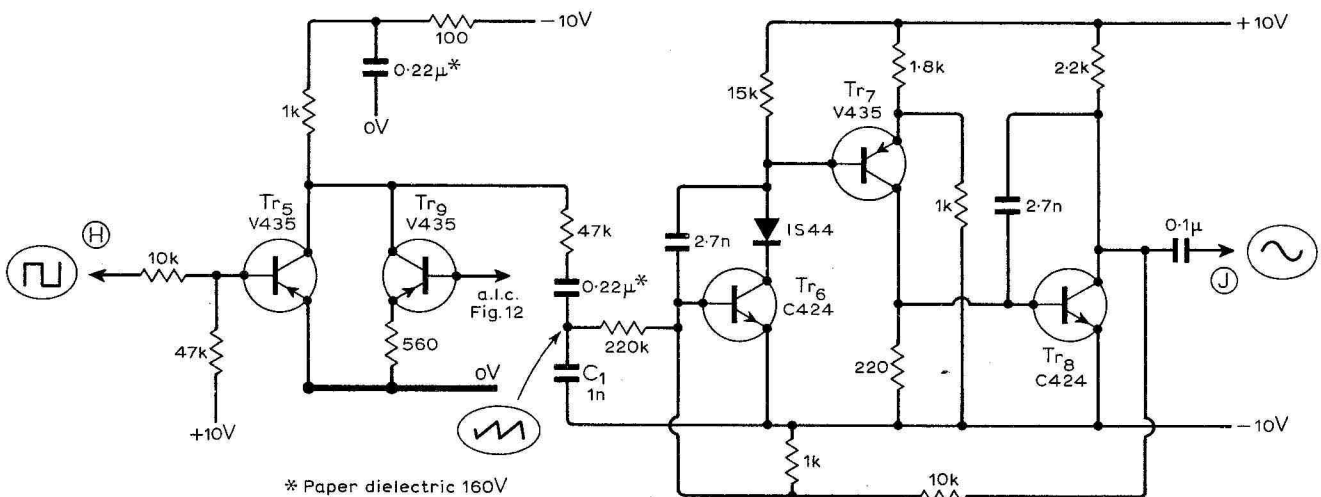
switch–Fig. 7–producing a square wave at its collector. Transistor $Tr_5$ is shunted by $Tr_9$ so that the amplitude of the square wave is controlled by the automatic-level-control current flowing from the base of $Tr_9$.

This square wave goes to the low-pass $RC$ filter consisting of $R_6$ and $C_1$ whose output is a sawtooth which is applied to the input of the tuned amplifier, containing $Tr_{6-8}$, due to Faulkner and Downe†. This amplifies the 20-kHz fundamental and attenuates harmonics. Amplitude of the second harmonic is 0.05% of the fundamental (0.8V); the amplitude of the other harmonics could not be measured directly but calculated values are 0.1% third harmonic and less than 0.02% fifth harmonic.

## Mixer

The mixer forms the product of this sine wave of frequency $f_2$ with the variable-frequency square wave. The output from the mixer can be expressed as a Fourier series of sine waves having frequencies $Nf_1 \pm f_2$, where $f_1$ is the frequency of the square wave and $N = 1,3,5 . . . .$ In this application $f_2 = 20$kHz, and $f_1$ varies from 20 to 30kHz, giving the following components:

| component | relative amplitude | frequency (kHz) |
|---|---|---|
| $f_1 - f_2$ | 1 | 0 to 10 |
| $f_1 + f_2$ | 1 | 40 to 50 |
| $3f_1 - f_2$ | $\frac{1}{3}$ | 40 to 70 |
| $3f_1 + f_2$ | $\frac{1}{3}$ | 80 to 110 |

There is a clear two octaves (10kHz to 40kHz) between the wanted signal and the lowest unwanted signals, so the latter can be easily filtered out leaving a pure sine wave. If the original sine wave is not pure but contains harmonics then the output from the mixer will have other components in the range 0 to 40kHz.

Operation of the mixer–Fig. 8–is explained with reference to Fig. 9 where a sine wave is being multiplied by a square wave of 1.5 times the frequency. In Fig.

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†'Second-order active filter circuit for tuned amplifiers and sinusoidal oscillators', E. A. Faulkner and Viscount Downe, *Electronic Engineering*, vol. 39 1967, p.287.
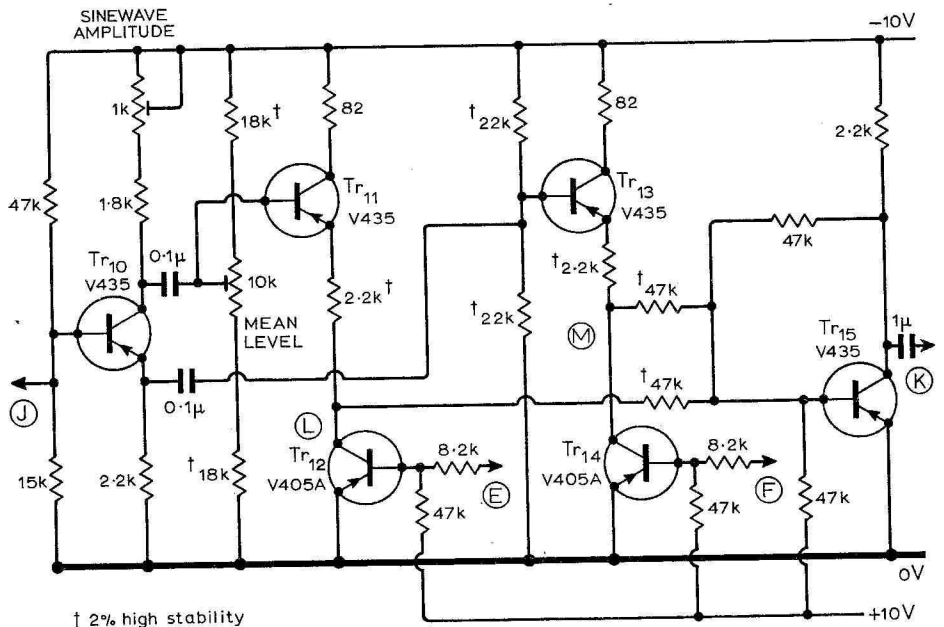


Fig. 8. Balanced mixer produces difference frequency between fixed and variable-frequency oscillators. Because the summing amplifier $Tr_{15}$ is linear, difference frequency signal amplitude is proportional to fixed-frequency signal amplitude and hence controlled by the a.l.c. circuit: In all these circuits alternative transistors are C724, BC107 for C424; C740, BFY76 for C450; V723, BSX29,36 for V405A and V723, BCY72 for V435A.
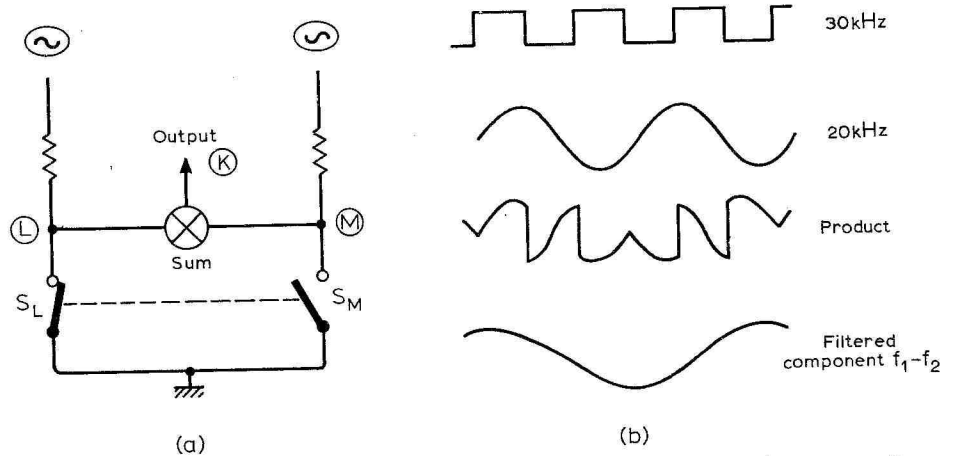


Fig. 9. Simplified circuit of switching mixer with points L and M at (a) corresponding to those in Fig. 8. Mixer waveforms at (b) are with fixed and variable frequencies in ratio 2:3. Bottom waveform is filter output.



Fig. 10. As well as difference frequency, there are additional components, e.g. $f_1+f_2$, $3f_1-f_2$, $3f_1+f_2$, which must be filtered. Low-pass filter shown has a cut-off frequency of 10kHz and a slope of nearly 36dB/octave giving 64dB attenuation at 40kHz.
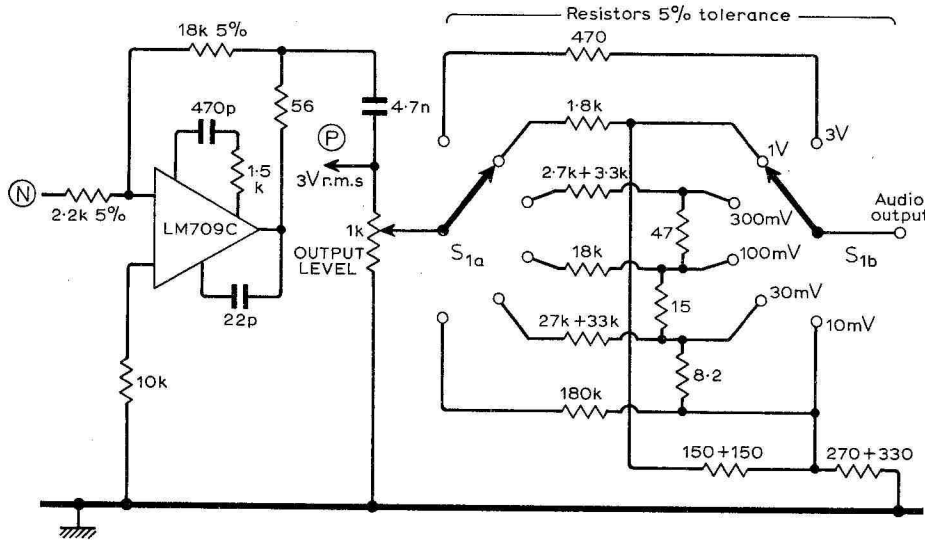
*Fig. 11. Signal level is amplified to 3V r.m.s. and fed to a switched attenuator, accurate to 5%.*

9(a) the switches $S_L$ and $S_M$ are closed alternately by the square wave while sine waves in antiphase are fed to the top ends of the resistors. When $S_L$ is closed $S_M$ is open and the output is essentially the inverse of the input sine wave. When $S_M$ is closed $S_L$ is open and the output is essentially the same as the input. So the required multiplying action is obtained. In practice $S_L$ and $S_M$ are transistors $Tr_{12}$ and $Tr_{14}$–Fig. 8–the sine waves are displaced from zero mean level so that these transistors are always forward biased.

Transistor $Tr_{10}$ is a phase inverter producing two equal and opposite signals to feed the emitter followers $Tr_{11}$ and $Tr_{13}$. The summing amplifier $Tr_{15}$ is
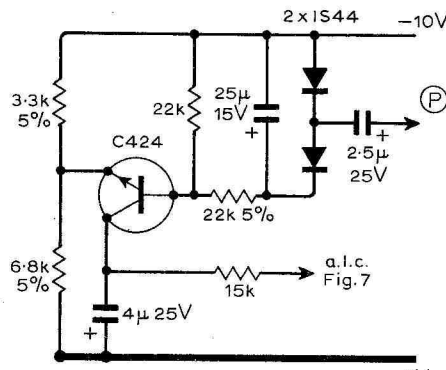


*Fig. 12. Filter characteristic has a rise of 2.6dB at 8kHz, but is reduced to 3% pk-pk with a.l.c. (at 4s sweep time). Level control signal is obtained by rectifying signal from D in Fig. 11.*



*Fig. 13. Sweep control circuit—between sweep generator and voltage tuned oscillator—enables various sweep modes to be set. The integrated circuit provides a low-impedance supply at 5V for setting zero-frequency output. Upper frequency limit is variable in switch positions W, S and X.*

linear in the sense that the amplitude of the component at $f_1-f_2$ is proportional to the amplitude of the sine wave $f_2$, thus the a.l.c. modulator controls the amplitude of the output.

## Filter

The filter used to separate the wanted component from the complex wave consists of three cascaded active low-pass sections–Fig. 10.

The measured overall gain of the complete filter rises from near unity (0dB) at zero-frequency to +2.6dB at 8kHz, after which it falls slowly to +1.3dB at the cut-off frequency 10kHz, and then rapidly at nearly 36dB per octave so that signals above 40kHz are attenuated by at least 64dB. The output from 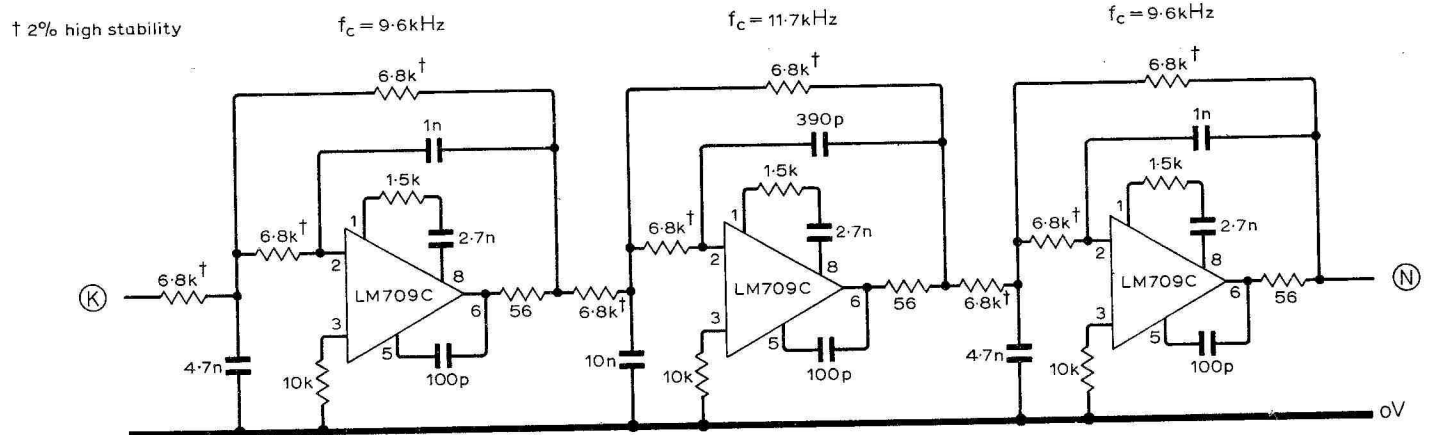the filter is amplified to 3V r.m.s. by a further LM709 used as a linear amplifier feeding the output attenuator–Fig. 11.

## Automatic level control

The output level will vary as the frequency sweeps across the range because of the filter characteristic. This variation is only about 3dB but it can easily be reduced with the automatic level control circuit–Fig. 12.

With this method the variation of amplitude over the entire frequency range was reduced to 3% pk-pk when the sweep period was four seconds. The levelling deteriorates at fast sweep rates because of the slow response of the detector caused by the smoothing components.

## Control network

Between the sweep generator and the voltage-tuned oscillator, the control network–Fig. 13–enables sweep frequency range to be set in several ways by the panel controls. A low-impedance supply at nominally −5V is provided by the operational amplifier–connected as a voltage follower. This potential, adjusted by the 2-k$\Omega$ potentiometer, is applied to one end of the frequency control and corresponds to zero-frequency output.

There are four sweep modes selected by $S_2$. In the 'full' mode the output frequency is swept from zero to 10kHz. The 1-k$\Omega$ preset potentiometer adjusts the amplitude of the sweep voltage across the frequency control so that this range can be set accurately. In the 'wide' mode the upper frequency limit is set by this frequency control. Some non-linearity in the calibration of this control is caused by loading but with the values shown this should be less than 1%.

In the 'symmetrical' mode the switch section $S_{2a}$ disconnects the sweep generator from the frequency control and applies a constant voltage, derived from the +10V supply, equal to the maximum positive excursion of the sweep waveform at A. If $S_3$ is set to 'c.w.', the

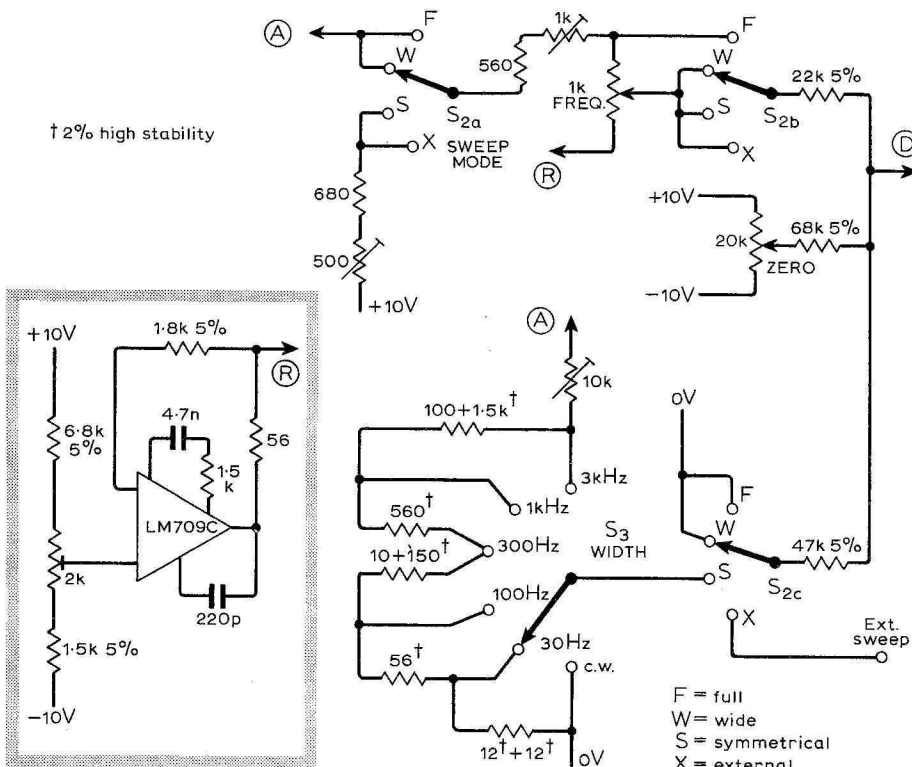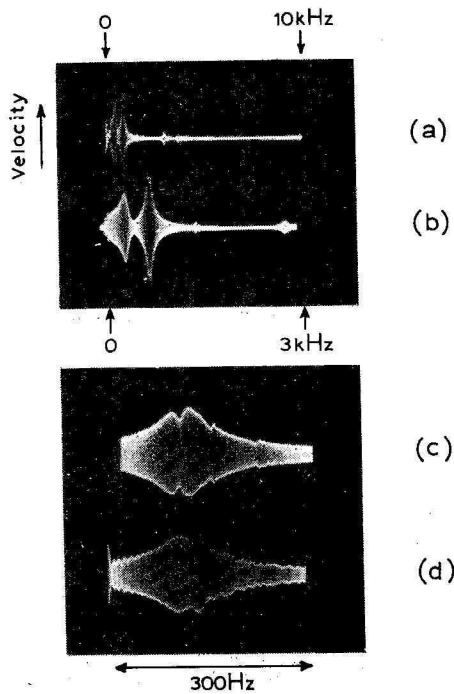*To illustrate the application of the sweep generator, the response of an earphone was investigated by another earphone 2mm from the diaphragm. Trace (a) shows voltage across the coil—proportional to velocity—with a 10-kHz frequency base. Trace (b) shows the trace expanded using 'wide' mode with a base of 3kHz. Second peak is resolved by using 'symmetrical' mode with a 770-Hz centre frequency and a 300-Hz sweep width (c). Decreasing sweep time from 7s per sweep to 0.5s per sweep shows the loss of resolution and spurious effects caused by too fast a sweep rate (d).*

frequency control adjusts the output frequency (unswept) with the same calibration as it had in the 'wide' mode. With $S_2$ and $S_3$ so set, the instrument thus behaves simply as an ordinary sine-wave signal generator. This is useful for setting the centre-frequency of a sweep to, say, the peak of a system's response and also for examining the output waveform of the oscillator itself. Symmetrical sweep about this mean frequency is provided in five calibrated widths by $S_3$. A similar arrangement holds for 'external' sweep waveforms with the mean frequency adjusted by the frequency control. The sensitivity for external sweep is approximately 800Hz per volt.

stabilized, both at less than 100mA The constancy of the ratio between these two voltages is more important for frequency stability than any common change in the supply voltages.

The oscillators were not sure-start if the two supplies were switched on simultaneously, and the +10V supply should be switched on first.

# Announcements

During the **British Association** annual meeting at Swansea there will be two public lectures at University College. The first, entitled 'The physics of musical sounds' will be given by Prof. C. A. Taylor (head of the department of physics, University College of South Wales) at 20.00 on September 6th. The second is devoted to fuel cells and will be given by F. T. Bacon (consultant to Energy Conversion Ltd) at 20.00 on September 7th.

A course of 20 weekly lectures (2.30-4.30) on **sound studios and recording** starts at the Polytechnic of North London, Holloway Rd, on October 28th (fee £10.50). On the same day at 6.30 a 15-lecture course on **audio and acoustic measurement** begins (fee £6.30). The lecturers include many well-known names in the audio world. Further particulars from the Dept. of Electronic & Communications Eng., Polytechnic of North London, Holloway Rd., London N7 8DB.

**Modern electronic techniques** is the title of a course of 10 afternoon lectures for engineers and technicians to be given at Portsmouth Polytechnic from October 8th. Fee £4.

Two one-day seminars on **computer-aided design** will be held at the Royal Garden Hotel, London, on September 21st and 22nd. Organized by our associate quarterly journal *Computer Aided Design*, the first day will be devoted to computer-aided design in shipbuilding and the second to c.a.d. in engineering. Fee £25 each seminar.

A residential vacation school on **lasers and their applications** is to be held at the City University, London E.C.1, from 13th to 24th September. The school is designed for physics and engineering graduates.

The **World Radio Club** programme, broadcast in the B.B.C's World Service on Thursdays at 12.45 G.M.T. and repeated on Fridays (23.45) and Sundays (08.15), is offering for the fourth year a DX award. The listening period is August 1-31, and entrants must give a concise reception report on one transmission from Great Britain and from each of the following: the Atlantic; East Mediterranean and the Far Eastern Relay Stations. Details from World Radio Club, B.B.C., Bush House, London WC2B 4PH.

**Racal** are negotiating to take over two more companies—Amplivox Ltd, manufacturers of hearing aids and other audio equipment, and Zonal Film (Magnetic Coatings) Ltd, the Ilford subsidiary manufacturing magnetic recording material and tapes.

The recently formed **International Radio and Electrical Distributors Association** is planning its first trade exhibition to be held at the Bloomsbury Centre, London W.C.1, from Sunday 21st May to Thursday 25th May next year.

**Specialist Switches**, who for almost 20 years have been providing a 24-hour service in custom built rotary and lever switches type H, DH, Hc and LO, have been taken over by Stoneleigh Electronics Ltd, and are now at Factory No. 8, Bridge Close, Romford, Essex.

**Texscan Instruments Ltd**, of Lord Alexander House, Hemel Hempstead, Herts, has been formed to handle the sales and service of the range of sweep signal generators, attenuators and filters produced by Texscan Ltd Inc., in America. The new company is also responsible for the marketing of wideband power amplifiers from Electronic Navigation Industries Inc., and function generators and r.f. power sources from Microdot Inc.

Emi has won a contract to develop for the Post Office experimental **digital transmitters and receivers** for the 10.7-11.7GHz band 'which could well provide a basis for the next generation of microwave equipment for the telecommunications network'.

An **airborne maritime radar** to detect submarines as well as to carry out general surveillance duties on maritime surface traffic is to be developed by EMI under a contract placed by the Ministry of Defence (Aviation Supply).

**GEC-Mobile Radio**, a division of Marconi Communication Systems Ltd, has moved from Coventry to the Chelmsford area. Their new headquarters is at Great Baddow, where the commercial and engineering departments are also housed, but the main service department will remain at Coventry Other service depots, such as those at Altrincham (Cheshire), Edinburgh, Leeds and London, will remain in operation.

Matsushita Electric are marketing a range of **printed wiring boards** through Steatite Insulations Ltd, of Hagley Rd, Birmingham, B16 8QW. The boards consist of phenolic resin impregnated paper, epoxy resin impregnated glass, and Duston plated—a process whereby powdered copper is fixed with adhesive ink to an insulated board which has a wiring pattern already on it.

R.E.W. Audio Visual Company, of London S.W.17. have been appointed sole distributors of the new series **7 Ferrograph-Dolby tape recorders** in the U.K.

Marconi Instruments Ltd are to supply to the Home Office an **automatic test system** for firemen's v.h.f. Personal Alerters. The 'Autotest' takes 30 seconds to automatically measure signal frequency, i.f. and audio frequency, as well as d.c.

FieldTech Ltd, of Heathrow Airport, have been appointed sole U.K. agents for a range of epoxy glass fibre **whip aerials,** manufactured by Valeriote Electronics (Guelph) Ltd, of Ontario, Canada. They cover the frequency range 2-30MHz, and are capable of handling 5kW average, 10kW peak.

**Gresham Recording Heads Ltd** have appointed Radio-Equipements, of 9 Rue Ernest Cognacs, 92-Levallois-Perret, France, agents to handle recording head sales throughout France.

**Motorola** have appointed GDS (Sales) Ltd to handle the sales of all their semiconductors in Eire and Northern Ireland.

# Books Received

**Transistor Circuits in Electronics,** second edition, by S. S. Haykin and R. Barrett. The book, written for students taking electronics in full-time or sandwich courses to degree, H.N.D. and H.N.C. level, will appeal also to electronics* engineers. In this new edition a chapter has been included on monolithic integrated circuits, and logic symbols have been redrawn to BS3939, 1969, Section 21. Chapter headings are as follows: transistor characteristics; graphical analysis; small signal equivalent circuits and parameters; amplifier circuits; feedback amplifier and oscillator circuits; switching circuits; regenerative switching circuits; logic circuits; modulator and demodulator circuits; and integrated circuits. There is a single page bibliography, an appendix explaining the super-position theorem and the theorems of Thévenin and Norton, and a five-page index. Pp. 367. Price £3.80 cased and £2.50 limp. Iliffe Books, Butterworth & Co. (Publishers) Ltd, 88 Kingsway, London WC2 6AB.

**A Dictionary of Electronics** by S. Handel, third edition. Five years have passed since the second edition of this reference work appeared. The first edition published in 1962 contained 384 pages and cost 7s 6d. The new edition contains 413 pages and costs 45p. Penguin Books Ltd, Harmondsworth, Middx.

# Helical V.H.F. Aerial

## Using twin helices with triangular cross-section

*by George J. Monser, M.S.E.E.*

To date the helical aerial, which has many desirable reception properties, has been overlooked to some extent for domestic v.h.f. reception, mainly because of the difficulties in building and installation. An attractive feature to recommend the helix is that it is circularly polarized, which means that it responds equally well to any linear polarization. As a result, fading effects due to propagation disturbances and multi-path effects tend to be minimized. In short, a gain of 8 to 10dB over the band can be provided, even under adverse conditions. Multipath propagation can change the plane of polarization. Thus, if you are using an aerial designed strictly for horizontal polarization, a signal loss of 3dB or 50% of the r.f. power may result.

Other attractive features of the aerial described are:

—It offers a nearly flat resistive impedance of 135 ohms, which means it can be connected directly to 300-ohm twin-lead with little loss.

—The phasing of the turns is such that it gives a maximum gain of 8 to 10dB over the band.

With such good features, why isn't the helix more frequently used? Mainly because conventional designs are cumbersome to build and difficult to install. First, as the name suggests, the radiating elements must be helical turns. When the size of these turns and the axial length are chosen for v.h.f. radio or television it is found that building such an antenna isn't so easy. Second, conventional designs are single-ended, or unbalanced. Thus, for proper operation, a sizeable ground-screen is required, posing difficult mounting problems.

By two simple modifications, the helix can be adapted for home construction. The first consists of changing the cross-section from circular to triangular. Thus, each turn is formed as a rigid triangle instead of a circular turn.

In the second modification, the turns are bifilar wound so that a balanced aerial is provided, requiring no ground screen. The cost of these modifications is slight.

Typically, conventional helices show 2 to 3dB variations in response with polarization. This model, when tested, showed 2 to 5dB variation, which is still quite acceptable.

The design, detailed in the illustrations, covers the band 88 to 170MHz which in the U.S.A. includes the f.m. sound broadcasting band and most of the v.h.f. television band. But the aerial has useful gain at higher frequencies e.g. about 6dB at 200MHz. It can be scaled for other frequency bands—I built a 1/15-scale model for testing.

By using a balun, it can feed receivers with unbalanced 75-ohm input circuits.

A

B

C
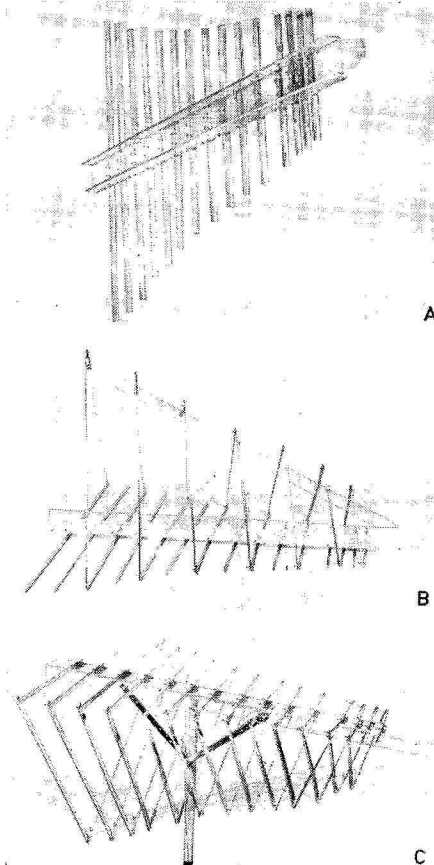
*Fig. A. Alternate elements on this U-channel form one side of the two tetrahedral helices. Model in this photograph has an extra element—extreme right—not shown in the table.*
*Fig. B. One complete helix with support attached to element vertices using either metal tie strips (three vertices on right) or with wire (vertices on left). The extra turn shown on this helix is optional.*
*Fig. C. Completed aerial with feeder cable attached to last two elements.*

*Fig. 1. Cut the aluminium strips to element lengths given in the table, and drill 3-mm holes as shown.*

*Fig. 2. For the element supports, cut the Perspex pieces and make notch as shown.*

*Fig. 3. Make the U-channel from three of the pieces shown in Fig. 2 (A, B & C) by drilling 3-mm holes, 25-mm deep at 76-mm intervals along both sides as shown and insert self-tapping screws.*

*Fig. 4. Attach the support—Perspex piece D—to the U-channel by drilling a 3-mm hole through notch and support, and fixing with nut and bolt.*

*Fig. 5. Cut 14 aluminium angle brackets to length shown, drill four 3-mm holes in each and bolt centrally to elements—numbers 1, 4, 7, 10, 13, 16 and 19 of both helices, see Fig. 6.*

*Fig. 6. Drill holes in U-channel to take elements and brackets as shown.*

*Fig. 7. Add remaining elements, completing one helix before starting the other. Open out lower support and fix to upper vertices in photographs with straps or wire. Support can be made more rigid by drilling a second hole at the notch and fixing with nut and bolt. Some holes at element joins may need redrilling.*

*Fig. 8. Bolt completed aerial to wood or plastics mast using four aluminium diagonal supports.*

### Table 1. Parts needed

**Perspex pieces**

1.15m × 38 × 6.5mm (45 × 1½ × ¼in) 2 off
1.2m × 76 × 10mm (48 × 3 × ⅜in)
1.2m × 38 × 6.5mm (48 × 1½ × ¼in)

**Aluminium strips**

1.83m × 13 × 1.6mm (72 × ½ × 1/16in) 13 off

**Aluminium right angle**

1.83m × 13 × 13 × 1.6mm (72 × ½ × ½ × 1/16in)
Also needed are self-tapping screws, nuts, bolts, tie strips and wire, wooden or plastics mast, 300-ohm balanced feeder cable.

### Table 2. Element lengths

| Element no. | first helix lengths (cm) | second* helix |
|---|---|---|
| 1 | 138 | 131 |
| 2 | 123 | 115 |
| 3 | 118 | 111 |
| 4 | 123 | 116 |
| 5 | 109 | 100 |
| 6 | 103 | 94 |
| 7 | 110 | 103 |
| 8 | 93 | 86 |
| 9 | 89 | 81 |
| 10 | 96 | 90 |
| 11 | 79 | 72 |
| 12 | 74 | 66 |
| 13 | 82 | 76 |
| 14 | 65 | 56 |
| 15 | 59 | 51 |
| 16 | 68 | 61 |
| 17 | 50 | 43 |
| 18 | 46 | 37 |
| 19 | 40 | 39 |

*Second-turn elements identified by asterisk in drawings.

**1**

13mm

13mm

6·5mm

13mm

3mm dia. holes

**2**

A,B 38mm

1·15m

2-off

6·5mm (1/4 in.) thick

C 76mm

1·15m

76mm

38mm

9·4mm (3/8 in.) thick

8 x 33mm notch

D 38mm

1·2m

6·5mm (1/4 in.) thick

**3**

Side 2

Base

A

C

Side 1

B

U-channel (side view)

**4**

Support
1·2m x 38mm x 6·5mm

D

Angle set during
final assembly

U-channel

**5**

50mm

3mm dia. holes

25mm

13mm x 13mm x 1·6mm
(1/2 in. x 1/2 in. x 1/16 in.)
76mm long

3mm dia. holes

Centre angle
pieces

**6**

Aerial elements
(mounted on U-channel)

U-channel

Angle set
at final
assembly

Lower
support

76mm spacing between these elements

51  25
mm  mm

90mm

77
mm

1   1*   4   4*   7   7*   10   10*   13   13*   16   16*   19   19*

**7**

Feed points

19*
19
16*
16
13*
13
10*
10
7*
7
4*
4
1*
1

17
18
14
15
11
12
8
9
5   6
2
3

**8**

U-channel

Approx.
53cm
from
notch

8mm dia holes

33cm aluminium
diagonal supports

Mast support

# Ceramic Discriminator for Narrow-band F.M.

## Product Application Note

*by D. Balfour*

Piezoelectric materials have been employed in the communications industry for many years, the most commonly known being the quartz crystal used widely as an oscillator and for filter networks. The ceramic resonator has achieved similar penetration as a frequency determining element or as part of a filter network at frequencies around 455 kHz. Both these devices are similar in that being piezo-electric their mechanical vibrations may be considered in terms of electrical parameters of inductance, capacitance and resistance. Their equivalent circuit in the vicinity of resonance is shown in Fig. 1.

Quartz, however, whether in its natural state or whether grown synthetically, has a fixed set of piezoelectric constants, which limit some of the electrical values obtainable, for instance the ratio of $C_S$ to $C_1$.

**Table 1**

| Type | Channel spacing | Resonances ref. 455kHz | $L$ mH | $C_1$ pF | $R$ $\Omega$ | $C_s$ pF |
|------|-----------------|------------------------|--------|----------|--------------|----------|
| TFD4 | 20-25 kHz | ± 18 kHz | 2.2 | 61 | 15 | 360 |
| TFD5 | 12.5 kHz | ± 12 kHz | 3.2 | 40 | 15 | 369 |

of $LC_1R$. Between these two frequencies $F_R$ and $F_A$ the impedance of the transfilter is largely inductive, becoming entirely resistive both at $F_R$ and at $F_A$. By altering the piezoelectric coupling of the material we can alter the spacing between $F_R$ and $F_A$ within wide limits, and have manufactured two devices (TFD4 and TFD5) suitable for 25 and 12.5 kHz channel spacing systems respectively. Brief details of the devices are given in Table 1.

Resonances have been chosen symmetrical to 455 kHz and are placed almost at the adjacent channel frequencies. This placing helps improve the overall rejection of the complete system. Typical values for the parameters of each device are tabulated. In general, the admittance of the device may be calculated exactly from the equation:—
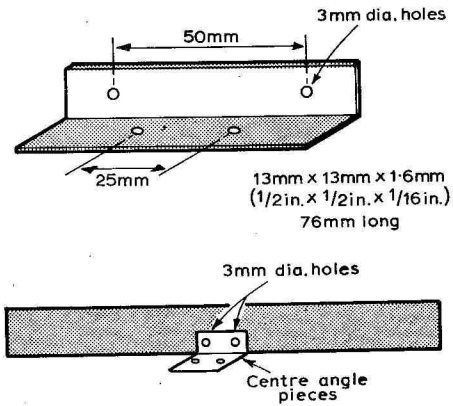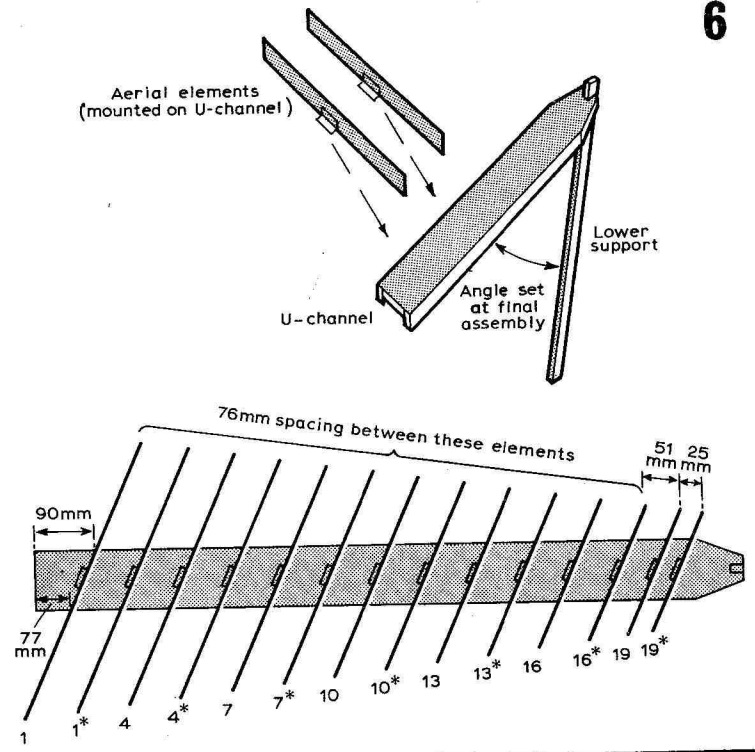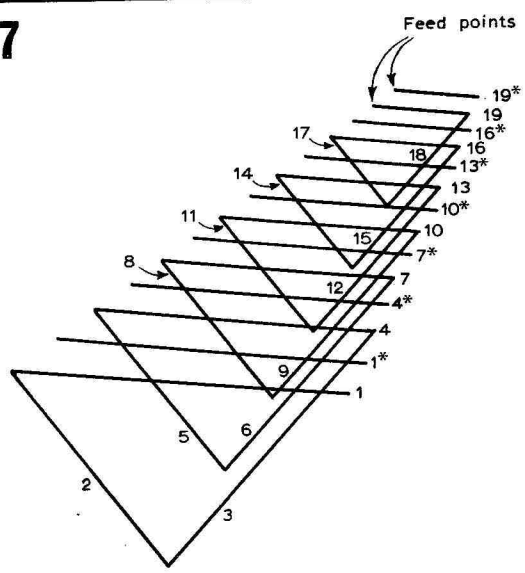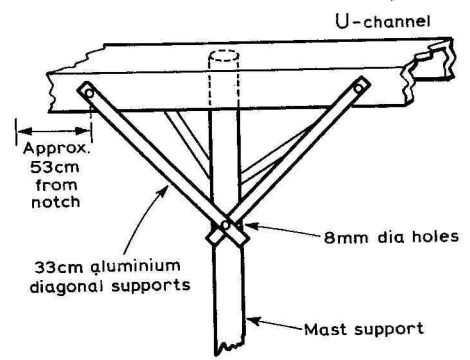
$$Y = \frac{1}{R + j\omega L + \frac{1}{j\omega C_s}} + j\omega C_s$$

This is cumbersome and it can be shown that within ±5 kHz of 455 kHz that the network can be considered lossless with little error. It can further be shown that the impedance may be expressed as follows:
$$jZ \text{ where } Z =$$
$$1.0 + 0.11f + 0.006f^2 \text{ k}\Omega \text{ (TFD5) or}$$
$$1.3 + 0.24f + 0.02f^2 \text{ k}\Omega \text{ (TFD4)}$$
where $f$ represents the deviating frequency in kHz. This impedance is approximately true for ±3 kHz for the TFD5 and ±5 kHz for the TFD4.
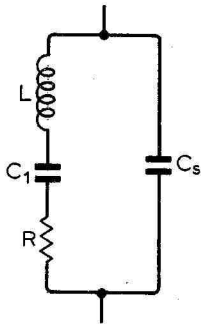
The ideal device would have a linear



*Fig. 1. Equivalent circuit of a piezoelectric resonator.*

Ceramic can have its piezoelectric values altered over a wide range, which enables a more flexible series of designs to be achieved. Ceramic is, however, less stable both with temperature and time than quartz, hence the use of ceramic filters at 455 kHz, where the stability with respect to temperature in absolute terms is equal to that of a quartz filter at 10.7 MHz.

An analysis of the equivalent circuit of Fig. 1 shows that the impedance plot is characterized by a minimum at a frequency $F_R$, very close to the resonance of the series arm $LC_1R$ and an impedance maximum at a frequency $F_A$, where $C_S$ has a capacitive impedance equal to the inductive impedance

* Vernitron Ltd



*Fig. 2 Circuit of the discriminator.*

change of impedance with frequency and the following procedure may be used to achieve a substantially linear output with the circuit of Fig. 2.

Assuming that both $A$ and $B$ are resistive the following expression gives the ratio of the output volts to the input volts:—

$$V_o = \frac{B}{A+B} \cdot \frac{V_{in}}{\sqrt{1 + \left(\frac{AB}{(A+B)Z}\right)^2}}$$

The problem then resolves itself into two forms:—

1. To linearize the expression:

$$\frac{1}{\sqrt{1 + \left(\frac{AB}{(A+B)Z}\right)^2}}$$

so that it becomes closely linear with frequency. This involves choosing a specific value for the expression $AB/(A + B)$, which is the value of $A$ in parallel with $B$ and is not definitive with respect to $A$ or $B$.

2. To maximize the value of the output by choosing $B/(A + B)$ to give the greatest sensitivity commensurate with the limits in 1. The value of $AB/(A + B)$ has been computed to give a linear relationship of output versus deviation for the points $f = -3$, 0 and $+3$ for the TFD5 and $= -5, 0, +5$
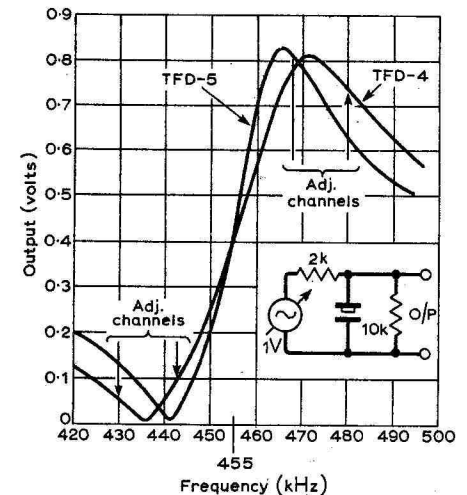


*Fig. 3. Performance of slope detector using TFD4 and TFD5.*

for the TFD4, using figures in Table 1. The results are as follows:—

$$\text{TFD5} \quad \frac{AB}{A+B} = 1.5 \text{ k}\Omega$$

$$\text{TFD4} \quad \frac{AB}{A+B} = 1.7 \text{ k}\Omega$$

For practical purposes values of 2kΩ for $A$ and 10kΩ for $B$ are satisfactory for both TFD4 and 5. The sensitivities achieved are 32mV/kHz for a 1V input for the TFD4 and 50mV/kHz for a 1V input for the TFD5, as shown in Fig. 3.

The adjacent channel sensitivity is much less than for wanted channel, typically 0.33. This means that the discriminator acts as a filter for adjacent channel signals giving 8 to 10 dB rejection.

# Dual-trace Oscilloscope Unit

## 2. Field-effect transistor amplifier

*by W. T. Cocking\*, F.I.E.E.*

**The basic requirements for a unit which enables two signals to be seen simultaneously on an oscilloscope were discussed in Part 1, where it was shown that two identical amplifiers with input attenuators and an electronic switch are required. A maximum overall gain of unity is needed but, to reduce the effective capacitance of the input cable, input attenuation must be used and so subsequent amplification must be included to offset this. It is important that the input resistance of the amplifier be well defined, which means that it must be provided substantially by a resistor, and so, with the usual parallel connections, the amplifier proper must have an input resistance which is very large in comparison.**

The junction field-effect transistor is the obvious choice for the input stage of any amplifier which must have a high input resistance. Its main drawback is its enormous tolerances. It is also rather more costly than the usual bipolar transistor.

Fig. 1 shows the characteristics of the

*\*Editor in Chief, Wireless World*

BFW10 f.e.t. At zero gate voltage, the drain current may be from about 7·8 mA to about 20 mA, while the gate cut-off voltage may vary from −2·1 V to −8 V. In an amplifier in which it is impracticable to use capacitance interstage couplings, it is imperative that the d.c. level of the output electrode be substantially constant and this is where the difficulty in using the f.e.t. arises.

If the f.e.t. is employed as a source-follower, it is necessary for the source to be always at some fixed voltage relative to earth. One can use a variable source resistor which is adjusted to suit the particular f.e.t. employed. For example, one might decide to operate at 1·5 mA to suit a low-tolerance f.e.t., which will demand a gate bias of −1·5 V. If, as is usual, the gate is returned to earth, the source must be 1·5 V above earth and at 1·5 mA, a source resistor of 1 kΩ is needed. With a high-tol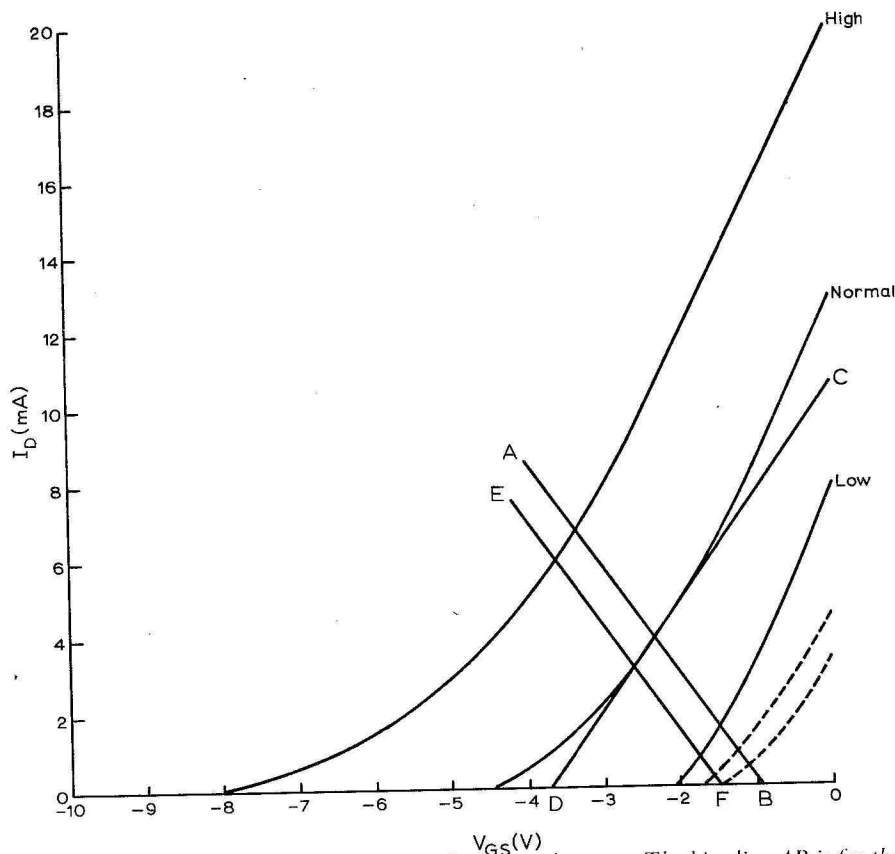erance f.e.t. the source must still be 1·5 V above earth but the current will be 13·9 mA and the source resistance must be 108 Ω only. This is far too low for good source-follower action and instead of the "gain" approaching unity, it will be around 0.23 only.

A better alternative is to use a fixed value of source resistance. Constant voltage then demands constant current, which must be chosen at a suitable value for a low-tolerance f.e.t., say, at 1·5 mA. Control must then be exercised by a variable negative gate bias which is adjusted to suit the f.e.t. It can be seen from Fig. 1 that for a high-tolerance f.e.t. −6 V bias will be needed to give 1·5 mA. This means −4·5 V bias additional to the 1·5 V source bias. The gate return now cannot be earthed, but must be taken to a source of up to 4·5 V negative to earth, which must be stabilized. This is inconvenient. The "gain" is still not constant, but is more constant than with a variable source resistor. This is because at constant current a high-tolerance f.e.t. has a much lower mutual conductance than a low-tolerance one, as can be seen from the slope of the curves in Fig. 1.

In the writer's view there is only one practicable way of coping satisfactorily with the tolerances of the f.e.t. when there must be an output point at a constant voltage to earth. This is to use it with a p-n-p transistor (if it is an n-channel f.e.t.) in the circuit shown in Fig. 2. The resistor $R_D$ is made variable and is adjusted to bring the collector of $Tr_2$ to a fixed voltage $V_C$ with respect to earth.

Ideally, the voltage amplification is $1 + R_C/R_S$. In practice, it is somewhat less. It can be within about 95% of this figure for low tolerance to normal f.e.ts, but it falls off more with high-tolerance ones because $R_D$ then becomes too small. The circuit is an admirable one for an f.e.t. with a tolerance range of about one-half of that of the



Fig. 1. *Curves of the BFW 10 f.e.t. showing the large tolerances. The bias line AB is for the circuit of Fig. 2 with $V_C = 4V$, $R_S = 470\ \Omega$ and $R_C = 1·5\ k\Omega$. The line EF is for the same values but $V_C = 6·3\ V$. CD is drawn tangentially to the f.e.t. curve at its intersection with AB and its slope gives the mutual conductance. The dotted curves are for two specimens of the T1S58 which were used in the experiments.*

BFW10. Such f.e.ts are available, but naturally tend to cost more.

The circuit has a low output impedance and so is not much affected by a load $R_L$ as long as this does not draw direct current. It has a good high-frequency response and works well up to at least 10 MHz.

Assuming, as usual, that the base current of $Tr_2$ is negligible compared with the collector current,

$$V_C = V_S + I_C R_C$$

$$V_S = (I_C + I_D) R_S$$

$$V_{BE} = I_D R_D$$

from which

$$V_S = V_C \frac{R_S}{R_C + R_S} + I_D \frac{R_C R_S}{R_C + R_S}$$

The f.e.t. thus behaves as if it were source biased by a resistance having the value of $R_C$ and $R_S$ in parallel returned to a voltage positive to earth by $V_C R_S/(R_C + R_S)$. If, for example, $R_S = 470\,\Omega$ and $R_C = 1.5\,k\Omega$, the parallel value is $358\,\Omega$. If $V_C = 4\,V$ the effective return voltage is 0.95 V. By drawing a bias line from this voltage to represent $358\,\Omega$ in the usual way, the intersections with the f.e.t. curves enable $I_D$ and $V_S$ to be
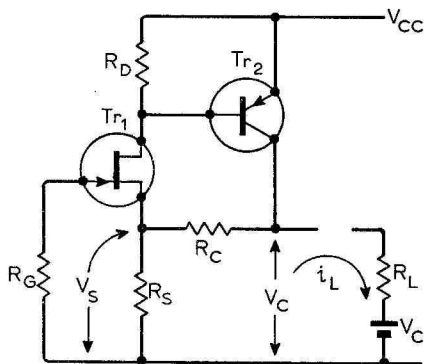


Fig. 2. Circuit for an n-channel f.e.t. with p-n-p bipolar transistor.

read off for low- and high-tolerance and normal f.e.ts. The line AB in Fig. 2 represents the conditions. For a low-tolerance f.e.t., $V_S = 1.5\,V$, $I_S = 1.5\,mA$; for a normal f.e.t., $V_S = 2.35\,V$, $I_S = 3.9\,mA$; and for a high-tolerance f.e.t., $V_S = 3.4\,V$, $I_S = 6.8\,mA$, within the accuracy limitations of small-scale graphs.

The required value of $R_D$ is $V_{BE}/I_D$. Taking $V_{BE} = 0.65\,V$, $R_D$ is $433\,\Omega$, $166\,\Omega$ and $95.5\,\Omega$ in the three cases. In practice, there is a manufacturing tolerance on $V_{BE}$ which is usually of the order of $\pm 75\,mV$. This is covered by providing $R_D$ with a somewhat greater range of resistance.

At this stage $I_D$, $V_C$, $V_S$ and $R_D$ are known. The collector current is

$$I_C = \frac{V_C - V_S}{R_C}$$

For the three cases of the BFW10, with $R_C = 1.5\,k\Omega$, the collector current is $1.67\,mA$, $1.1\,mA$ and $0.4\,mA$ respectively.

The signal conditions are more complex and the equations for gain are developed in the Appendix. The performance depends very largely upon a quantity which is there

termed $y$. It is the effective current gain of $Tr_2$ and in the limit becomes $h_{fe}$. This only occurs when $R_D$ becomes infinite and is only approached when the input resistance $h_{fe}r_e$ of $Tr_2$ is very small compared with $R_D$ so that substantially all the signal current of $Tr_1$ flows into the base of $Tr_2$.

For given values of $V_C$, $R_C$ and $R_S$, $I_D$ and $V_S$ are much greater for a high-tolerance f.e.t. than for a low, and so $I_C$ must vary inversely. As $I_D$ rises, $R_D$ must be reduced, and as $I_C$ becomes less, $r_e$ increases. The result is that $y$ varies very greatly between low- and high-tolerance f.e.ts.

Taking $V_{BE} = 0.65\,V$, which is typical, the value of $y$ given in the Appendix can be expressed in d.c. terms as

$$y = \frac{h_{fe}}{1 + \frac{h_{fe}I_D R_C}{25(V_C - V_S)}}$$

It is clear from this that $V_S$ must not approach $V_C$ too closely. If it does, the denominator will become large and will vary very much with small changes of $V_S$. This means that the collector current must not be too small. For a large value of $y$ it is necessary that the collector current be much larger than the drain current, but this is not always practicable.

In general, the larger $V_C$ the better, but there is a limit set by the requirements of a low-tolerance f.e.t. It is essential that the value of $V_C R_S/(R_C + R_S)$ be numerically less than the cut-off bias of a low-tolerance f.e.t.

Tables 1 and 2 give the calculations step by step for the BFW10 using the curves of Fig. 1 and taking $g_m = 3\,mA/V$ in all cases, since for the particular conditions it varies very little. In all cases, $h_{fe} = 100$, $R_C = 1.5\,k\Omega$ and $R_S = 470\,\Omega$; for Table 1, $V_C = 4\,V$, while for Table 2, $V_C = 6.3\,V$. In the two cases, the bias lines in Fig. 1 are AB and EF respectively.

With $V_C = 4\,V$, the gain varies from 2.24 to 3.93, a ratio of 1.75:1, whereas with $V_C = 6.3\,V$, it varies from 3.48 to 4.09 only, a ratio of 1.17:1. With the higher voltage, the output resistance is also much less.

If the circuit has a load $R_L$ this load must not draw direct current for the analysis of the Appendix to hold. The load can be fed through a capacitor, or it can be connected directly if its earthy end is taken to a voltage equal to $V_C$. The practical difficulty is then to ensure that temperature changes do not upset matters.

**Table 1**

| | Low | Normal | High | |
|---|---|---|---|---|
| $V_S$ | 1.5 | 2.35 | 3.4 | V |
| $I_D$ | 1.5 | 3.9 | 6.8 | mA |
| $V_C - V_S$ | 2.5 | 1.65 | 0.6 | V |
| $I_C$ | 1.67 | 1.1 | 0.4 | mA |
| $R_D$ | 433 | 166 | 95.5 | Ω |
| $r_e$ | 15.6 | 23.6 | 65 | Ω |
| $R_D/h_{fe}$ | 4.33 | 1.66 | 0.955 | Ω |
| $r_e + R_D/h_{fe}$ | 19.99 | 25.26 | 65.955 | Ω |
| $y$ | 21.65 | 6.6 | 1.45 | |
| $g_m R_S(1+y)$ | 31.9 | 10.7 | 3.45 | |
| $\dfrac{g_m R_S(1+y)}{1 + g_m R_S(1+y)}$ | 0.97 | 0.915 | 0.775 | |
| $1 + \dfrac{R_C}{R_S} \cdot \dfrac{y}{1+y}$ | 4.05 | 3.77 | 2.89 | |
| $A$ | 3.93 | 3.45 | 2.24 | |
| $R_o$ | 122.5 | 345 | 905 | Ω |

**Table 2**

| | Low | Normal | High | |
|---|---|---|---|---|
| $V_S$ | 1.575 | 2.62 | 3.65 | V |
| $I_D$ | 0.7 | 3.16 | 6 | mA |
| $V_C - V_S$ | 4.725 | 3.68 | 2.65 | V |
| $I_C$ | 3.15 | 2.45 | 1.77 | mA |
| $R_D$ | 925 | 206 | 108 | Ω |
| $r_e$ | 8.25 | 10.6 | 14.7 | Ω |
| $R_D/h_{fe}$ | 9.25 | 2.06 | 1.08 | Ω |
| $r_e + R_D/h_{fe}$ | 17.5 | 12.66 | 15.78 | Ω |
| $y$ | 52.9 | 16.3 | 6.85 | |
| $g_m R_S(1+y)$ | 76 | 24.4 | 11.1 | |
| $\dfrac{g_m R_S(1+y)}{1 + g_m R_S(1+y)}$ | 0.987 | 0.96 | 0.916 | |
| $1 + \dfrac{R_C}{R_S} \cdot \dfrac{y}{1+y}$ | 4.13 | 4 | 3.79 | |
| $A$ | 4.09 | 3.84 | 3.48 | |
| $R_o$ | 53 | 161 | 337 | Ω |

Rather surprisingly, there is little published information on the temperature characteristics of the f.e.t. Two different effects exist. There is the usual negative effect of a junction which results in the drain current increasing with temperature, but there is a positive effect in the bulk material which has the opposite effect. At one particular low current the two can cancel, but at normal drain currents the junction has the greater effect, and normally drain current increases with temperature.

The bipolar transistor itself behaves normally, of course. However, the change of $V_{BE}$ affects $I_C$ by an amount which depends greatly upon the value of $R_D$. If $R_D$ is very large, $V_{BE}$ hardly affects $I_C$ at all, but it can exercise almost its full effect when $R_D$ is small. It can be expected, therefore,
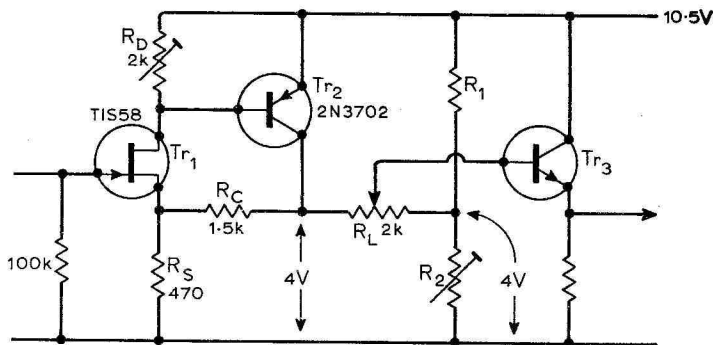


Fig. 3. The circuit of Fig. 2 elaborated to include a gain control; $R_D$ is adjusted to bring the collector of $Tr_2$ to a design value and $R_2$ is adjusted for zero volts across $R_L$

that the overall temperature coefficient will be greater with a high-tolerance f.e.t. than with a low. An intelligent guess would put the overall temperature coefficient at about 3 mV/°C referred to the gate. At the output this will appear multiplied by the gain as a change of $V_C$. It may thus be 12 mV/°C or about $\pm 0.15$ V for $\pm 12.5$ °C temperature change.

Fig. 3 shows a circuit which was used experimentally. It was designed for $V_C = 4$ V. The f.e.ts used had the characteristics shown dotted in Fig. 1. Two rather similar specimens were used and both were much "lower" than a low-tolerance BFW10; obviously $V_C = 6.3$ V would not be appropriate with these. To set up the circuit an Avo on its 10 V range was connected between earth and the collector of $Tr_2$ and $R_D$ was adjusted for a reading of 4 V. The meter was then connected across $R_L$ and $R_2$ adjusted for zero volts, first on the 10 V range, then on the 2.5 V, and finally on the 50 $\mu$A range. It was found desirable to use an emitter-follower after the stage, partly to reduce capacitance loading on $R_L$ but mainly to reduce the base current in $R_L$. The high-frequency response can be extended by adding a small capacitance (e.g., 25 pF) across $R_S$.

With the following stages an overall response almost flat to 5 MHz, and about $-3$ dB at 10 MHz, was readily obtainable. The only fault of the circuit lay in the difficulty of maintaining an adequate balance of the voltages at the two ends of $R_L$. An out-of-balance current of 10 $\mu$A in $R_L$ is about as much as can be tolerated and this corresponds to only 20 mV cross $R_L$. One would expect this to occur with a temperature change of only around 2°C.

To maintain good balance both ends of $R_L$ must be connected to points which vary in voltage by the same amount. The only way which seems likely to give this reasonably well is to replace $R_1$ and $R_2$ by a duplicate amplifier and this requires the two f.e.ts to be fairly closely matched. This was not done because it was considered undesirable to use matched f.e.ts.

The circuit was, thus rather regretfully, abandoned. It should be understood that this was only because of the gain control. If that were not needed, and low-tolerance f.e.ts could be guaranteed, then $V_C$ could be 2.7 V only and the stage could drive the switched transistor directly. The temperature coefficient would not be important because it would only affect the position of the trace on the screen and a shift control is needed in any case and could correct it. The shift control would, in fact, be $R_D$, or a portion of it.

Before concluding, it is desirable to point out that a bipolar transistor can be used instead of an f.e.t. This is shown in Fig. 4. For simplicity, we shall treat this as an extension of the f.e.t. analysis and so shall call the collector current of $Tr_1$ $I_D$ instead of the more usual $I_{C1}$. The previous equations apply, but additionally,

$$V_{B1} - V_{BE1} = V_S$$

There is now no reason why the current of $Tr_1$ should not be much less than that of $Tr_2$; $R_D$ can be large and $y$ can be large.
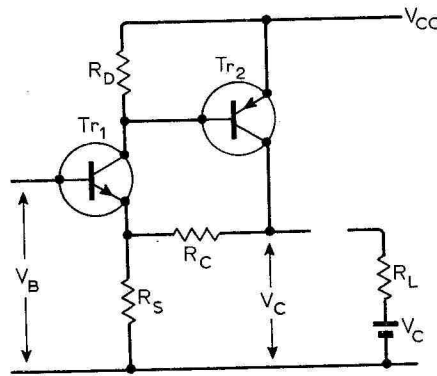


Fig. 4. The circuit of Fig. 2 but with an n-p-n transistor in place of the f.e.t.
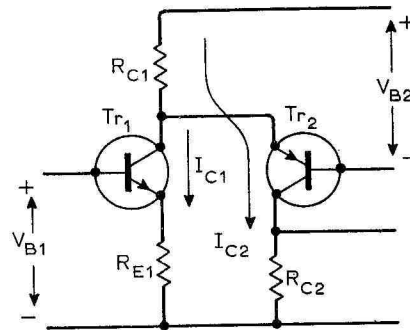


Fig. 5. A form of cascode circuit using n p n and p n p transistors which is useful for passing a signal between points at different d.c. voltage levels. A shift control which does not affect gain can be obtained by adjusting $V_{B2}$. The gain tends in the limit to $R_{C2}/R_{E1}$ but in practice is a little lower.

Ignoring the internal resistance of the emitter junction of $Tr_1$, the input resistance is

$$R_{in} = h_{fe1} R_S (1+y) \frac{R_C + R_C/(1+y)}{R_C + R_L + R_S}$$

and it can easily be several megohms. Notice, however, that $R_{in}$ is reduced by a finite value for $R_L$. This is physically obvious for when $R_L$ is present some of the collector current of $Tr_2$ flows through it instead of it all passing through $R_S$. The feedback, upon which the high input resistance depends, is thus reduced.

The gain equation is now very nearly the ideal of $1 + R_C/R_S$, since $g_m$ for a transistor is usually 30 mA/V or more and $y$ can easily be made 50 or more. It will rarely be less than the ideal by more than 5%. Also the output resistance is much smaller and is nearly

$$R_o = \frac{R_C}{1+y}$$

With this circuit it is possible to obtain with a load of 2 k$\Omega$ an input resistance of at least 2 M$\Omega$, an output resistance of around 15 $\Omega$ and a gain of about four times and it is useful up to at least 10 MHz. If $R_D$ is fairly large, as it can be, (e.g., 1 k$\Omega$) changes of $V_{BE2}$ have little effect. However, changes of $V_{BE1}$ are subject to the full gain. Thus, changes of 2 mV/°C become changes of 8 mV/°C at the output.

It should be understood that the gain variations of the f.e.t. are dependent on the particular f.e.t. used. Design must be carried out so that the minimum gain is greater than the required gain and a pre-set gain control included. It is, of course, possible to reduce the variations by using a fixed value of $R_D$ and adjusting $V_C$ by negative bias on the gate. As mentioned earlier, this is not a complete cure and the need for a stabilized negative supply line is undesirable. It is a merit of the arrangement of Fig. 2 that the performance and $V_C$ are substantially independent of $V_{CC}$ so that a stabilized supply is unnecessary.

We have not so far discussed the possibility of obtaining higher gain. For $A = 10$, $R_C/R_S = 9$ ideally, and in practice probably about 12. The effective bias line of 358 $\Omega$ is about the optimum for minimizing variations of $g_m$, so keeping this figure, we find $R_S = 390$ $\Omega$ and $R_C = 4.7$ k$\Omega$ approximately. For the bias line to start from 0.9 V, as before, $V_C$ must be 11.7 V. As $V_{CE}$ must be at least 2 V, $V_{CC}$ must be at least 14 V. All this is quite possible.

It is unlikely, however, that the frequency response would be adequate, and $R_L$ could certainly not be increased proportionately to $R_C$ without seriously affecting the response. In view of the difficulty of maintaining an adequate balance of the voltages at the ends of $R_L$ it was regretfully decided to abandon the circuit, and no work was done in an attempt to obtain higher gain.

It should be pointed out that further transistors are needed to couple the stage to the output stage. The base of the output stage is to be at 2.7 V; the output of the amplifier would be at 6.3 V. The amplifier output is in the same phase as the input, but the output stage gives a phase-reversal. It is desirable that there should be no overall phase-reversal, so the intermediate stage should be phase reversing or a common phase-reversing stage can be used after the common outputs of the two channels.

Fig. 5 shows a very useful circuit for connecting two difficult voltage levels. From the signal point of view it is a form of cascode stage and gives phase reversal. For $Tr_1$

$$V_{B1} - V_{BE1} = I_{C1} R_{E1}$$

and for $Tr_2$

$$V_{B2} - V_{BE2} = (I_{C1} + I_{C2}) R_{C1}$$

$$V_o = I_{C2} R_{C2}$$

and the gain is nearly $R_{C2}/R_{E1}$. Also

$$V_{CE1} = V_{CC} - I_{C1}(R_{C1} + R_{E1}) - I_{C2} R_{C1}$$

$$V_{CE2} = V_{CC} - I_{C1} R_{C1} - I_{C2}(R_{C1} + R_{C2})$$

The usual practical difficulty is to make $V_{CE1}$ and $V_{CE2}$ large enough. Suppose, $V_{B1} = 6.3$ V, $V_o = 2.7$ V, and $V_{BE1} = V_{BE2} = 0.65$ V with $V_{CC} = 10.5$ V min. Then $I_{C1} R_{E1} = 5.65$ V. Suppose, $R_{E1} = 1.5$ k$\Omega$, then $I_{C1} = 3.76$ mA. If $R_{C2} = 1.5$ k$\Omega$, $I_{C2} = 2.7/1.5 = 1.8$ mA. Now for $Tr_1$, let $V_{CE1} = 3$ V. The collector is then 8.65 V above earth and we can drop only $10.5 - 8.65 = 1.85$ V in $R_{C1}$ with a current of $3.76 + 1.8 = 5.56$ mA, so $R_{C1} = 332$ $\Omega$. We then have $V_{CE2} = 10.5 - 1.85 - 2.7 = 5.95$ V. The only thing wrong with this is that 330 $\Omega$

is rather low for $R_{C1}$. The loss of signal can be corrected by increasing $R_{C2}$ and frequency correction can be obtained by shunting $R_{E1}$ by a small capacitance ($\approx$ 25 pF). A shift control which does not affect the gain can be obtained by making $V_{B2}$ variable. It must, of course, be nominally 0·65 V plus the drop across $R_{C1}$ or 2·5 V negative to $+V_{CC}$ and the supply must be stabilized with respect to $+V_{CC}$. This is easily done with a zener diode.

Thus, with an f.e.t. input stage we need a minimum of one f.e.t. and three bipolar transistors prior to the output stage. The arrangement has been fully tried out and with small capacitances across $R_S$ (Fig. 3) and $R_{E1}$ (Fig. 5) an overall frequency response down by only 3 dB at 10 MHz was readily obtainable. The only fault lay in the inability to maintain the current in the gain control resistor small enough. It was felt that if the circuit was used it would be necessary to provide a balance adjustment as a panel control. In view of this it was decided to investigate other methods.

It may be asked at this point why the gain control was not capacitatively coupled to remove d.c. from it. This was actually tried and abandoned. In the first place, because of the low resistance values needed to maintain the high-frequency response, at least 500 $\mu$F is needed. This means electrolytic types must be used and these have a leakage current. This can be small initially if their voltage rating is high compared with the actual voltage across them, and a trial showed it to be negligible. However, according to the books an electrolytic capacitor used on a low voltage gradually reforms to a working voltage near to that applied and then passes a relatively high leakage current. If this does occur, it means that after three months or so, there would be excessive current in the gain control.

A second reason for avoiding coupling capacitors is that it would be necessary to include protective diodes and resistors. Without them, there is no more certain way of obtaining a heavy mortality in transistors! The trouble occurs when switching on and off. Protective circuitry not only adds to the cost, but tends to reduce the high-frequency response. We tried capacitors without such circuitry and several transistors died!

Before concluding this part, it may be well to say a few words about another circuit

which was tried. The merit of this circuit, Fig. 6, is that ideally there is no current in $R_{B1}$, which solves the gain control problem. The circuit is usually used without $R_{E1}$ and $R_{E2}$, but they were included so that the currents in $Tr_1$ and $Tr_2$ would be better determined.

Transistors $Tr_1$ and $Tr_2$ are supposed to pass equal currents. Their base voltages must be the same except for any difference between $V_{BE1}$ and $V_{BE2}$. Ignoring base currents, $R_{B1}$ and $R_{B2}$ must thus be returned to substantially the same voltage. Now if current flows in $R_{B1}$ and $R_{B2}$ from $Tr_3$, there must be a voltage drop across $R_{B2}$ and the base of $Tr_2$ will not be at the same voltage as the base of $Tr_1$. But the base voltages cannot differ appreciably and so there cannot be current in $R_{B1}$ and $R_{B2}$. Thus the collector voltage of $Tr_3$ to earth is the same as the base voltages of $Tr_1$ and $Tr_2$.

With the particular conditions of Fig. 6, the base supply voltage for $Tr_1$ had to be 3·6 V compared with the base supply of 2·8 V for $Tr_2$, a difference of 0·8 V. In part this may be accounted for by differences of $V_{BE1}$ and $V_{BE2}$, but it was largely caused by the high base current of $Tr_1$ (9 $\mu$A) in the high base resistance (100 k$\Omega$). This alone gave a bias difference of 0·9 V. In fact, the transistor used for $Tr_1$ had $h_{fe} = 55$ only.

The gain of the stage is nominally $1 + R_{B1}/R_{B2}$ and this is 3·45 for the values used. In practice it is very close to this. The input resistance is high and was measured to be about 1 M$\Omega$. Both the input resistance and bias difference could easily be improved by using a higher $h_{fe}$ transistor for $Tr_1$. An improvement of about four times should easily be obtained.

The gain increased with frequency and was at least twice the low-frequency value of 10 MHz. A flat response was secured by adding the $RC$ circuit across $R_{C1}$. The circuit is a feedback one with three transistors in the feedback loop. It is thus potentially unstable. Theoretical design for stability is very difficult because it would require a detailed knowledge of all the transistor and circuit parameters up to 100 MHz or so, and even then would be very laborious. No difficulty was experienced in obtaining the required frequency response in the bread-board model but positive feedback symptoms were certainly present and it was felt that difficulties might well arise over

component tolerances. Further, the input resistance was lower than desired and although it could be made higher, it was doubtful if it could be made high enough.

The circuit is unquestionably an interesting one and it was abandoned rather regretfully because it was felt to be too subject to variation of performance from one amplifier to another. We may be wrong about this but we felt that we could not recommend its use until we had built 20 or 30 specimens to prove it. This was impracticable.

We, therefore, turned finally to an entirely different kind of circuit. It had been in our mind from the start, for it is an eminently designable circuit. It readily gave the required performance and its only fault is that it requires rather a lot of transistors, but they are inexpensive bipolar types. The development of this final amplifier will be treated in Part 3, and all component tolerances will be taken into account. In the main these tolerances have not been considered in this article because the procedure is rather tedious and one normally applies it only when a design is approaching finality.

## Appendix

Under signal conditions, as distinct from d.c.,

$$V_{in} = V_s + V_{gs} = V_{gs} + i_d R_s(1 + i_c/i_d) - i_L R_s$$

Now $V_{gs} = i_d/g_m$

and

$$\frac{i_L}{i_d} = \frac{R_s + (R_c + R_s) i_c/i_d}{R_c + R_L + R_s}$$

Therefore,

$$A = \frac{V_s}{V_{in}} = \frac{g_m R_L i_L/i_d}{1 + g_m R_s(1 + i_c/i_d - i_L/i_d)}$$

Now

$$\frac{i_c}{i_d} = \frac{R_D}{r_e + R_D/h_{fe}} = y$$

where $r_e = 0.026/I_C$ = emitter junction resistance. A little algebra then gives

$$A = \frac{g_m R_s(1+y)\dfrac{R_L}{R_c+R_L+R_s}}{1 + g_m R_s(1+y)\dfrac{R_L + R_c/(1+y)}{R_c+R_L+R_s}}$$

$$= \frac{g_m R_s(1+y)}{1 + g_m R_s(1+y)}\left[1 + \frac{R_c}{R_s}\cdot\frac{y}{1+y}\right]\frac{R_L}{R_L+R_o}$$

where $R_o = (R_c + R_s)\dfrac{1 + g_m \dfrac{R_c R_s}{R_c+R_s}}{1 + g_m R_s(1+y)}$

= output resistance.

If $g_m R_s(1+y) \gg 1$, $y \gg 1$ and $R_L \gg R_o$

$$A \approx 1 + \frac{R_c}{R_s}$$

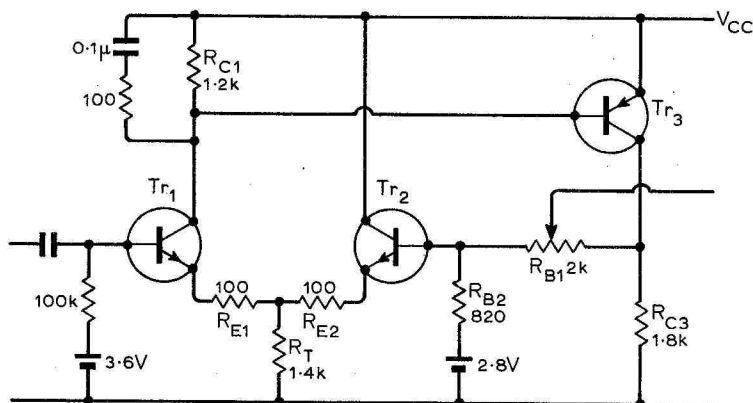If, also, $g_m \dfrac{R_c R_s}{R_c+R_s} \gg 1$

$$R_o \approx \frac{R_c}{1+y}$$



*Fig. 6. Three transistor circuit which gives a gain of about 3·5 times with an input resistance of at least 1 M$\Omega$ and, ideally, has zero current in the gain control.*

# News of the Month

## Far East hold on TV market tightening

Sales of U.K. manufactured colour television receivers to the trade jumped by 46% in the first half of this year compared with the same period last year; the respective figures being 278,000 and 191,000. As expected there was a fall in monochrome receiver sales during the same period amounting to 16% from 789,000 to 666,000. It will be interesting to see what effect the recent relaxation in H.P. restrictions and purchase tax will have on sales for the second half of the year.

Looking at the overall picture things are not so bright. In the first quarter of the year total sales of British-made television receivers showed a decrease of 16% over the same period last year, 401,000 (484,000). In contrast imported receivers are selling at treble the rate they did last year.

The importers increasing dominance of the radio receiver market again caused decreases in U.K. produced equipment for the first six months of the year, 323,000 (342,000).

These figures were provided by the British Radio Equipment Manufactuers' Association.

## If you can't beat them . . .

At a conference held in London recently, to discuss international harmonization of component standards, delegates from all nations present agreed that a world-wide agreement on standardization should be based on the system established by the Comité Européen de Coordination des Normes Electriques (CENEL). Delegations representating the following governments were present at the Conference: Belgium, Denmark, the Federal Republic of Germany, France, Italy, the Netherlands, Sweden, U.K. and the U.S.A.

Several nations have in the past few years set up various committees with the object of bringing national standards in line with international standards. Our own

BS 9000, based on the second report of the 'Burghard Committee' which was published in 1965, is fully compatible with the recommendations of the International Electrotechnical Commission (I.E.C.) of which we were one of the creators. The countries of the E.E.C. and E.F.T.A. got together to form CEN and CENEL and, in addition, the governments of France, West Germany and the U.K. formed a Tripartite Committee to discuss component standard harmonization.

At that time the American Electronic Industries Association attacked the Tripartite Agreement. Mr I. D. Secrest, executive vice-president, made the following statement: "The Tripartite Agreement creates an absolute embargo against exports of U.S. electronic components to the U.K., France and West Germany. The agreement is not yet fully implemented. There is time to prevent this blatant violation of U.S. rights under existing trade agreements from occuring if there is strong and determined action by the United States" (See *Wireless World*, July 1969, p. 303).

The action, we are pleased to say—to complete the heading . . . is to join them. Recently, two years after the E.I.A. outburst over the Tripartite Agreement, the I.E.E.E., gave its support to a proposed bill, S.1798, before the Foreign Commerce and Tourism Subcommittees of the United States Senate, the purpose of this bill is "to foster fuller U.S. participation in international trade by the promotion and support of representation of U.S. interests in international voluntary standards activities, and for other purposes".

Mr Sherr, manager of standards operations of the I.E.E.E., in commenting on the bill said that "it should provide a mechanism to allow professional societies to effectively carry out such activity [international standardization], an effort for which technical societies are best able to provide appropriately qualified manpower".

As the U.S.A. have now expressed the desire, and this bill will give them the means, perhaps we shall see, at last, some truly international component specifications.

## Two British i.c. plants to close

A cold wind blew through the i.c. industry recently when GEC Semiconductors

*The result of a recent fire at KEF Electronics packing and despatch department at Maidstone. The fire was thought to be caused by some kind of electrical failure in the roof and damage in excess of £80,000 was estimated. Within 24 hours of the disaster KEF were delivering stocks to a new temporary warehouse and a new cabinet assembly line is being set up with improved test facilities.*
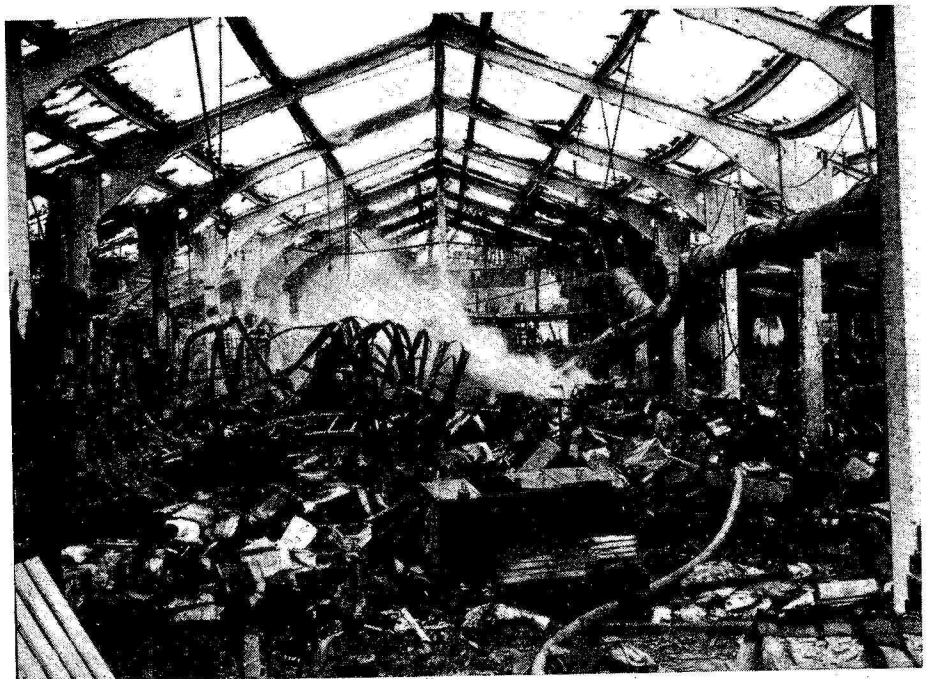
announced that it was to close two of its factories producing microcircuits. The closures were announced because increasing costs and falling prices led to heavy losses. One of the plants to be closed is at Witham, Essex; it has been open only two years and cost upwards of £2M to build. Also to be closed is a factory at Glenrothes in Scotland. GEC now intends to concentrate its microcircuit manufacture at the Hirst Research Laboratories at Wembley, Middlesex. A Witham engineer said that this does have advantages in that they will be in close contact with Hirst Labs where a good deal of semiconductor research is done and they will be able to use an 'in house' Myriad computer instead of having to rely on a rented terminal as they do at present. It is likely that we will see GEC pull out of standard i.cs, in which fierce competition exists, and concentrate its resources on custom designed i.cs which might result in an expansion of its m.o.s. activities.

A large American microcircuit manufacturer recently told *Wireless World* that the British i.c. industry is being killed by its own customers. In America and Germany, apparently, customers seem to be prepared to pay a reasonable price realizing that if the source of supply is not to dry up the manufacturer must have some profit margin, if only to recoup some of the development costs. According to the American manufacturer this does not apply in Britain and customers tend to beat down the price to rock bottom. This argument does not apply to such lines as t.t.l. where the manufacturers are waging a fierce price cutting war themselves to the customer's advantage, but usually, to their own disadvantage.

## Electronic clocks and watches

A number of semiconductor firms are actively engaged in research on all-electronic clocks and watches with no moving parts at all. Producing an electronic timing 'movement' is easy, but the real problem is how to display the information—one can hardly go round with four neon tubes strapped to one's wrist. In an attempt to solve this problem a great deal of work is going on with liquid crystals, but the potential market is so huge that few people are saying anything at this stage. One firm has said that the future of one of its entire manufacturing departments depends on achieving a successful and economic answer.

Motorola have produced a development prototype electronic clock that uses a crystal oscillator for timing and integrated electronics for division and display driving, which will be the standard pattern of things in all clocks and watches.

It employs 72 GaAs diodes for the display. There is an outer circle of 60 to display minutes and seconds and an inner circle of 12 for the hours. Two miniature batteries will last a year.

Another approach we have heard about is the use of micro-miniature stepping motors, consuming only a few microamps, which drive conventional hands. We can look forward to a great deal of interesting activity as microcircuit manufacturers strive to develop devices for the very large consumer market (watches, clocks, cars, plus who knows what) in an effort to stay in business. A company who can 'steal a march' on its competitors in this direction could reap rich rewards and perhaps use the extra income to finance unprofitable industrial device production lines.

## New hybrid resistor pastes

The Electrical Research Association (E.R.A.) has been trying to find materials which can be used as resistors for thick film hybrid microcircuits to replace the precious metals which are employed at present. Resistor pastes of precious metals are normally used because they retain their electrical conductivity after being fired in air. The electrical characteristics of a large number of materials are seriously affected by oxidization under these conditions.

Work of E.R.A. has shown that certain transition metal interstitials and some of their oxides retain their conductivity after being exposed to an oxidizing environment. Transition metals are those with the atomic numbers 22 to 30, 40 to 48 and 72 to 80 and an interstitial compound of these is one where atoms of small physical size (hydrogen, boron, carbon, nitrogen, oxygen, etc) are situated in the interstices of the parent metal lattice.

E.R.A. have successfully made resistor pastes with molybdenum boride and are now proceeding to find other materials with better performances and which are easy to process.

The reason for the behaviour of the transition metal interstitials is not fully understood, but E.R.A. think that the interstitial material may act as a reducing agent on the transition metal counteracting the oxidizing effect of the atmosphere during firing.

## Facsimile transmission to police cars

The Home Office and Bristol Constabulary are co-operating in an experiment to discover the value of transmitting documents from headquarters to police vehicles using the v.h.f. radio system. Ten vehicles have been fitted with facsimile receivers connected to the normal mobile radio installations. The system is capable of transmitting documents of unlimited length but only 108mm wide such as sketches, maps, typescript, photographs, etc.

## Tall buildings v microwave links

Post Office engineers are carrying out a series of tests to find out what effects tall buildings have on microwave links and how these effects can be calculated. A large number of factors are involved including the position of the building relative to the microwave link and the height, shape and materials used to construct the building. A helicopter has been fitted with a 9.4GHz radar modified by the Radio and Space Research Station for the job.

A ground receiver picks up a direct signal from the helicopter and the signal which has been reflected by the building under investigation. By altering the position of the helicopter it is possible to measure the building's radiation pattern. At Romford one building produced a reflection which was only 8dB down on the direct signal; enough to cause severe interference.

## Ideas catalogue

A directory of computer programmes for solving scientific problems is available from Peter Peregrinus Ltd (P.O. Box No. 8, Southgate House, Stevenage, Herts, SO1 1HQ) following an agreement with Science Associates International (New York). The directory, called 'Computer programmes in science and technology', enables information to be obtained on how others have used a computer to solve particular problems.

## Heatsink court case

Marston Excelsior Ltd has won a court action, under the design copyright act, against Waycom Semiconductors Ltd and Advance Electronics Ltd. The case concerned the manufacture of extruded aluminium heatsinks which were registered as Marston Excelsior model 10D. The court order restrains Waycom and Advance from manufacturing heatsinks to this design and instructs these companies to surrender to Marston Excelsior the heatsinks which infringe the copyright. In addition related drawings, catalogues etc., have to be destroyed.

# Letters to the Editor

*The Editor does not necessarily endorse opinions expressed by his correspondents*

## F.E.T. audio oscillator

The design by Mr. A. J. Ewins of his f.e.t. audio-frequency oscillator in *Wireless World* for March, 1971, was most interesting.

The appended circuit may prove of further interest as the simple, economical arrangement gives extremely good results.

The direct-coupled amplifier has its quiescent operating bias conditions set by adjustment of the preset resistor $R_6$. Initially this is adjusted to give half the supply voltage at the emitter of $Tr_3$. Ultimately this control can be further adjusted for minimum distortion from the oscillator providing that a suitable distortion measurement instrument is available.

The amplifier has moderate gain but low distortion when the overall negative feedback loop via the thermistor is open. This is due largely to the local inverse feedback circuits in the individual stages. The thermistor feedback loop is then relieved from controlling large and violent variations in gain due to transient conditions such as range switching or rapid tuning dial excursions.

Jitter accompanying frequency adjustment, familiar (and annoying) in most thermistor controlled *RC* oscillators, is considerably improved by the above means. This improvement is also assisted by the fact that only one *RC* time constant

$R_{53}$-$C_2$, is present in the negative feedback loop. The combination gives an oscillator substantially free from tuning jitter.

There may have been good reasons for the choice of a $\sqrt{10}$ tuning ratio in Mr. Ewins' design, but it is generally more useful for a 10:1 frequency range to be available. Using a 450 pF double-gang variable capacitor and 22MΩ resistor for the lowest range, frequencies from 15 to more than 150 Hz can be generated. For the other four ranges the resistors are progressively reduced in decade steps so that the top range of 15 to 150kHz employs 2.2k resistors.

The frame of the variable capacitor must be insulated from earth as it is connected to the gate of the f.e.t. The tuning capacitor, range switch and associated resistors are vulnerable to hum and other stray field pick-up and thus should be carefully positioned. It is preferable to locate these components within a shielded compartment which should, however, not add too greatly to the stator-to-earth capacitance of the tuning capacitor. This would limit the highest frequency attainable on each range. The stator-to-earth stray capacitance and the input capacitance of the f.e.t. should in any case be compensated by adding a trimmer of similar capacitance across the top section of the tuning gang.

The oscillator described has a range

of 15 Hz to more than 1.5 MHz covered in five decade ranges. It has an output of 1V r.m.s. at low impedance. Total harmonic distortion, measured at random spot frequencies on each range, using a Hewlett Packard 333A Distortion Analyser, was between 0.05 and 0.09%.

V. R. KRAUSE,
Johannesburg,
South Africa.

## Ceramic pickup equalization

While I would endorse Mr Burrows' conclusions in his article 'Ceramic pickup equalization', part 1 (July issue), I would like to point out that his efficiency calculation is not valid. He has calculated the input power to the cartridge by multiplying the r.m.s. velocity by the component at 45° of the tracking force. Unfortunately he has overlooked the fact that there is no net work done against the tracking force, because the work done against the force by the groove modulation on one half-cycle is returned on the next half-cycle. The tracking force is merely holding the stylus in the groove and has nothing to do with the cartridge input power.

If this is difficult to visualize, imagine a mono signal. This is at 90° to the tracking force and therefore can do no work against it, but yet there is still an output from the cartridge.



Fig. 1.
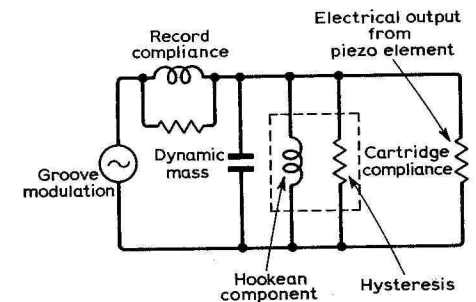
What the groove modulation 'sees', in terms of mechanical impedance, is the dynamic mass of the stylus assembly, the compliance (or rather the stiffness) of the cartridge, and the resistance to movement which is converted into electrical output. The question is whether or not the latter is significant in comparison with the former two.

The compliance can be resolved into two parts; the 'Hookean' component in which displacement from the mean position is proportional to the force applied, and the 'hysteresis' (or damping) component, in which displacement from any position is a function of applied force and change of applied force. A simplified electrical analogue of the system is shown in Fig. 1.

No work is done against the dynamic mass or the Hookean component of compliance, because the forces and velocities involved are 90° out of phase, e.g., stylus acceleration is zero when the velocity is maximum, and vice-versa. Therefore, the only work done against the compliance is against the hysteresis part. Forces and
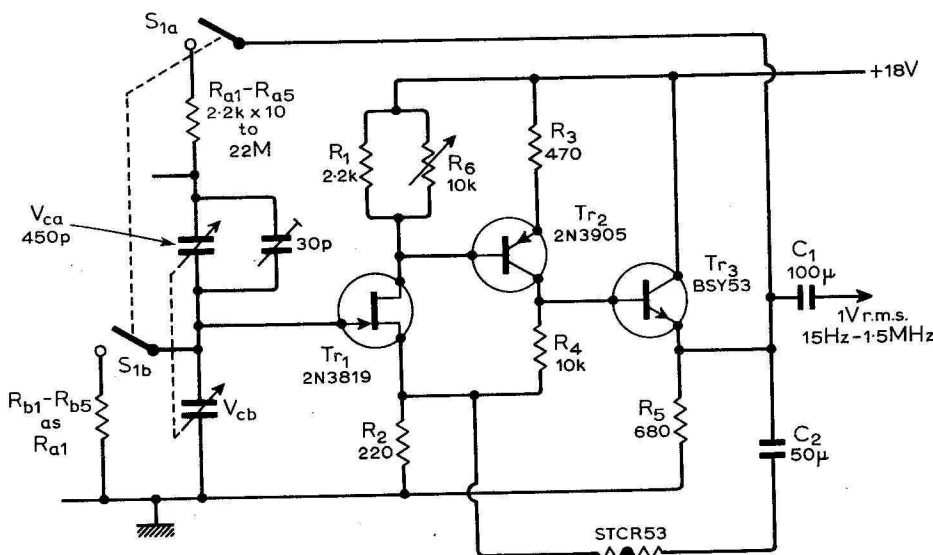
velocities associated with dynamic mass and Hookean compliance result only in 'reactive' dyne cm, no actual ergs of work being done. Without knowing details of the hysteresis, the efficiency cannot be calculated, but what can be done is to compare the electrical output in ergs/sec with some of the 'reactive dyne cm/sec'.

For a Deram, the stiffness (1/compliance) at 45° is about $0.16 \times 10^6$ dynes/cm. At 1kHz and 20 cm/sec r.m.s. velocity, the r.m.s. force required to overcome the stiffness, assumed Hookean, is 500 dynes, resulting in 10,000 'reactive dyne cm/sec'. Assuming the stiffness to be Hookean results in the minimum number of dyne cm/sec for the given value of stiffness.

Now the maximum output from a Deram under these conditions is $1.1\mu$W into 270 k$\Omega$. This is $1.1 \times 10^{-6}$ J/sec or 11 ergs/sec, so taking maximum power from this cartridge has a similar effect on the damping to connecting a resistor taking 11W across a tuned circuit involving 10 kVA, i.e. a very small effect indeed!

Unfortunately, if we regard a cartridge as a series of black boxes, we conclude that the effect of loading is dependent on the characteristics of the last black box (i.e. the piezoelectric element in the case of a ceramic cartridge) and its coefficient of coupling with the previous black box, rather than on the efficiency of the whole system. Presumably manufacturers realise this and ceramic cartridges are independent of loading, not inherently but as a result of design.

H. C. MIRAMS,
Bradford,
Yorks.

*The author replies:*
I was interested to read Mr. Mirams' comments on the efficiency calculation, and he has rightly pointed out that the basis of the calculation was not sufficiently well explained to be rigorous. He is right in saying that there is no net work done against the tracking force (i.e. power $= \bar{v} \; \bar{F}$ and not $\bar{v} \times \bar{F}$) but of course there was nothing in the article to infer that the calculation was made assuming it to be a cross product.

Mr. Mirams is, I am sure, mistaken in believing that the tracking force has nothing to do with the cartridge input power. On the contrary, the tracking force is a good measure of the force necessary to keep the needle in contact with the groove walls (mono or stereo) and is therefore a *direct* measure of the lateral forces on the needle. So by knowing the tracking force, one knows the force necessary to keep the stylus in the groove at maximum modulation velocities and the calculation was performed for this case.

Taking Mr. Mirams' second point that no work is done against the dynamic mass or the Hookean component of compliance, this is of course true, work is done when there is hysteresis or damping present. I think where I would disagree is in the relative magnitude of the resistive and reactive components of the needle tip mechanical impedance. In the vicinity of 1kHz this impedance is mainly resistive and was assumed wholly resistive for the calculation. It is

mainly resistive here because at around 1kHz the two reactive impedances (i.e. the compliance and the dynamic mass) cancel out as in a series tuned circuit at resonance; so, to avoid a high $Q$ resonant system, considerable damping has to be added to the pickup system.

Mr. Mirams' very simplified electrical analogue—which rather confusingly used capacitance as the analogue of mass, where it is conventional to use inductance as the analogue—should be compared to Mr. S. Kelly's electrical analogue as published in *W.W.* December 1969 which shows the number of damping elements present in a conventional stereo ceramic pickup.

Finally with reference to Mr. Mirams' last paragraph, it can be seen that this point was made in the original article under the heading 'Reasons for low efficiency'.

I certainly endorse Mr. Mirams' conclusion that ceramic pickups are independent of loading by design, but the main point which I tried to put over in the article is that this is a natural outcome from making an aperiodic transducer, and is not achieved by special design effort separate from the essential one of achieving non-frequency-dependent action.

B. J. C. BURROWS.

# Diagnosis of logical faults

I read with great interest the first part of 'Diagnosis of Logical Faults' by R. G. Bennetts (July issue) and readers may find the following comments pertinent. The circuit used to illustrate the various

methods is reproduced in this letter as Fig. 1. This circuit was used as data for one of our standard programmes, CLOIS*, that generates automatically a multi-flow testing procedure from the circuit description using the fault matrix method. The resulting testing procedure is shown in Fig. 2. It can be observed that five tests are used to detect all faults, namely $t_4$, $t_1$, $t_3$, $t_2$ and $t_7$ instead of the four tests that Mr. Bennetts suggests. The differences arise from the fact that when a node has a fan-out of greater than one Mr. Bennetts does not consider any extra faults whereas CLOIS inserts more faults. For instance input (a) is connected to $G_1$, $G_2$ and $G_3$ and a fault could be that gate $G_2$ is not connected to node $C_1$ but it is connected to logical 0 instead. By considering such faults another ten faults can be introduced into' matrices $F$ and $G_D$. However, the problem can also be reduced considerably as Mr. Bennetts later suggests. By deleting identical rows matrix $F$ needs only thirteen rows including the extra faults due to fan-out. The two extra rows are:

(1) $f_{17}$—connection to $C_2$ of gate $G_1$ s-a-l which is identical to connection to $C_1$ of gate $G_2$ s-a-0, and

(2) $f_{18}$—connection to $C_1$ of gate $G_3$ s-a-0.

The extensions to the $F$ and $G_D$ matrices are shown in Figs. 3 and 4. From Fig. 4 it can be seen that fault $f_{18}$ is detected only by $t_4$ and therefore $t_4$ is also an essential test; $t_4$ also detects the presence of fault $f_{17}$.

This demonstrates that the minimal detection test set of Mr. Bennetts is only minimal for the particular faults he considered and that some simple faults are not detected by such a test set.
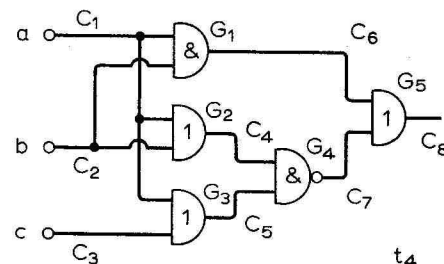
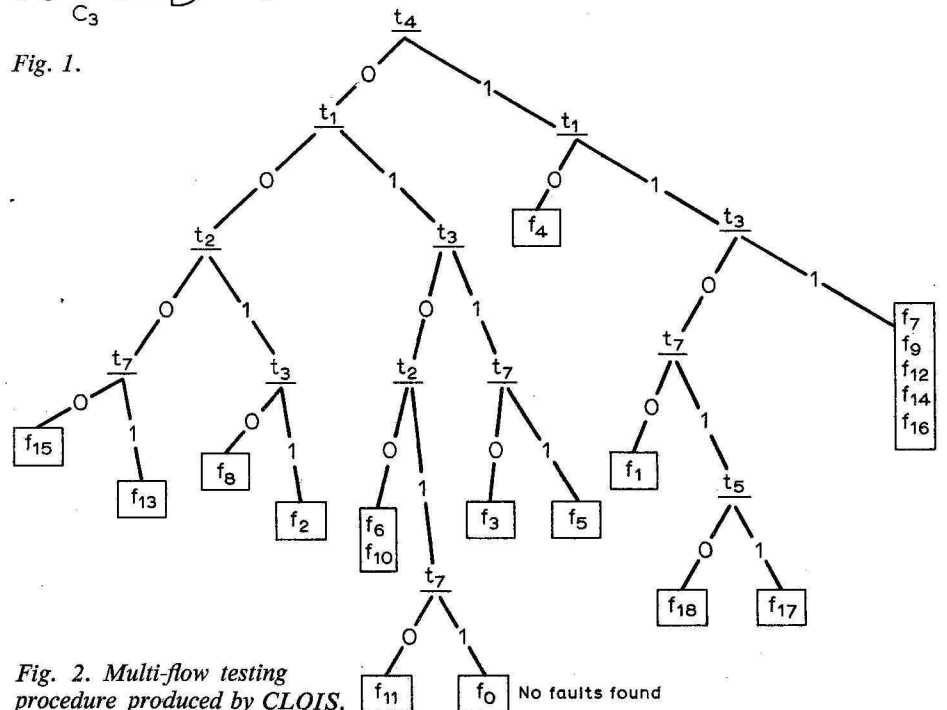*Boyce, Emmerson, Stringer and West: 'Simulation of binary logic circuits by digital computers', *The Marconi Review,* Vol. XXXIV, No. 181, 2nd. Quarter 1971, pp. 121-142.



Fig. 1.



Fig. 2. Multi-flow testing procedure produced by CLOIS.

| Test | $f_{17}$ | $f_{18}$ |
|---|---|---|
| $t_0$ | 1 | 1 |
| $t_1$ | 1 | 1 |
| $t_2$ | 1 | 1 |
| $t_3$ | 0 | 0 |
| $t_4$ | 1 | 1 |
| $t_5$ | 1 | 0 |
| $t_6$ | 1 | 1 |
| $t_7$ | 1 | 1 |

*Fig. 3. Extra columns for matrix F.*

| Test | $(f_0 f_{19})$ | $(f_0 f_{18})$ |
|---|---|---|
| $t_0$ | | |
| $t_1$ | | |
| $t_2$ | | |
| $t_3$ | | |
| $t_4$ | 1 | 1 |
| $t_5$ | 1 | |
| $t_6$ | | |
| $t_7$ | | |

*Fig. 4. Extra columns for matrix $G_D$.*

It may be of interest to mention that for this example CLOIS took eight seconds to compile the circuit data and fifteen seconds to generate the multi-flow testing procedure. The computer used was an ICL 4/70.

There are two further points in the article which are misleading. The extension of the fault matrix method to produce multi-flow testing procedures does not require as much storage as Mr. Bennetts mentions. The CLOIS programme used two matrices $F$ and $G_D$ of sizes $8 \times 14$ and $8 \times 13$, respectively, as compared to Mr. Bennetts' $8 \times 136$.

During the discussion of the path sensitizing method for fault $f_1$ it is found that $t_4$, $t_5$ and $t_7$ all can detect the presence of $f_1$. It is then suggested that $t_4$ (or $t_5$) is used as this detects seven faults. If columns had been ignored if they are identical with previously entered columns then he would argue that $t_4$ and $t_5$ detect three faults each and $t_7$ detects four faults, therefore he would have picked $t_7$ instead of $t_4$. The resulting minimal set would then have been $t_7$, $t_3$, $t_2$, $t_1$. It was just by chance, although fortunate, that his minimal set by the path sensitizing method is the same as that obtained by CLOIS.

The final point is that the footnote on p.327 should be

$$n\,[(n + 1)/2] = (n^2 + n)/2$$

A. H. BOYCE,
Research Division,
Marconi Company,
Great Baddow,
Essex.

*The author replies:*
It would appear that Mr. Boyce has misunderstood the purpose of the article. It was written as a tutorial introduction to digital circuit fault diagnosis and was not intended to be an exhaustive treatise

—indeed, if he now reads Part II published last month, he will see that I have in fact referred the reader to not only a more general review paper but also separate papers for each technique.

Returning to Mr. Boyce's comments in detail, he is of course absolutely correct in considering separate faults on fan-in/fan-out lines—a point I made in the footnote to col. 3 p.326.

I believe also that Mr. Boyce has confused 'multi-flow testing procedures' (defined on p.326) with the formation of the $G_L$ matrix. The theoretical maximum for this matrix is, as stated, $8 \times 136$. This is given by $^{n+1}*(2)$ and although some reduction based on indistinguishable fault sets can be effected, it is inconceivable that this would reduce to the same size as the $G_D$ matrix. What Mr. Boyce has in fact done, is to create a partition based on the $F$ and $G_D$ matrices, i.e. he has combined the two techniques of fault matrix and partitioning to derive a detection test set in a similar manner to that indicated by me in Part II (compare the form of his Fig. 2 with my Fig. 11).

The times quoted by Mr. Boyce for deriving the test set using CLOIS are interesting. The approach at Southampton University is based primarily on the Boolean Difference technique and for the circuit in the article the programme takes 11 seconds to accept a topological description, derive the Boolean expression and then proceed to manipulate this to eventually generate a detection test set. This is using the University's ICL 1907 computer.

One final point. The equation in col. 1, p.327 should read:

$$t_i f_j + t_i f_k = 1$$

R. G. BENNETTS.

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*This is an alternate form (used by Kautz incidentally) for $^{n+1}C_2$ and unfortunately, the printer thought I had omitted the dividing line.

## Sonic scanning for tubeless TV

It was with interest that I read the article, 'Sonic scanning for tubeless TV' by J. J. Belasco, in the July issue. It reminded me of the work done some 10 years ago by Stephen Yands.

A similar flat device was built by him that utilized sonic scanning to display video information on an electroluminescent phosphor (see: 'A solid-state display device', *Proc I.R.E.* Vol. 50, No. 12, Dec. 1962). Basically, it consisted of a piezoelectric ceramic sheet covered with an electroluminescent phosphor, and sandwiched between a transparent viewing electrode, and a ground electrode. This was scanned by launching travelling elastic waves into the piezoelectric material. A spot could be produced by launching two travelling waves orthogonally, and selecting the increased amplitude at the intersection by a discriminating medium. By varying the relative

timing between the orthogonal waves, the spot could be made to scan a raster.

To my knowledge, although crude Lissajous figures were displayed on such a panel, it never reached the stage of producing acceptable TV images. This panel was a continuous sheet of piezoelectric material since provision was made for two-dimensional sonic scanning. The one-dimensional sonic scanning array of 625 horizontal strips proposed in the article might possibly provide a better solution. However, the complexity of the number of interconnections and transducers could prove to be a stumbling block.

G. O. TOWLER,
Broomfield,
Chelmsford,
Essex.

## Broadcasting frequencies

I should like to endorse, with one reservation, the sentiments of Mr. Higham's remarks on B.B.C. medium-wave broadcasts ('Letters', June issue).

The bad reception, owing to East German and Albanian interference, and phase distortion, renders intolerable reception in many parts of the country. The proffered alternative, the f.m. service, is always 'loud and clear'—but a weakness lies in the poor choice of programmes provided. For example, on one occasion recently, tuning into v.h.f., there was only one programme (jazz) to listen to. Radio 2 was being relayed on Radio 3, and Radio 4 had closed down. This broadcast occupied no fewer than nine frequencies in Band II together with two a.m. outlets; a grand total of 11 simultaneous broadcasts! Three of the above frequencies were those of B.B.C. local radio stations—which relay from Radios 1-4 on average 60-70% of their broadcasting time. One wonders what could be less local than the relaying of national programmes.

Possibly, the long-awaited 'shot in the arm' for the B.B.C. could well lie in the creation of healthy competition with the promised commercial radio services.

However, I must condemn the concept of 'pirating' any odd frequency to hand. This is the law of the jungle, and causes interference. What is needed—after 23 years—is a complete re-appraisal of the broadcasting plan by the countries involved, and a new scheme drawn up. Following this, coupled with the new commercial stations, sound broadcasting could have a very bright and interesting future.

STEFAN WORONIECKI,
Lancing,
Sussex.

# Circuit Ideas



## Level-conscious trigger system

Schmitt triggers can be coupled together to make a channel selector governed by input signal amplitude. Although shown for d.c. triggering, adaptation for a.c. operation is possible. In Fig. 1 three Schmitt circuits are set to trigger at different voltage inputs. As shown, the higher trigger voltage will also trigger the circuits requiring lower trigger voltages. Fig. 2 shows inhibit feedback current circuits. These are used to short circuit the unwanted outputs as shown in Fig. 3. Diodes.
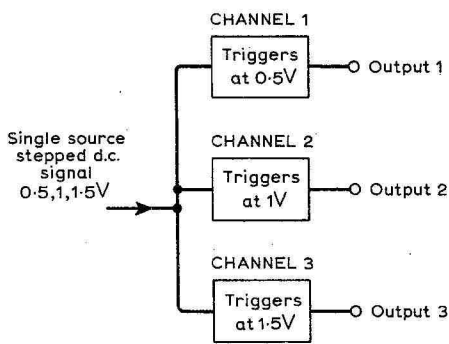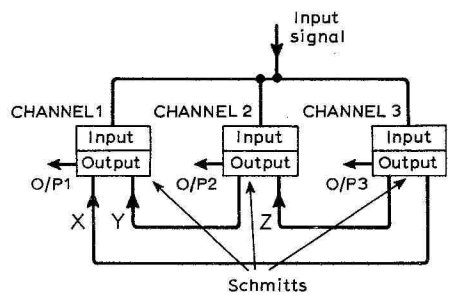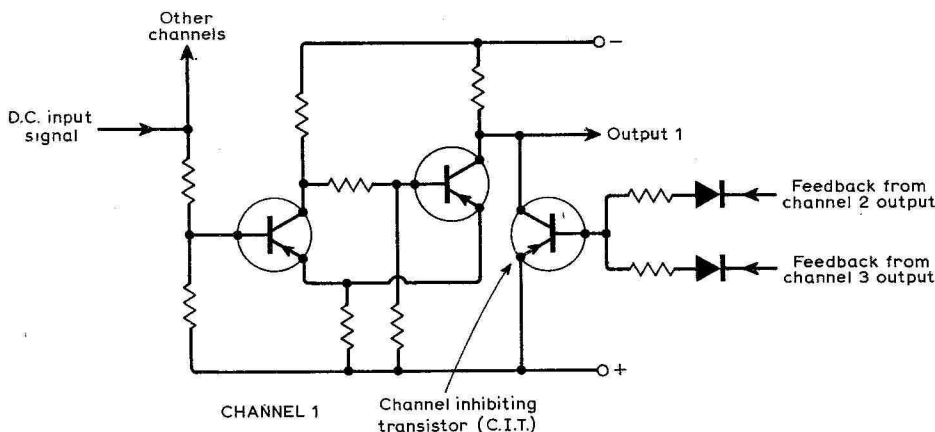


**Fig. 1**



**Fig. 2**



**Fig. 3**

are required to isolate the output circuits of channels 2 and 3 from each other. Signal differentiation greater than 0.2V can be achieved with careful trigger design.

A. R. BIDWELL,
West Molesey,
Surrey.

## High input-impedance Schmitt trigger

The need for a high input-impedance trigger circuit is quite common and the usual approach involves using a field effect transistor as a buffer for a bipolar transistor Schmitt or an i.c. comparator. Designs using a junction f.e.t. or m.o.s.f.e.t. in both stages of the Schmitt are not common due to the wide spreads and low mutual conductance of these devices. Recently silicon-gate field-effect transistors have become available with threshold voltages of 1 to 2V. This spread is sufficiently low to enable the conventional Schmitt circuit to be used. In the circuit shown a silicon-gate pair (M1202, G.E.C. Semiconductors) is used in a standard Schmitt configuration. The circuit differs from normal bipolar transistor practice in only two respects. The resistance values are an order of magnitude higher to allow for the lower mutual conductance of the field-effect transistors and a series input resistance is provided to limit the forward current of the internal protection diode of the M1202. The series resistance is necessary if the input signal is allowed to have a positive polarity with respect to ground. For a negative-going signal the input current to the Schmitt is typically less than 100pA. The input current
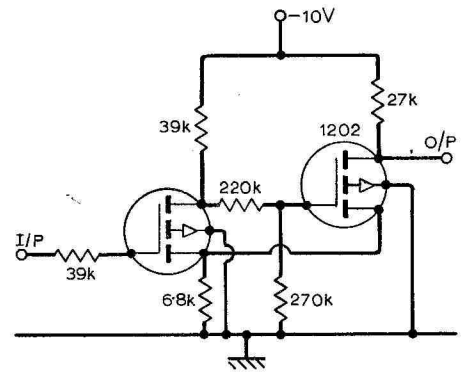
is due to the reverse leakage current of the M1202 protection diode. The low threshold voltage of the silicon-gate transistors enables the circuit to operate from supply voltages as low as 5V. With the supply voltage and resistance values shown the circuit provides an upper trip point of 4V, a lower trip point of 3.1V and rise and fall times of less than $1\mu s$.

J. A. ROBERTS,
J. DRISCOLL,
Witham,
Essex.

## Wirewound 'log' pot

Carbon-track potentiometers when used as volume controls often have a very limited life and develop 'intermittents' and crackles. Wire-wound controls are much better in this respect, but unfortunately only linear laws are commonly available, and these are not suitable for faders. An approximation to a
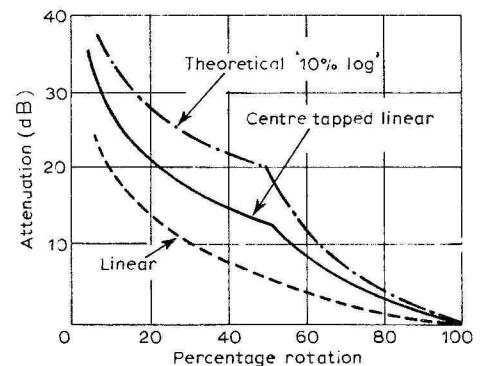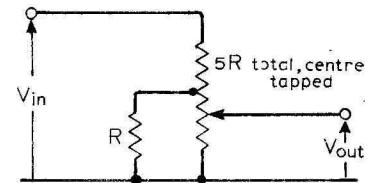




logarithmic law can be obtained by using the arrangement shown. The wirewound track can be centre-tapped quite easily in cheap controls by taking the back off and exercising some ingenuity! Although a better approximation to the '10% log' law could be obtained, (the kink in the curve shows up as a noticeable jump in the sound when doing a fade) the present arrangement seems to be the best compromise.

D. C. HAMILL,
Southampton.

# Frequencies for Space Communication

## World radio conference in Geneva

*by D. E. Baptiste**

The first administrative radio conference to allocate frequencies for space telecommunications was held by the International Telecommunication Union in 1963, only six years after the original sputnik first orbited in space. That conference successfully provided frequencies and the necessary technical and regulatory provisions to enable Intelsat to come into being as a commercially viable organization. The facility by which hundreds of millions of the world's population have seen the Olympic games and the various Apollo missions on television has become so familiar in a relatively short time that it is easy to forget that radio communication through outer space was unknown a decade ago.

Apart from communication satellites, such as Intelsat and the Russian Molniya, there have been meteorological satellites, satellites used for space research, and the use of space techniques by amateurs. The 1963 conference also provided additional frequencies for Radio Astronomy.

### Second space conference

The rapid operational and technical development of these space services and the possibility of using satellites for new services made it necessary for the Administrative Council of the I.T.U. to convene a further world administrative conference. Its main purpose was to provide more frequencies for existing space systems (like Intelsat) to allow for growing traffic needs for international telephone and telegraph traffic and the relaying of television programmes; and for the growing needs of other services such as space research, radio astronomy, meteorological satellites, amateurs and the aeronautical satellite service. Furthermore frequencies were needed for new satellite services: maritime-mobile, broadcasting and earth exploration. In addition the conference had to draw up the necessary technical provisions to enable the new frequency allocations to be used successfully; and to provide regulatory procedures for co-ordination between administrations and the notification and recording of frequency assignments.

*Head of the Radio Regulatory Division, Ministry of Posts & Telecommunications, and leader of the U.K. delegation to the Geneva conference.

The conference assembled, with over 700 delegates from 100 countries, at the Palais des Expositions, Geneva, on the 7th June. In attendance there were the usual officials of the I.T.U. and a sprinkling of observers of the United Nations and other interested international organizations to see fair play. The conference got off to a good start under the experienced chairmanship of Gunnar Pedersen, Director General of the Danish P.T.T., who had also been chairman of the 1963 conference. No time was lost in breaking down into committees and thence into working groups, so that in a matter of days delegates were deep in discussion on the main aspects.

From the technical point of view it was essential to get down to an early examination of the technical criteria for sharing between space systems and terrestrial services and for sharing between various space systems so that the delegates concerned with frequency proposals, particularly proposals for new services, should know what was practicable. There were other delegates, concerned with regulatory procedures—co-ordination between administrations, notification and recording of assignments etc.,—who needed to know what technical factors should influence their thinking. This was not a one-way process. As the conference progressed there was inter-action between the frequency and regulatory committee and the technical committee. In addition there was the main task for the technical committee of considering specific technical proposals from administrations in the light of the preparatory work of the Special Joint Meeting of the C.C.I.R. held in Geneva in February/March 1971; and necessary revision of the technical provisions of the Radio Regulations.

The frequency committee and its main working groups broke down the many frequency proposals of administrations into subject matter. The most important task was to find more space for the communication-satellite service (to be known as the fixed-satellite service). The conference recognized that the frequency spectrum up to 10GHz was so crowded that there was no scope for introducing more wideband space services. The first relief bands to provide for the next

generation of fixed satellites were therefore found between 10.95 and 14.5GHz. The importance of a band below 15GHz is that it is not so affected by rain in temperate climates as frequencies higher up the spectrum. In Region I, which includes western Europe, these bands consisted of three separate 250 MHz segments at the lower end, mainly for use in the space-to-earth direction, and one 500MHz band (14-14.5MHz) in the earth-to-space direction. The apparent imbalance between the down and the up bands is explained by the fact that different down bands might be employed in working to an inter-continental satellite between the Americas and Europe from those needed for a European satellite occupying a different arc of the sky. Two of the down bands are also allocated in the earth-to-space direction so that they can be used for feeding broadcasting satellites. This makes for maximum use of the spectrum.

The next relief band—for the third generation of fixed-satellites—was found between 17 and 31 GHz. A total of 2,500MHz of space each way (1000MHz shared with terrestrial services and 1,500MHz exclusive) was provided. At these frequencies local rain storms can blot out reception. It will be necessary to provide more than one earth station at each terminal, separated by sufficient distance from each other, to avoid this hazard. These frequencies are therefore likely to be used only for high cost intercontinental traffic of the Intelsat type.

### Allocations up to 275GHz

The present Radio Regulations allocate frequencies only up to 40GHz. The spectrum above that is affected by the earth's atmosphere so that communication between earth and space is not generally practicable. There are, however, some exceptions to this rule inasmuch as at certain frequencies there are windows in the atmosphere that permit communication. The conference provided allocations in these windows for fixed-satellites (a total of 32GHz) stretching from 40 to 275GHz. In addition frequencies were allocated for space-to-space links, (over 50GHz) on the space side of the atmospheric fence, away from these

windows. Although these frequencies are not likely to be brought into use within the next 10 years, both the U.S.A. and Japan stated they were working on satellites which would use them. It was important that the conference should fix the allocations so that system design could proceed.

Some countries had a need for a small allocation of frequencies for fixed satellites at around 2000MHz to enable a satellite system carrying a small traffic load to be used in sparsely populated regions, like Alaska and the Yukon, where there is no existing terrestrial network to conflict with the earth stations. The conference found two small frequency bands 35MHz wide in the band 2500-2690MHz for this purpose outside Region I and provided safeguards for countries whose terrestrial systems might be affected.

### Space research and radio astronomy

Additional frequency space was provided for space research and radio astronomy ranging from a small 20kHz radio astronomy band at 21MHz right up to a band for Radio Astronomy and Space Research at 230-240 GHz. Of particular concern to the U.K. was a U.S. proposal for space research in the important 1750-2290MHz band. The American requirement for an additional 185 MHz in the up direction and 90 MHz in the down direction could have played havoc with this band, which is heavily used in Europe for public telecommunication radio-relay services. There is a fundamental technical need for frequencies for the penetration of deep space to be kept below 2300 MHz. The conference recognized this but kept the frequencies out of Region I except for 85MHz allocated to Spain. This effectively limits the location of the one high-power station required in Europe to the country in which it is at present located.

### Maritime mobile satellites

For the first time frequencies were allocated to the maritime mobile satellite service. A small allocation, in the v.h.f. band used for international shipping, was made for safety and distress purposes. Two bands, 7.5MHz in each direction, were allocated between 1535 and 1660MHz, with two small bands (1MHz each) for combined use by maritime and aeronautical mobile satellite. This should provide a satisfactory service for the larger ocean-going ships. It is not likely to be introduced before 1978 but would provide welcome relief for the congested and unsatisfactory h.f. band. It could provide a reliable high-quality 24-hour-a-day service integrated into the public automatic telephone network.

### Aeronautical mobile-satellites

The conference allocated two 15-MHz bands for use by aeronautical mobile-satellites for civil aircraft in the 1535-1660MHz band. This should provide adequate frequency space for the development of satellite communications for aircraft.

### Earth-exploration

This is a new type of satellite service including
(a) the meteorological satellite, controlled from the U.S.A. but giving information to world weather forecasting centres, one of which will be in the U.K.; and
(b) other earth-exploration satellites used for obtaining information about the earth-mineral resources, land and sea use, detection of agricultural diseases, atmospheric and water pollution, etc. The information is obtained by satellites from sensors on the earth or in the air and relayed to earth stations.

Frequencies for all these uses were allocated by the conference.

### Broadcasting satellites

Frequencies were allocated for the first time for this service in which distinction was drawn between individual reception, requiring very high powers, and community reception in which relatively low powers would be needed. The latter is important as the conference would accept the use of broadcasting satellites in certain bands only on the basis that community reception would be used.

The conference accepted the use of broadcasting satellites in the television u.h.f. band between 620 and 790MHz, subject to agreement among administrations concerned and affected, and laid down a stringent power limitation to protect the terrestrial broadcasting receivers of other countries. The interest of western European countries was to avoid interference from satellites in this band with their extensively developed broadcasting networks.

Band 2500-2690MHz was allocated to broadcasting satellites for domestic and regional systems for community reception only, with power limits to protect terrestrial services of other countries. This should be the main band for developing countries and sparsely populated territories in advanced countries where a terrestrial broadcasting network would be too costly.

The main band for broadcasting satellites for use by advanced countries in western Europe will be from 11.7-12.5GHz. This 800MHz has been allocated in Region I on an equal primary basis to broadcasting satellites, broadcasting, fixed and mobile (except aeronautical mobile) services. The conception is that there should be a frequency assignment planning conference as soon as practicable. At this conference the countries of Europe could decide how much of the 800MHz should be devoted to European or regional coverage and how much to national coverage; for example, 800MHz is wide enough to enable each country in western Europe to have four programmes, because at these frequencies very narrow beams can be used and channels

can be repeated at suitable distances. A new footnote in the Radio Regulations provides that the terrestrial services will be in effect on a secondary basis to the broadcasting satellite service during the frequency planning process so as not to inhibit the planning. Once the plan has been settled, countries will know what frequencies remain outside the channels allocated to them and neighbouring territories for broadcasting satellites. These can then be planned on a national basis for their terrestrial services. Broadcasting satellite channels can be exploited in the first instance for community reception and later used for more powerful satellites giving individual reception to homes when this becomes technically and economically feasible.

The conference also allocated frequencies for broadcasting satellites higher up the spectrum, at 22.5-23GHz (Region 3 only) 41-43GHz and 84-86GHz. But these are for long-term study and development rather than for use in the foreseeable future. As regards broadcasting satellites generally, the technical and regulatory constraints prevent broadcasting to other countries without their consent.

### Amateur satellites

The conference agreed to the use of satellites by amateurs in the h.f. bands allocated exclusively to amateurs on a world-wide basis (7, 14, 21 and 28 MHz) and one higher band at 24-24.05 GHz. But the most useful allocation was at 435-438 MHz which can be used in conjunction with the existing 144–146 MHz band.

### Summary

To sum up, the conference, which concluded its six weeks sitting on 17th July, allocated all the important frequencies needed for the continued growth of the Intelsat system for the foreseeable future and beyond; for the new European system if it is required; and provided frequencies for use by new services with adequate safeguards to terrestrial services where safeguards are needed. The revised Radio Regulations will come into force on 1st January 1973.

# Elements of Linear Microcircuits

## 11: F.M. radio receivers

*by T. D. Towers,* M.B.E.

The electronics design engineer working in the domestic radio field is turning away from discrete transistors to the numerous special-purpose linear i.cs which are now available. However, it is evident that an i.c. for domestic radio application must meet quite a number of special constraints.

● It must be lower cost to the set manufacturer than discrete-component assemblies.

● Must be capable of being 'second sourced'.

● Its throw-away value must not be too high to permit economic servicing.

● Reliability should be higher than discrete assemblies.

● It should be able to work over widely different voltage rails (which usually means internal voltage regulator stages).

● Current consumption should be as low as discrete designs because dry-battery operation is often required. (This can conflict with the different voltage rail requirement.)

● It should be designed for easy handling, testing, installation and removal.

Before the linear i.c. arrived, a.m./f.m. set manufacturers had already had experience of block modules made with discrete components in the Mullard 'LP' range (LP1169/79 f.m. tuner blocks and LP1164/65, 1170/71 a.m./f.m. i.f. blocks). As a result, they had already solved some of the assembly problems involved in changing over from traditional separate component assemblies to the use of functional assemblies—which is after all what i.cs are.

### Partitioning a.m./f.m. receivers

Different manufacturers adopt different approaches to the problem of how to divide up receiver functions for the separate i.c. packages required to make up the set. Until some degree of standardization is reached all we can do at this stage is to look at some typical examples.

If you are interested in the detailed problems of partitioning f.m. domestic radios, you will find a useful discussion of the topic in 'A.M./F.M. monolithic receivers' by P. E. Hermann, L. H. Hoke,

* Newmarket Transistors Ltd

R. L. Petrosky and R. Wood (of Philco-Ford) in *I.E.E.E. Transactions on Broadcast and Television Receivers,* July 1968, Vol. BTR-14, No. 2 pp. 95-103.

Initially set designers tried to use general purpose professional linear i.cs (such as the $\mu$A703 and MC1550) for domestic receivers, but were unsuccessful because they were too costly.

Next, industry turned to developing special i.cs for high-performance professional f.m. applications, such as the RCA CA3076 10.7MHz high-gain amplifier limiter and the CA3075 amplifier limiter detector. These could be integrated into excellent high-gain f.m. systems but the assembly costs could not compete with

conventional discrete transistor assembly in domestic f.m. sets. (A full description of the CA3075/6 and their applications can be found in 'High-performance integrated circuits for high-gain f.m.-i.f. systems' by R. T. Peterson in *I.E.E.E. Transactions on Broadcast and Television Receivers,* Nov., 1970, Vol. BTR-16, No. 4 pp 257-263.)

Another interesting development that pointed the way to current practice was the Fairchild set of i.cs $\mu$A717, 718, 719, and 720. These were all the same basic monolithic chip with different internal metallizing interconnection patterns which produced devices for various television, f.m. and a.m./f.m. applications. You can find more detail of these in 'Novel
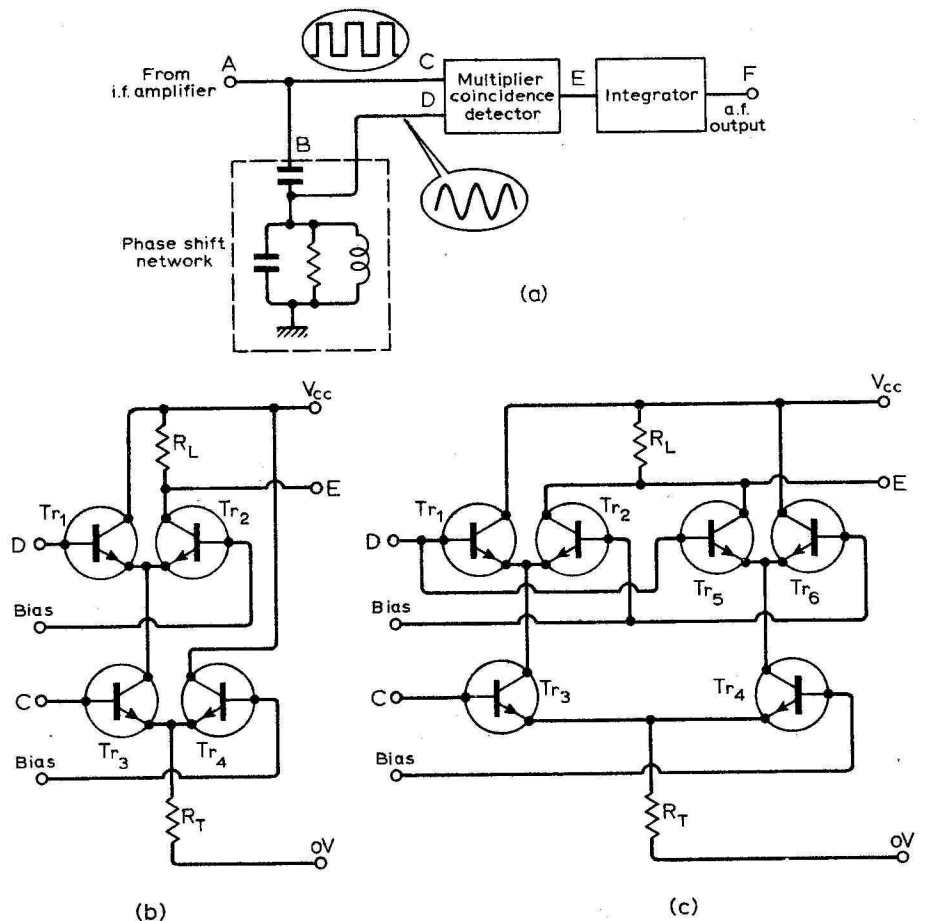


Fig. 1. Coincidence (quadrature) f.m. detector system suited to i.cs in f.m. receivers; (a) simplified block diagram; (b) practical i.c. realization of half-wave detector; (c) full-wave detector.
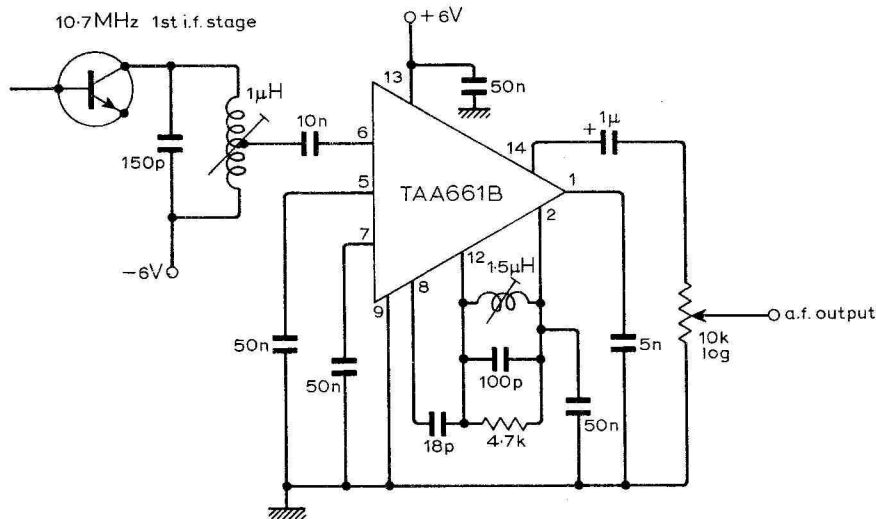
*Fig. 2. Typical circuit using the TAA661 f.m. i.f. amplifier and coincidence detector.*

multi-purpose l.i.cs introduce new concepts into circuit design' by David Bingham in *I.E.E.E. Transactions on Broadcast and Television Receiver,* July, 1967, Vol. BTR-13, No. 2 pp 108-115.

## Detection

Over the years many different types of f.m. detection have been used. Most of them, such as the 'Fremodyne' (single detuned *LC* circuit drive to detector), the 'Travis' (two *LC* circuits detuned on each side of i.f.), and the 'super-regenerative' detector, have fallen out of favour. With discrete transistors, the two systems with the widest commercial use are the Foster-Seeley discriminator (common in the U.S.A.) and the ratio detector (common in Europe). Neither of these is ideally suited to monolithic i.cs because they require carefully tuned balanced *LC* circuits. With i.cs they are tending to be replaced by the coincidence (quadrature)

detector requiring only one tuned *LC* circuit; by the phase-locked-loop detector, dispensing with inductances altogether; and by the diode-pump detector (also inductorless).

The diode pump or pulse-counter detector is attractive because it is so easy to set up, but to be really efficient it calls for a low intermediate frequency, around 100kHz, which tends to rule it out for low-cost domestic receivers.

The coincidence detector appears to be preferred by most designers for 10.7MHz/i.f.,f.m. detection with i.cs. Fig.1 illustrates its working. In Fig.1(a), the 10.7MHz signal, built up to a square wave in a preceding limiting amplifier, is fed into terminal A. From A it passes direct to one terminal C of the coincidence multiplier in one direction and it is also split off into a second channel B which contains a single tuned circuit (externally connected to the i.c.), the action of which restores the 10.7MHz square wave to sine-wave form at terminal D. Thus both the square wave

and the sine wave are fed to the multiplier circuit. The signal frequency modulation varies the instantaneous frequency of both signals and, since the sine wave is subjected to a phase displacement due to the action of the tuned circuit, the coincidence detector produces an output at E consisting of a series of pulses of mean value proportional to the modulation frequency. Thereafter the integrator (a capacitor shunting the output resistance of the coincidence detector) recovers the audio from the f.m. r.f. signal and provides the necessary de-emphasis or top cut (European time constant $50\mu$s, American $75\mu$s). In i.c. form the detector multiplier circuit can provide half-wave or full-wave detection.

In Fig. 1(b), a half-wave detector, the average value of the output current in $R_L$ is proportional to the frequency deviation of the input signal. The full-wave version (more complex, but less affected by noise) is given in Fig.1(c) and uses three, instead of two, pairs of differential long-tail transistors, but is similar in action.

The coincidence detector is becoming popular with i.cs in f.m. sets because the setting up of the detector involves the adjustment of only a single external coil, while giving performance similar to the more traditional, but more difficult to set up, Foster-Seeley and ratio detectors. Besides decreasing assembly and alignment time, the coincidence detector reduces the number of external passive components required.

## TAA 661

One example of an i.f. amplifier using a coincidence detector is the SGS TAA661 which incorporates 25 transistors and 18 resistors in a single silicon chip. It is housed in a 14-lead dual-in-line package and includes a three-stage limiter amplifier, an f.m. detector and an emitter-follower audio buffer pre-amplifier, with an internal voltage regulator circuit permitting operation on rail supplies from 6 to 18V. How simply it can be used in practice is demonstrated in Fig.2 which shows the practical circuit for taking the output from a discrete component 10.7MHz f.m. first i.f. amplifier stage and delivering a.f. to the volume control.

## TBA 690

The TAA661 is for f.m. only. Some domestic receivers covering f.m. also incorporate a.m. This points to a different line of i.c. development characterized by the Mullard TBA690. This i.c., in a 16-pin plastic dual-in-line package, comprises the functions within the shaded area of Fig.3, and can be seen to contain everything except the f.m. front end, the f.m. first i.f. (which can be switched to operate as a mixer oscillator on a.m.) and the f.m. and a.m. detectors. The integrated audio amplifier in the TBA690 can provide 500mW into an 8 $\Omega$ speaker on a 9V battery, although the supply can be anything from 4.5 to 9V. The quiescent current drain on the battery is only 22mA; this is comparable with discrete device
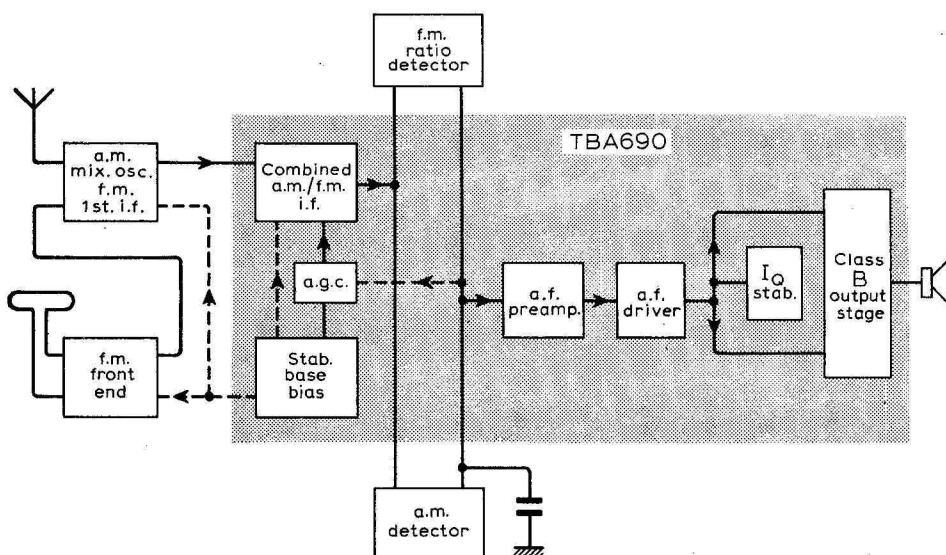


*Fig. 3. How the Mullard TBA690 integrated circuit incorporates in a single package all stages of an a.m./f.m. portable 9V receiver except the f.m. front end, the a.m./f.m. input stage and the a.m. and f.m. detectors.*

designs. (A companion i.c., the TBA700, can operate from 4.5 to 12V on the rail and at 12V can supply an output of 1.5W into 8 $\Omega$ .)

It is not immediately evident from Fig.3 that the TBA690 does not of itself supply the selectivity provided in a discrete-component receiver by the r.f. front end tuned circuits and the fixed tuned i.f. circuits. In the arrangement of Fig.3, the r.f. selectivity is provided by the separate external input blocks, and the i.f. selectivity is provided by lumped $LC$ or ceramic filter circuits between the a.m. mixer oscillator /f.m. first i.f. and the i.c. input. Equally the two detectors call for external tuned $LC$ transformers. [A new i.c. has just been announced (CA3089, see Literature Received) which combines the following functions: i.f. amplifier, quadrature detector a.f. pre-amp,. with outputs for a.g.c., a.f.c., muting (squelch) and tuning meter.—Ed.]

## Stereo decoders

One area where monolithic i.c. techniques lend themselves is in stereo decoders. An example of this is the Siemens TBA450. Of the three standard decoder systems (matrix, switch and envelope), the TBA450 uses the matrix decoding system outlined diagrammatically in Fig.4. The output from a standard f.m. front end is taken after the detector, but without de-emphasis applied, and fed into three filters which separate the M (mono L + R) signal below 15kHz, the S (stereo L-R) signal from 23-53kHz, and the pilot signal at 19kHz. The pilot signal is frequency doubled to 38kHz and then controls a synchronous a.m. demodulator while the M and S signals are matrixed to give independent outputs of the stereo left and right audio signals. The same system is employed in the TBA450. In this circuit the bandpass amplifiers are active filters which do not use inductances.

## Phase-locked-loop

The phase-locked-loop technique referred to in the last article on a.m. receivers offers a way of avoiding the fixed tuned i.f. filters of the f.m. receiver. Fig.5(a) shows the functional p.l.l. sections in the Signetics NE561B linear integrated circuit, which will provide a demodulated audio output if fed directly with the 10.7MHz output from a conventional f.m. mixer without any 10.7MHz tuned circuits. The tuning element in the circuit is a voltage controlled relaxation oscillator whose frequency is determined by non-inductive components. The oscillator is designed so that the operating frequency can be varied over a limited range by a d.c. bias voltage. If the oscillator is rough-tuned near to the 10.7MHz and its output is applied to the phase comparator, the comparator will give an output determined by the frequency and phase deviation of the v.c.o. from the input signal. This comparator output is amplified and filtered and fed back round the loop through the limiter to adjust the
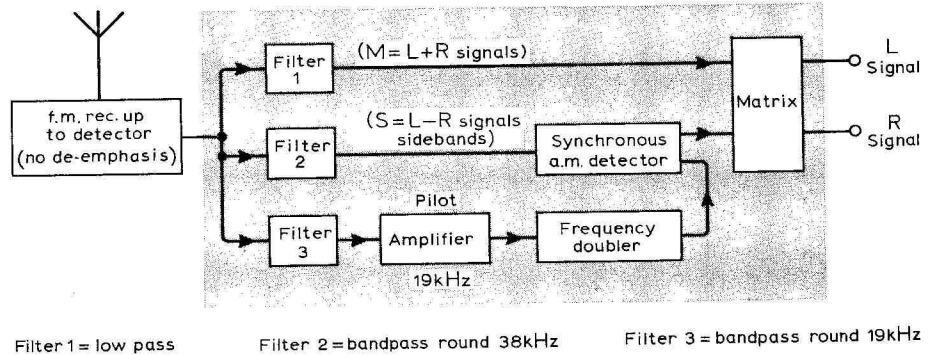


Fig. 4. Stereo decoder integrated circut block diagram of matrix decoder system.

v.c.o. frequency to bring it into frequency and phase step with the f.m. input. Thus the oscillator tracks the input signal and produces a strong continuous signal even if the input is discontinuous or noisy.

So far the circuit has produced a cleaned up, greatly amplified, copy of the input f.m. signal without using inductances. But the main interest of the circuit to us is that within the phase comparator loop an output signal has been obtained which is dependent on the carrier shift. Apart from its use to lock the v.c.o. onto the carrier, it also represents the

audio output of the demodulation system, because the amplitude of the loop control signal is proportional to the carrier frequency deviation . . . which is just the f.m. modulation. This enables the NE561B to be set up in a simple system such as Fig.5(b) to replace the complete 10.7MHz i.f. strip up to the f.m. detector. As yet, phase-locked-loop i.cs operating directly at the f.m. broadcasting frequencies around 100MHz are not practicable with existing monolithic technologies, but as the art develops it is possible that the local oscillator too can be dispensed with.
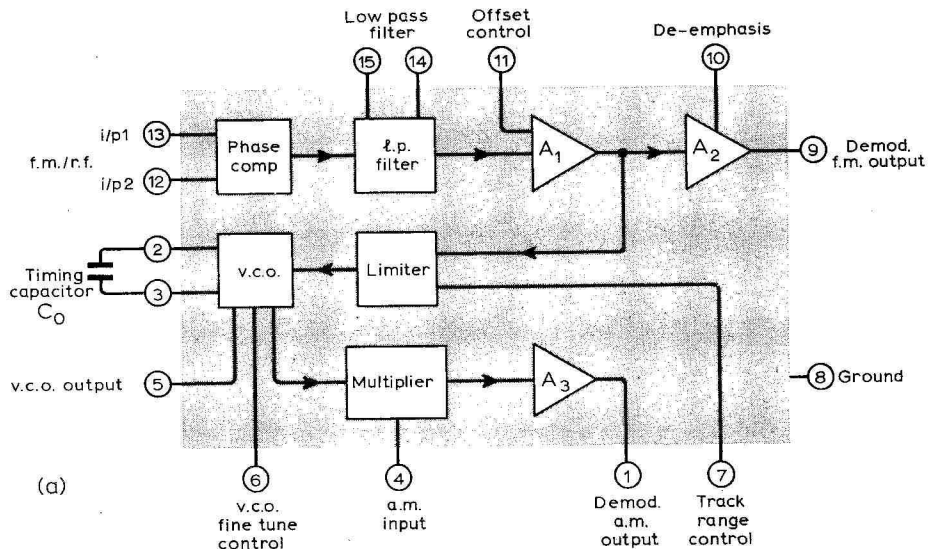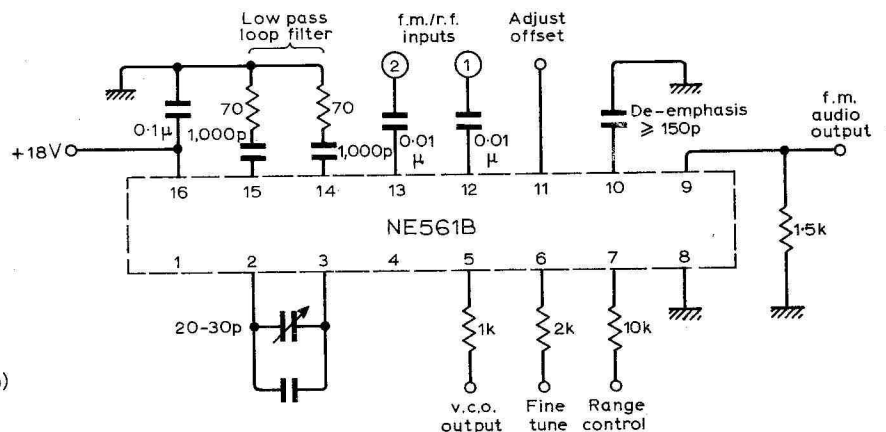


Fig. 5. Phase-locked-loop integrated circuit (Signetics NE561B) enabling complete inductorless substitute for conventional f.m. i.f. strip plus f.m. detector. (a) Internal functional sections of i.c., (b) connection as 10.7MHz a.m. /f.m. i.f. strip up to volume control.

The multiplier feeding the amplifier $A_3$ in Fig.5(a) is an additional a.m. detector that enables the NE561B to be switched to a.m. to provide an a.m./f.m. system. For this the a.m. is fed directly at the broadcast frequency into terminal 4 and the v.c.o. locks onto the carrier; the system providing detected audio output at terminal 1. [In last month's issue an article described a complete p.l.l. stereo decoder on a single chip which is manufactured by RCA.—Ed.]

### Digital synthesis

Another approach to eliminating tuned circuits with i.cs is the digital frequency synthesizer, which combines the phase-locked-loop with a master crystal oscillator. Frequency synthesizer circuits have been designed that generate the required local oscillator frequencies for an a.m. and f.m. broadcast receiver. Selection of a station is accomplished by positioning switches to indicate the station's frequency. Fine tuning is not necessary. The receiver will not (for all practical purposes) drift, because the local oscillator is crystal controlled. Low-cost medium-scale integrated circuits (m.s.i.) are the building blocks of this. For a detailed account of this design, you should consult J. Stinehelfer and J. Nichols' 'A digital frequency synthesizer for an a.m. and f.m. receiver' in *I.E.E.E. Transactions on Broadcast and Television Receivers*, October 1969, Vol.BTR-15, No.3, pp 235-243.

### Thick film hybrid

Instead of packaging i.c. chips in multilead packages and supplying them to set manufacturers to mount with passive components, such as resistors and capacitors, on a printed circuit board, we are already seeing a logical development in which semiconductor manufacturers are themselves mounting the chips in thick film hybrids with the passive components to complete their functions printed and fired on the ceramic substrates. This produces a neat microcircuit suitable for plugging into sockets on the printed circuit board (which now tends to become merely an interconnection network between the microcircuits and the large non-integratable components) and will remove many of the servicing problems found with discrete components or even linear monolithics soldered into position. With new subminiature i.f. transformers, about 5mm square, it is now possible to mount them directly onto thick film hybrids.

### Conclusions

Much is happening in the application of i.cs to domestic f.m. receivers, and developments are taking place along several different lines at once. It is difficult to see how things will finally develop, but in the not-very-distant future we can expect to find the set makers indicating a preference which will show itself by semiconductor manufacturers beginning to 'second source' some items.

*(The concluding article in this series will deal with i.cs in television receivers.)*

# Sixty Years Ago

September 1911. Our predecessors on the staff of *The Marconigraph* devoted much space to the social implications of the ever-increasing acceleration of the technology machine. Sometimes the only intention was to make technical reports more readable. Today this is seldom done because the average engineer is bombarded with so much printed material he has time only to glance at a small fraction of it and to read even less. The change in approach is emphasized if one reads (if time allows) early technical articles.

For instance, in a report on the massive radio station at Cape Cod about half a page was devoted to the antics of two dogs kept at the station and the rescue of one of the animals, who had been caught in a trap, was described in detail.

The Cape Cod station was used to transmit the daily news to ships in the Atlantic and had the advantage of an automatic morse transmitter using paper tape input. Apparently once the huge generators were started the noise of the spark transmitter was 'terrifying' and the spark itself could be seen as a flickering light fifteen miles away.

At the receiving end on board a ship the transmission was recorded on paper tape and it was reported that a female passenger who said she understood all, after being shown around the wireless installation, wanted to know how the paper tape went from shore to ship without getting wet!

# Conferences and Exhibitions

*Further details are obtainable from the addresses in parentheses*

**LONDON**
Sept. 1-3                                    Imperial College
**Artificial Intelligence**
(British Computer Soc., 29 Portland Pl., London W1N 4AP)
Sept. 6-10                                    City University
**Electrical Network Theory**
(I.E.E.E. Symposium, c/o The City University, St. John St., London EC1V 4PB)
Sept. 8 & 9                                    Savoy Pl., W.C.2
**High Voltage Insulation in Vacuum**
(Inst. Phys., 47 Belgrave Square, London S.W.1)
Sept. 13-17                        U.S. Trade Center, S.W.1
**U.S. Electromechanical & Electronic Components**
(U.S. Trade Center, 57 St. James's St, London S.W.1)
Sept. 20-24                              St. Katherine's, E.1
**Control and Instrumentation Exhibition**
*(Control & Instrumentation, 28 Essex St., London W.C.2)*

Sept. 28-Oct. 1                              Savoy Pl., W.C.2
**Centralized Control Systems**
(I.E.E., Savoy Pl., London WC2R OBL)

**BRIGHTON**
Sept. 7-10                              University of Sussex
**Human Locomotor Engineering**
(I.Mech.E., 1 Birdcage Walk, London S.W.1)
Sept. 8-10                              University of Sussex
**Electron Mean-free Paths in Metals**
(Inst. Phys., 47 Belgrave Sq., London S.W.1)

**CARDIFF**
Sept. 17-19                              University College
**Physics—From School Through Higher Education**
(Inst. Phys., 47 Belgrave Sq., London S.W.1)

**CRANFIELD**
Sept. 1-5              Cranfield Institute of Technology
**Business and Light Aviation Show**
(ITF-Iliffe Exhibitions Ltd., 1-19 New Oxford St., London WC1A 1PB)

**LANCASTER**
Sept. 14-16                                    The University
**Solid State Devices**
(Inst. Phys., 47 Belgrave Sq., London S.W.1)
Sept. 23 & 24                                    The University
**Data Processing and Display for Inspection Purposes**
(Inst. Phys., 47 Belgrave Sq., London S.W.1)

**LOUGHBOROUGH**
Sept. 7-10                        University of Technology
**Displays**
(I.E.E., Savoy Place, London WC2R OBL)

**MANCHESTER**
Sept. 1-3                                    The University
**Multivariable Control System Design and Applications**
(U.K.A.C. 1971 Convention Secretariat, Savoy Pl., London WC2R OBL)

**SHEFFIELD**
Sept., 7-9                                    The University
**Computers in Medical and Biological Research**
(I.E.E., Savoy Place, London WC2R OBL)

**SWANSEA**
Sept. 1-8                                    University College
**British Association Annual Meeting**
(B.A., 3 Sanctuary Buildings, 20 Gt. Smith St., London S.W.1)

**TEDDINGTON**
Sept. 22 & 23                        National Physical Lab.
**High Voltage Electron Microscopy**
(Inst. Phys., 47 Belgrave Sq., London S.W.1)

**OVERSEAS**
Sept. 1-3                                    Sendai
**Antennas and Propagation**
(Dr. K. Nagai, Inst. of Electronics and Communication Eng., Kikai-Shinko-Kaikan Bldg., Shiba Park 21-1-5, Minato-ku, Tokyo 105)
Sept. 4-12                                    Milan
**Radio-TV Show**
(Associazione Nazionale Industrie Elettrotecniche ed Elettroniche, Via Donizetti 30, Milan)
Sept. 10-19                                    Amsterdam
**Firato Electronics Exhibition**
(RAI Gebouw N.V., Europaplein 8, Amsterdam)
Sept. 13-19                                    Budapest
**Micronica 71—Electronic Component Show**
(Micronica 71, Budapest 5, P.O. Box 454)
Sept. 19-23                                    Chicago
**Electrical/Electronics Insulation**
(E. A. Boulter, G.E.C., 1100 Western Ave., West Lynn, Mass. 01905)
Sept. 21-23                                    San Diego
**Engineering in the Ocean Environment**
(G. K. Tajima, Bissett-Berman Corp., 3939 Ruffin Road, San Diego, California 92112)
Sept. 23-25                                    Washington
**Broadcast Technical Symposium**
(R. M. Morris, 60 Sunset Lake Rd., RD1, Sparta, N.J. 07871)
Sept. 27-29                                    Turin
**Elettronica '71—Conference on Applications of Electronics in Industry**
(Elettronica 71, Corso Massimo d'Azeglio 15, 10126 Torino)

# The Liniac

## A new linear inverting and amplifying circuit and some other applications of low-level Darlington transistors

*by J. L. Linsley Hood*

One of the most interesting of recent developments in the discrete semiconductor components field has been the use of integrated circuit techniques to provide small-signal Darlington-connected transistors of the general form shown in Fig. 1(a). A suggested symbol is given in Fig. 1(b), and this is used in the remainder of this article.

While it is practicable to construct Darlington pairs from separate transistors if the collector current of the second transistor is fairly large, at the sort of current levels typically employed in small signal circuitry it is much more difficult. If the second transistor has, say, a current gain of 400 and a collector current of 0·5 mA, the collector current of the first device must be less than 1·25 $\mu$A, and at this order of collector current the current gain of most normal discrete small-signal transistors is very low, and their other characteristics are also impaired.

When, however, a monolithic Darlington transistor is made, the junction areas and doping levels of the input transistor are adjusted so that it will function effectively at a very low collector current. Also, because of the very low collector-to-input base capacitance, it is possible to obtain good performance at moderately high frequencies, even with high dynamic impedance

collector loads, which give high stage gain values.

Ideally, a low-level amplifier element should have a high input impedance, a relatively low output impedance, a high gain, a low noise level, should be linear, should be simple and tolerant in its power supply requirements. The normal (bipolar) junction transistor does not meet the input and output impedance requirements at all well, and in addition is intrinsically non-linear as a voltage amplifying element, so that it is almost essential to arrange stages in cascade with substantial amounts of overall negative feedback to remedy these defects. However, on consideration it is apparent that the non-linearity of the bipolar transistor is an input characteristic effect, and for any given base-emitter circuit impedance is directly related to the magnitude of the input signal voltage. Within limits, the output signal swing is unimportant in this respect. It follows from this that for any given output signal level, the higher the gain of the stage the better its linearity will be. The monolithic Darlington transistor offers a satisfactorily high input impedance with a very high value of current gain, and if an arrangement can be found in which this can be induced to give a high voltage gain the major circuit requirements will have been met. Moreover, such a stage will be phase inverting which is very convenient for a number of applications, whereas the conventional transistor feedback pairs of Fig. 2 are non-inverting systems.

### Methods of obtaining high stage gain

Several techniques are available for increasing the stage gain of a conventional transistor amplifier. However, some of these are unhelpful in preserving the linearity of the system, and the principal remaining technique is to employ a collector load which has a dynamic impedance substantially larger than its d.c. resistance. This could be a "bootstrapped" load resistor, an "active" (i.e. signal dependent) load, or a constant-current source. Of these arrangements the third is by far the most straight-forward and free from side-effects, and such a constant-current load can be provided by the use of a conventional junction field-effect transistor, for which the circuit required, as shown in Fig. 3, is simplicity itself. The characteristics of this arrange-

ment are shown in Fig. 4 for various values of the source resistor $R_1$.

Since the dynamic resistance of such a system is, effectively proportional to the reciprocal of the slope of the drain-current/drain-voltage graph (i.e., the flatter the higher) it can be seen that there are conditions when this dynamic impedance is very high, and it could then be employed as the load in the collector circuit of a transistor amplifier stage. This would give a very high
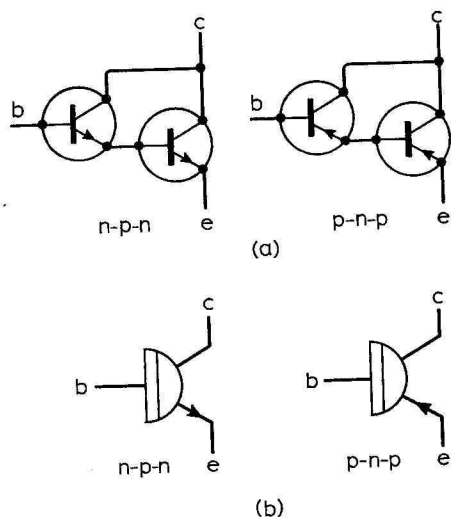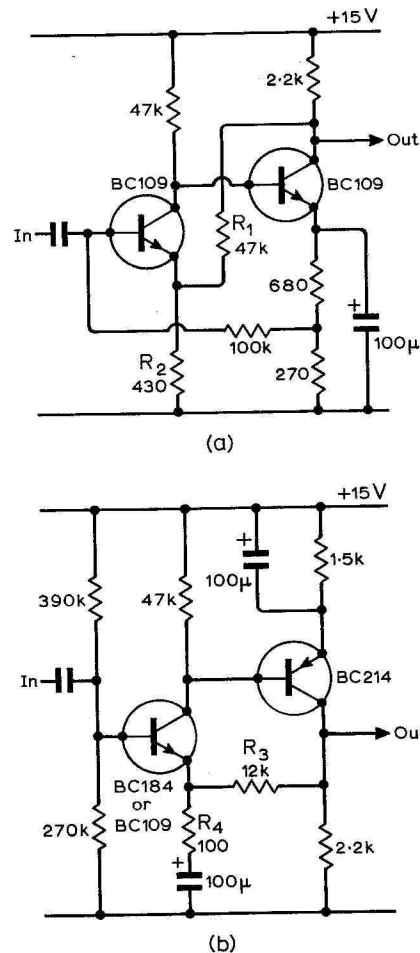
(a)

(b)

*Fig. 2. Feedback stabilized non-phase-inverting transistor pairs. (a) n-p-n/n-p-n feedback pair. Gain depends on $R_1$, $R_2$ (as shown $M \approx 100$). Input impedance $\approx 68\,k\Omega$. Open loop gain $\approx 2000$.*
*(b) n-p-n/p-n-p pair. Gain depends on $R_3$, $R_4$ (as shown $M \approx 100$). Input impedance $\approx 50\,k\Omega$. Open loop gain $\approx 2000$.*

(a)

(b)

*Fig. 1. (a) Darlington transistor arrangements; (b) suggested symbol for monolithic Darlington devices.*

stage gain while still allowing a reasonable value for the collector current, and a convenient range of voltage drop values across the load. Moreover, by the suitable choice of f.e.t. or source resistor the collector current of the amplifying transistor can be precisely defined, which is frequently an advantage.

## Circuit conditions for high stage gain

The stage gain of a simple single-stage transistor amplifier is given by the formula.

$$M = \cfrac{1}{h_{re} - \cfrac{h_{ie}}{Z_L}\left(\cfrac{1 + h_{oe}.Z_L}{h_{fe}}\right)}$$

If the terms $(h_{ie}.h_{oe} - h_{fe}.h_{re})$ are written as $\Delta h_e$, the so-called "$h$ determinant" for the common emitter configuration, this equation simplifies to

$$M = \frac{h_{fe}.Z_L}{\Delta h_e.Z_L + h_{ie}}$$

and if $\Delta h_e$ is sufficiently small, as is mostly the case, this approximates to

$$M \approx \frac{h_{fe}.Z_L}{h_{ie}}$$

If the dynamic value of $Z_L$ is large, and the input impedance of the amplifier transistor is small the stage gain can be very
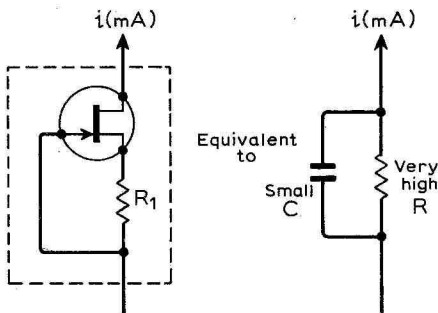
Fig. 3. Constant current load using f.e.t.— i depends on f.e.t. and value of $R_1$. Dynamic impedance in range $200\,k\Omega$–2 M$\Omega$.

high. (However, $h_{ie}$ depends on the collector current of the transistor, and increases as this is reduced. For this reason, high gains normally require both a certain minimum of collector current and also a drive impedance which is small in relation to $h_{ie}$.)

As will be seen from Fig. 4, an f.e.t. will act as a high dynamic impedance constant-current source even when the source resistance $R_1$ has zero value, provided that the source-drain voltage exceeds what is known as the "pinch-off" voltage, which is typically two or three volts. The current which will flow in this condition (zero source-gate bias) is known as the $I_{DSS}$ and will depend on the device. For f.e.ts such as the 2N4302 and the 2N5457 this will be in the range 1–3 mA—a convenient value of collector current at which to operate a typical small signal Darlington amplifier stage. When such a transistor amplifier is employed with an f.e.t. collector load it is found that stage gains of the order of 2500 to 5000 can be
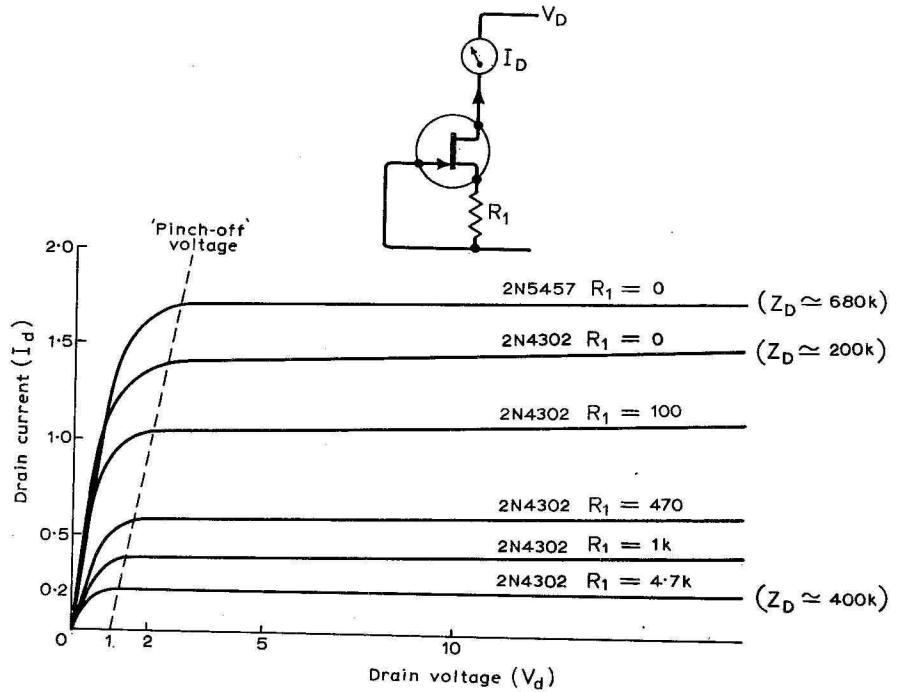
Fig. 4. Drain current characteristics for sharp cut-off f.e.ts.

obtained, even with source impedances of the order of 100 k$\Omega$ or more.

It will be appreciated that an amplifier stage of this type using a high dynamic impedance collector load will have an output impedance which is so high that the shunting effect of almost any external load would lead to a serious reduction in gain. To complete the practical circuit, therefore, an output emitter follower is required, and this can with advantage be a further monolithic Darlington transistor, although in practice a normal high-grain small signal transistor may be nearly as good and somewhat cheaper.

The final form of the proposed high gain circuit combination is shown in Fig. 5(a), and for convenience as a "shorthand" form in Fig. 5(b). This circuit arrangement has been found to be very versatile as a rela-

tively low-frequency amplifier stage, and to possess a number of useful qualities as a phase-inverting circuit element, and the name "liniac" (linear inverting amplifying circuit) is suggested for this configuration.

## Liniac circuit characteristics

General considerations. In its simplest form, the liniac consists of a bipolar transistor connected as a grounded-emitter amplifier, an f.e.t. used as a constant current load, and an output emitter follower. If the output circuit impedance is fairly high, say 10 k$\Omega$ or greater, this can be a normal small-signal transistor such as the BC109 or BC184. Also, if a source resistor is used with the f.e.t. of a value sufficient to reduce the load current to some 10–50 $\mu$A (at which level the dynamic impedance is extremely high)
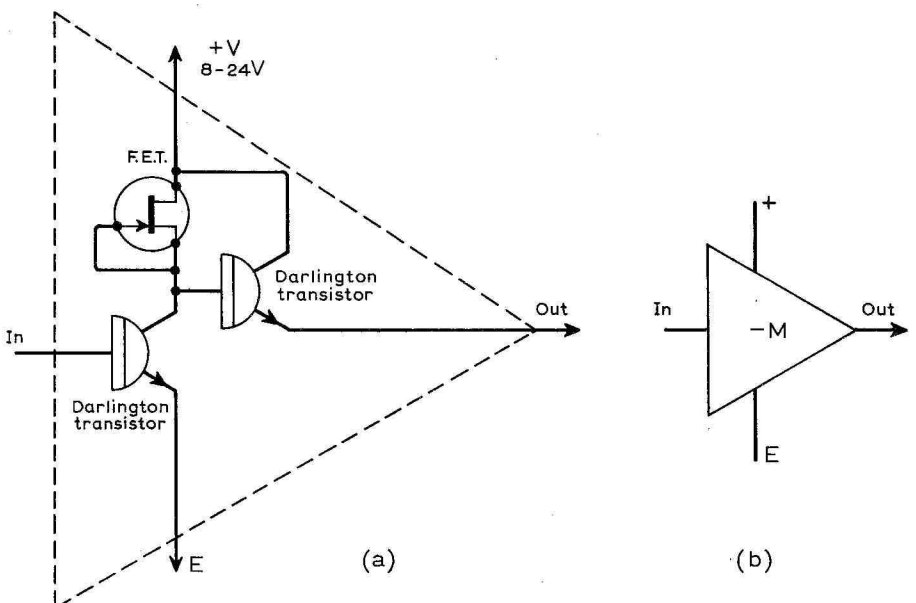
Fig. 5. (a) Basic liniac configuration; (b) symbol proposed for liniac.

and if a very high input impedance is not required, a simple bipolar transistor of similar type can also be used as the amplifier stage. This is the system which is to be preferred if the lowest possible noise level is required, and is still capable of very high stage gains if the drive impedance is fairly low. But, for most applications, a monolithic Darlington device is preferred in this position since this has a lower collector/base feedback capacitance and therefore gives a better open-loop h.f. response.

The liniac arrangement can be made with devices of complementary symmetry, with appropriate adjustments to supply polarity, and since the f.e.t. is used as a two-terminal unit either n-channel or p-channel devices can be employed provided that they have suitable $I_{DSS}$ and pinch-off voltage values. A suitable arrangement using a single very low noise p–n–p input transistor is shown in Fig. 6.

**Stage gain.** Because of the low emitter-circuit impedance of the amplifier transistor when a Darlington device is used in this position, and because of the high dynamic impedance of the collector load, the gain of the circuit is very high—typically of the order of several thousands—even when fed from a high source impedance, and is limited, at low frequencies, mainly by the output impedance ($Z_{oe}$) of the amplifier transistor, which is effectively in parallel with the collector load. At higher frequencies, the effect of the collector shunt and Miller capacitances causes the gain to fall at −6 dB/octave. Typical gain/frequency characteristics are shown in Fig. 7.

**Distortion characteristics.** For the reasons mentioned above, this configuration will be expected to possess a significantly lower order of non-linearity than the conventional bipolar transistor amplifier using a normal resistive load. In the event, the non-linearity is reduced by the same factor by which the gain of the stage is increased in comparison with the normal bipolar transistor operated at the same collector current. This is typically 10–15 times, which is a valuable feature in audio amplification circuitry. The output-voltage/total-harmonic distortion characteristics are shown in Fig. 8. Since in normal circuit applications overall negative feedback will be employed, and this will reduce the non-linearity even further, a stage with a gain of 50× can be built with less than 0·005% t.h.d. at 1 kHz at 1 V r.m.s. output.

**Noise levels.** The noise characteristics of the circuit, at gain levels in excess of some 20× (assuming some externally applied negative feedback) depend mainly upon the characteristics of the device used as the amplifier transistor, and on the relationship between the collector current and the input circuit impedance. The best available low-noise small-signal transistors give noise figures which are about twice as good as the equivalent monolithic Darlington connected devices. For this reason, when the liniac circuit is to be used under conditions where the noise level is of importance, such as in the input stage of a high-gain amplifier, it
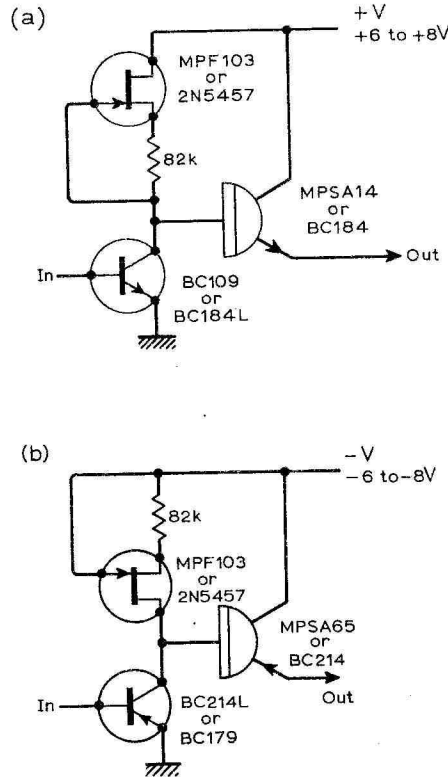


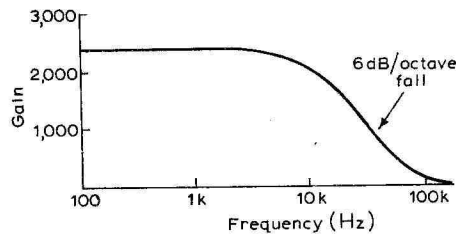Fig. 6. Very low noise modified liniac arrangements. Gain 2000–4000.



Fig. 7. Typical open-loop gain/frequency characteristics of liniac using Darlington input stage (as circuit of Fig. 9(a)).
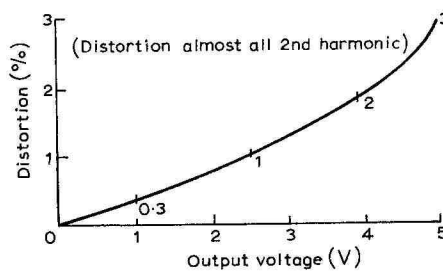


Fig. 8. Output signal voltage/distortion characteristics of liniac stage without negative feedback—Fig. 9(a). $V_c(Tr_1)$ = 8 V.

may be preferred to use the simple bipolar type, but in this case a lower input circuit impedance is essential.

In common with other transistor types the noise level at the output is reduced as the collector-emitter potential is reduced. For example, reducing the collector voltage from 8 V to 2·5 V reduces the broad band noise by about a factor of two, but also, of course, reduces the available output voltage swing. This technique should, therefore, be used with discretion.

At stage gains less than 20, the noise

contribution due to the f.e.t. may also become important, since the circuit can equally well be visualized as an f.e.t. amplifier with a bipolar constant current load, and if it is intended to use the stage with an output voltage of less than 100 mV, a low-noise f.e.t. should be used. The use of an unbypassed source resistor in the f.e.t. circuit will also reduce its noise contribution.

**Supply-line ripple rejection.** One of the more desirable qualities of small-signal amplifying stages is that they should not be affected to any large extent by ripple, voltage fluctuations or signal feedback from the h.t. supply line. This helps to eliminate hum, instability, and unexpected sources of distortion or cross-talk. Since the collector load of the transistor amplifier stage is a good constant-current source, and in typical circuit applications the input bias is not derived from the h.t. line the output signal is largely isolated from supply fluctuations. This advantage is diminished somewhat by the fact that the amplifier transistor has also a high dynamic impedance, but nevertheless the supply line rejection characteristics—assisted by externally applied negative feedback—are much better than those of the normal bipolar amplifier circuit.

**Supply and output voltages.** In typical liniac circuit applications, such as those shown in Fig. 9 et seq, closed-loop d.c. negative feedback is employed to stabilize the working voltage levels. This allows precise control of the collector potential of the first transistor stage, and thereby determines the potential drop across the f.e.t. collector load. Since it is undesirable that this should operate on the curved portion of its characteristic (cf. Fig. 4) the h.t. voltage level should be chosen so that there is at least 3 V across the f.e.t. at the peaks of the signal swing. Since the amplifier transistor should also be biased so that there is a minimum of some 2 V across it at the bottom end of the signal swing, the appropriate voltage levels may be determined simply if the output voltage swing is specified.

For example, if it is desired that the output should be 2 V r.m.s., which is 2·83 V peak, the collector voltage of the amplifier transistor should be at least 2 plus 2·83 V—say 5 V. Similarly the h.t. supply should be 3 V plus 2·83 V above this level—say 11 V. Since the forward base-to-emitter voltage drop of the Darlington transistor is some 0·9 V, the output level corresponding to the desired first transistor collector potential will be 4·1 V, assuming a Darlington device is used as the output emitter follower. If a simple transistor is employed the desired output voltage level will be 4·5 V.

The Darlington transistor used in the first stage will conduct when the base emitter potential exceeds 0·8–0·9 V $R_1$ and $R_3$ are chosen to give this—Fig. 9(a). Because of phase shift introduced by the interaction of $C_1$ and $C_2$ in this particular circuit, there will be a "hump" in the gain curve at about 10 Hz (with the capacitor values quoted) if the circuit is driven from a low-impedance source. If this is inconvenient it can be removed by a suitable input time constant

high-pass *CR* circuit.

In Fig. 9(b) the circuit has been elaborated to incorporate loop negative a.c. feedback to give a very-low-distortion amplifier with a gain of 50 and a wide bandwidth—10 Hz to 80 kHz at 3 dB—with the same d.c. levels and an input imepdance of 1 MΩ.

A simpler wide bandwidth arrangement using a lower input impedance is shown in Fig. 9(c). In this and the previous circuit a "virtual earth" feedback arrangement is employed. It should be remembered that in such cases the gain is dependent on the input circuit impedance as well, and an allowance should be made for this in the design considerations. There are obviously a large number of permutations of these basic circuits, but some specific applications are
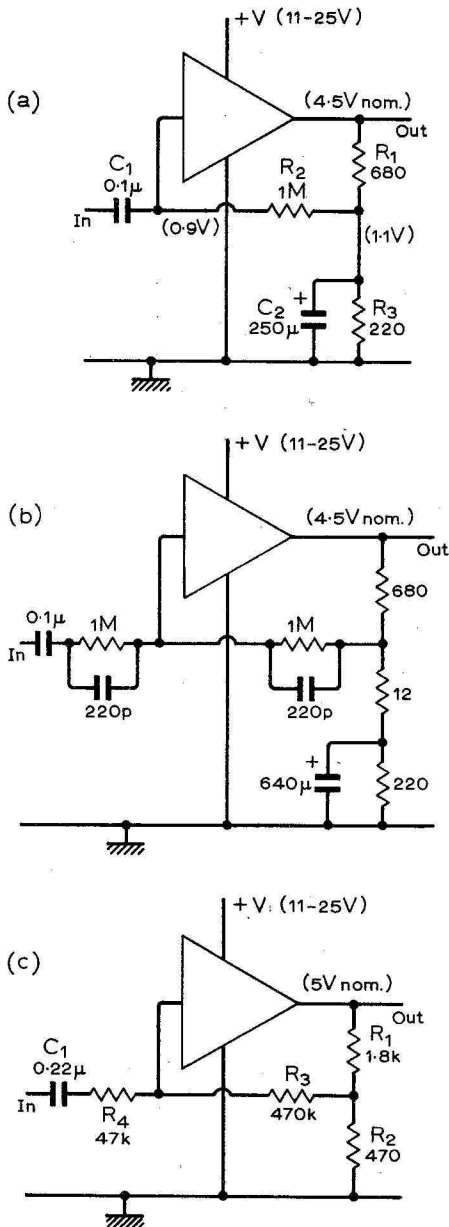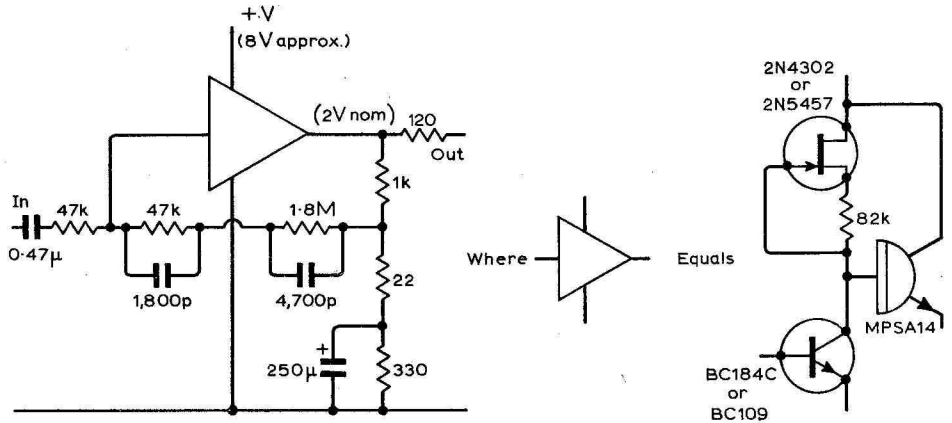
Fig. 10. Very low noise, low-distortion magnetic pickup input equalization stage. $Z_{in} = 47 k\Omega$. Gain = 50 at 1 kHz. T.H.D. <0·01% at 0·5 V r.m.s. output at 1 kHz.

shown below, in which facility for output to input loop negative feedback is exploited.

## Liniac applications

**Magnetic pickup (R.I.A.A.) equalising stage.** Because of the very high loop gain which can be obtained with this stage, even when a simple bipolar input transistor is employed, a very low noise, low distortion R.I.A.A. characteristic correction circuit can be made with this arrangement giving a gain of 50 at 1 kHz, and less than 0·01% t.h.d. at up to 0·5 V r.m.s. output. A suitable circuit arrangement is shown in Fig. 10.

**Low-distortion oscillator.** A very low distortion oscillator, employing a pentode valve amplifier, was described by A. R. Bailey in 1960[1]. In this the phase shift in a slightly unbalanced parallel "T" circuit is used to provide the necessary positive feedback to sustain oscillation, with the advantage of very good frequency stability. A circuit based on the same principle, but employing a liniac, is shown in Fig. 11. Since the number of variables is somewhat inconvenient for a continuously variable frequency oscillator, it is suggested that the capacitors should be switched to give a series of fixed frequencies.
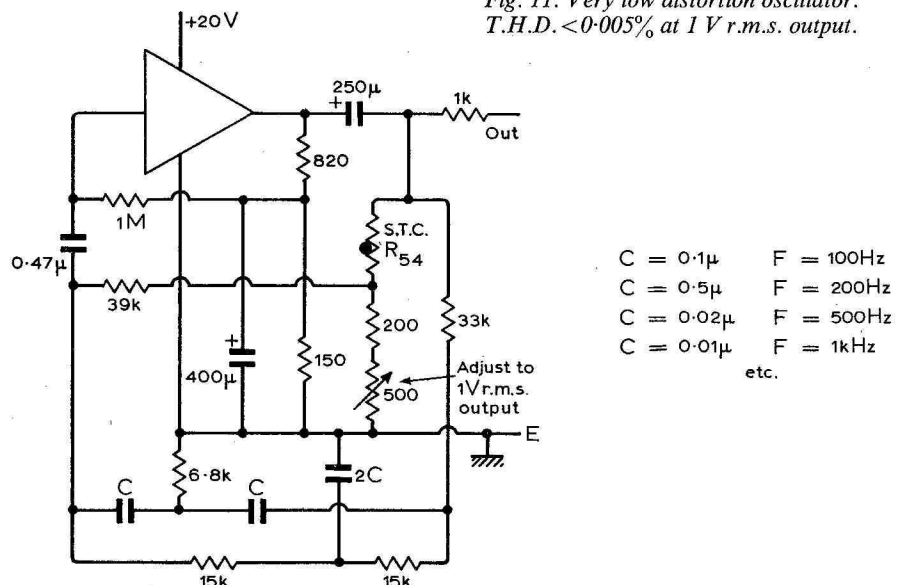
The distortion given by the prototypes of this, in the frequency range 200 Hz–5 kHz, is certainly below 0·005% at 1 V r.m.s. output. As such this circuit provides a useful reference standard for testing amplifiers, distortion meters and notch filter circuits. Incidentally the resistors used were normal high-quality carbon-film types, and no advantage was found, in terms of any measurable improvement in distortion, in changing over to wire-wound units as originally recommended by Bailey. However, the performance of the thermistor has been found to have an important influence on the overall distortion figure (of five units tried one was found to worsen the distortion to some 0·05%). It is thought that the electrolytic capacitors should also be of high quality.

**Pre-amplifier tone control circuit.** The very high gain, high input impedance and low noise and distortion characteristics of this circuit make it a natural choice for a Baxandall-type of negative feedback pre-amplifier tone control circuit, and a suitable arrangement giving approximately 20 dB of bass and treble lift and cut at 40 Hz and 15 kHz with respect to 800 Hz, is shown in Fig. 12. The worst case (maximum lift) distortion of this circuit is better than 0·02% at

Fig. 9. (a) Typical high-gain liniac amplifier stage $V_{out} = 2 V$ r.m.s. (max.). Gain ≈ 2500. Input impedance ≈ 100 kΩ. (b) High input impedance liniac arrangement. $V_{out} = 2 V$ r.m.s. (max.). Gain ≈ 50. Input impedance ≈ 1 MΩ (and 120 pF). Bandwidth ( −3 dB) 10 Hz– 80 kHz. (c) Low distortion liniac amplifier stage. Gain ≈ 50. T.H.D.< 0·01% at 2 V r.m.s. output (1 kHz).

Fig. 11. Very low distortion oscillator. T.H.D. <0·005% at 1 V r.m.s. output.

| C = 0·1μ | F = 100Hz |
|---|---|
| C = 0·5μ | F = 200Hz |
| C = 0·02μ | F = 500Hz |
| C = 0·01μ | F = 1kHz |
| | etc. |

1 V r.m.s. output. This is at least 20 times better than the conventional (and very widely used) single transistor circuit under similar worst case conditions.

### Other circuits using Darlington transistors

**F.E.T.—bipolar feedback pair.** Because of the relatively high output impedance of the normal grounded-source junction f.e.t. amplifier, it is not possible to construct f.e.t.
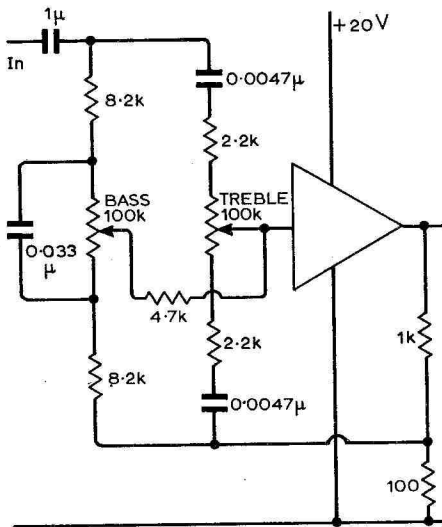


*Fig. 12. Liniac employed in tone control stage. Max. output 3 V r.m.s. Source impedance $\leqslant 10\,k\Omega$. Midpoint gain $10\times$ $\pm 18$ dB lift/cut at 50 Hz and 15 kHz w.r.t. 800 Hz. Worst case t.h.d. $<0.02\%$.*

—bipolar feedback pairs of a form analogous to the excellent circuit arrangements typified by Figs. 1(b) and 1(c), without the overall gain being much reduced by the inevitable mismatch at the drain of the f.e.t. However, if the second transistor is a Darlington device, the mismatch is avoided, and open loop gains of $4000\times$ are feasible, in the non-inverting mode. The circuit arrangements is shown in Fig. 13. For comparison, the same circuit with a 2N4058 or BC214 as $Tr_2$ has only a gain of 100.

**Improved bipolar feedback pair.** The circuit of Fig. 2(b) can itself be improved by the
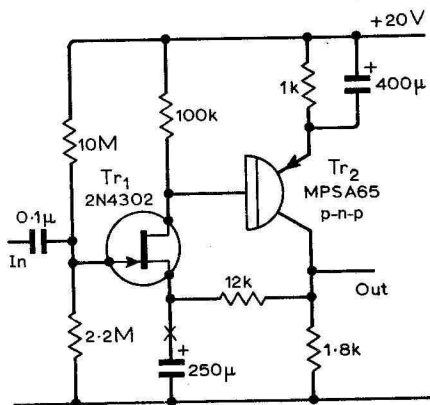


Feedback resistor inserted at X to provide feedback control of gain

Open loop gain $\simeq 4,000$

*Fig. 13. F.E.T./Darlington pair. High-gain high-input impedance.*

use of a Darlington transistor as $Tr_2$. The use of an MPSA65 p-n-p device gives loop gains in excess of 6000, for example. A suitable circuit of this general type is shown in Fig. 14.

**D.C. bootstrap circuit.** The fact that the emitter of a Darlington transistor will follow the base signal level very accurately, with a constant potential difference of about 1 V, allows the connection of a load resistor between the base and emitter as shown in Fig. 15, which multiplies the effective dynamic impedance of the resistor at all frequencies down to d.c. by a figure which approaches the Darlington transistor current gain. The f.e.t. amplifier circuit has a gain of about 250.
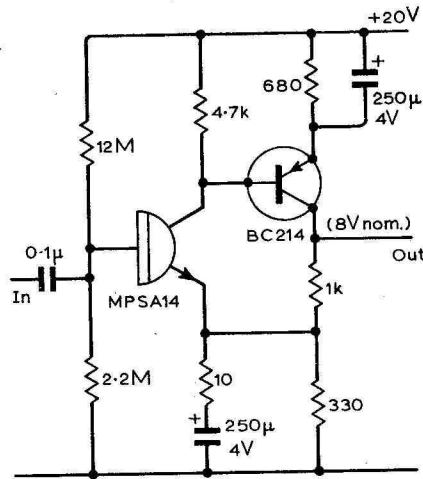


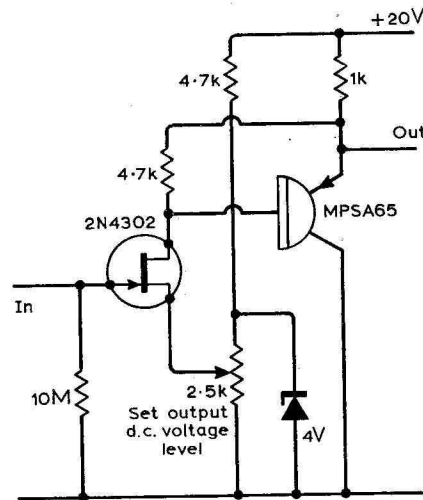*Fig. 14. Improved bipolar transistor feedback pair. $Z_{in} \approx 1.5\,M\Omega$. Gain $\approx 100$.*



*Fig. 15. D.C. bootstrap circuit (phase inverting). Gain $\approx 250$.*

Inexpensive plastic encapsulated and other relatively low-cost devices of this type are available from Motorola, Fairchild, SGS, and GE. Type numbers are MPSA 12, 13 and 14, BFX 66 and 67, and D16P4 for n-p-n types; and MPSA 65 and 66 (Motorola) for p-n-p devices. The MPSA 12 Motorola unit is a low noise pre-amp type.

**Reference**
1. Bailey, A. R., *Electronic Technology*, Feb. 1960, pp. 64–67.

# H.F. Predictions— September

Solar activity is now steadily declining as this table of Ionospheric Index IF2 shows.

| | 1966 | 1969 | 1971 |
|---|---|---|---|
| Jan. | 15 | 95 | 94 |
| Feb. | 21 | 104 | 85 |
| Mar. | 33 | 127 | 83 |
| Apr. | 37 | 122 | 74 |
| May | 46 | 118 | 70 |
| June | 55 | 119 | (70) |
| July | 55 | 114 | (68) |
| Aug. | 53 | 122 | (65) |
| Sept. | 42 | 115 | (63) |
| Oct. | 47 | 110 | (59) |
| Nov. | 64 | 106 | (57) |
| Dec. | 66 | 108 | (55) |

Forecast values are given in brackets. The years 1969 and 1970 were almost identical and constitute the maximum of the current sunspot cycle. A minimum is expected in 1975.
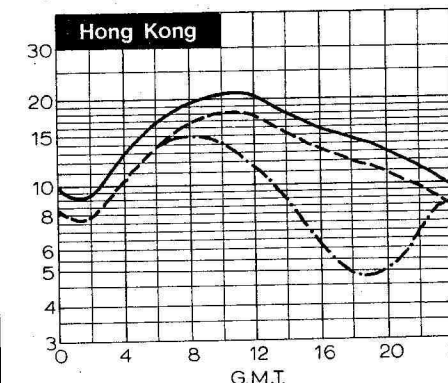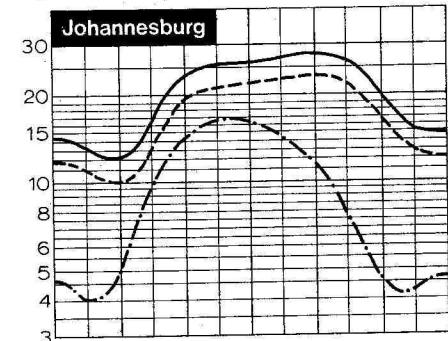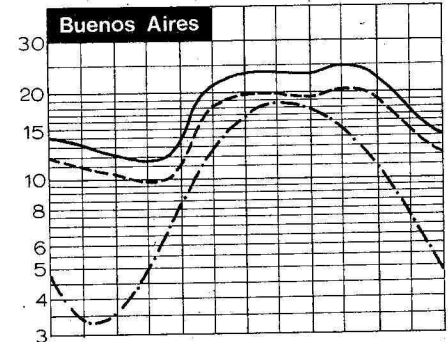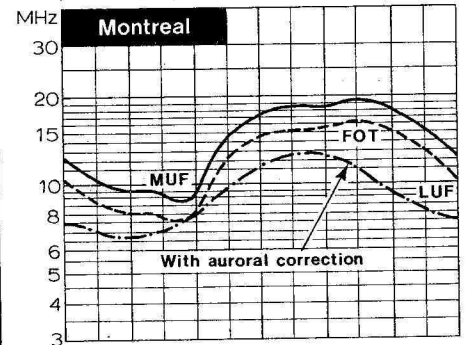
# Letter from America

The Consumer Electronics Show (C.E.S.) was held at the end of June in the exhibition centre at Chicago's McCormick Place. A few of the 300 exhibitors had extra-mural demonstrations in local hotels and although there was a well-organized charter bus service the humid, steamy heat with temperatures around 100°F made travel somewhat uncomfortable. Inside the show the scene was similar to the old Radiolympia in London with rows of elaborate stands, TV displays and loudspeakers making a continuous babel of sound. And, of course, each stand had its group of aggressive salesmen in newly pressed suits with here and there a gaily dressed (?) girl giving out leaflets and carrier bags. But there was a difference—the C.E.S. is for trade only and so the atmosphere was, in some respects, more serious. Also, the large hall was well air-conditioned and, note this, free champagne was given to the visitors on a terrace overlooking Lake Michigan. It certainly beats coffee on Hammersmith Road! Attendance for the four-day show was 36,200; more than 20% higher than last year's figure. Some very interesting TV sets were to be seen—including one from JVC shaped like a ball—but emphasis was definitely on audio, and four-channel sound in particular. Almost every stand boasted some kind of demonstration room.

The majority of exhibitors were using matrix, or synthetic, four-channel systems which are proliferating at an alarming rate causing a great deal of confusion. One dealer summed it up by saying "The situation has now got out-of-hand and we don't know which system is best and what will work with which encoder". On the other hand, another dealer was more optimistic and in his view "Most systems are compatible enough for a record or tape made by system A to produce an acceptable four-channel surround sound when played back via a decoder intended for system B". The long-awaited C.B.S. SQ disc system* was being demonstrated at a nearby hotel and comparisons were made with 15 i.p.s. master tapes. One of

the records had a commentary by David Frost (very popular this side of the Atlantic) and in spite of the high volume levels, it was one of the most convincing demonstrations I have yet heard. C.B.S. have already announced that they will release at least 50 SQ discs by the end of the year and that agreements have been made with Sony for the production of decoders and playing equipment.

Ampex were using both discrete and matrix systems but most tape recorder firms were content to use discrete four-channel tapes and at least three had cassette machines. The 4/8 track format was also popular.

Pioneer released details of a new miniature (Hipac) stereo cartridge. This is one-quarter the size of a standard eight-track type and smaller than a normal cassette. Koss were showing four-channel headphones. Triumph had headphones with a built-in five-transistor radio.

A number of f.m. stations are using Electro-Voice encoders and, Allied, a large chain concern with several hundred shops throughout the country, are busily demonstrating the EV system. Their competitors Lafayette, are equally committed to the Dynaco system which has the merit of requiring little extra equipment.

An extra-mural demonstration was given by Ray Dolby in conjunction with f.m. station WFMT. This station played a selection of tapes some of which used the Dolby mode. Several Dolby 'black box' equalizers have been lent to listeners in
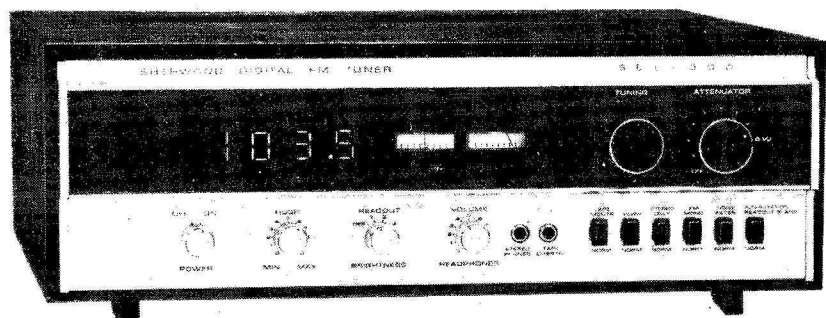
various locations and comments invited. As might be expected, those equipped with Dolby units—especially in fringe areas—reported spectacular improvements in signal/noise. Most of the listeners without Dolby units (who were advised to turn down their treble controls to produce a more balanced sound) preferred the extra brightness. So far then, tests show that the Dolby system as used for broadcasting is compatible enough to avoid conflict with the F.C. C.

Back at the show for a quick look at TV. Last year many observers predicted a big swing to i.cs but this has not materialized. The main reason is the higher cost involved. No doubt, higher production will bring down prices—but this is like the old chicken-and-the-egg story. Meanwhile, RCA have dispensed with the valve e.h.t. rectifier in their colour sets, thus making them all solid-state. One model uses no fewer than 12 modules that plug into two p.c. 'mother' boards—fine for the service man. The great majority of exhibitors of TV receivers were Japanese, which underlines the extent of Far-East competition. GE say they will discontinue production of radio receivers next year—leaving no large U.S. maker of domestic receivers.

The 60th anniversary edition of the *Wireless World* brought back some memories and I was especially interested in John Gilbert's letter mentioning Ted Rosen of Ultra. I was a tester for that firm at their Harrow Road factory around 1930 and I well remember a radio receiver called "The Switchboard to Europe". I also have fond memories of Brownie Wireless, makers of crystal sets and the Wates Company where an Everyman Four was used for testing phono-pickups These monsters tracked at four ounces and were fitted with an attachment for 'swans neck' gramophone tone arms!

But my clearest memory is listening to KDKA with a home constructed 1-valve set which had a coil wound on a wine glass (low-loss!). These days we have colour TV, videotape, quadraphonic sound, satellite communications and so on. All these are exciting enough but, for me, nothing can compare with the thrill of listening to KDKA from that attic in Camden Town more than 40 years ago.

G. W. TILLETT

* The compatible C.B.S. 'SQ' (stereo/quadraphonic) system uses a method of circular modulation of the two sides of the disc groove for the left and right back signals, as well as normal modulation for the front signals.—Ed.



*Sherwood digital read-out f.m. tuner.*

# Field-sequential Colour Television Receiver

## 1—Introduction and basic principles

*by T. J. Dennis, B.A.*

All systems of colour TV in general use today have as their display a system whereby the three primary coloured pictures are spatially superimposed, whether by projection of the red, green and blue images using the Schmitt system, by the use of three c.r.ts and half-silvered mirrors, or with the three pictures on one c.r.t. whose screen consists of triads of independently controllable phosphor dots, as in the R.C.A. Shadowmask[1] tube.

All three systems are capable of excellent results, but are difficult and expensive to set up. For example, in the projection system complex distortions have to be introduced into the scanning waveforms to correct for the fact that the projectors cannot be co-sited. Much the same problem is encountered with Shadowmask tubes, hence the joys of convergence adjustments. Any system using separate electron sources is prone to grey-scale tracking errors.

The Shadowmask is able to reproduce a range of colours because the spatial colour resolution of the eye is poor: close to a screen the dots can be easily perceived, but the overall impression is still one of the additive colour resultant. Temporal colour resolution is equally weak, as may be seen by rotating a disc carrying segments of, say red and blue, when the colours rapidly merge to magenta as the speed of rotation is increased. This is the basis of the field-sequential process, whereby the three coloured images are presented to the eye in turn. It is the oldest form of colour display, a version having been demonstrated by J. L. Baird in 1928.[2] In the author's opinion it is capable, within its limitations, of giving results of the highest quality.

Perhaps the major of these limitations is caused by the eye itself: perception of luminance, or brightness, changes in time, as well as in space, are particularly good. While a rotating disc of red and blue will appear magenta, it also carries a marked brightness flicker due to the luminance difference between red and blue. Flicker only disappears when its frequency is

higher than the flicker-fusion rate of the eye, a highly variable quantity found on average in the region of 30Hz.

For this reason it is normally considered necessary to increase the basic field rate from that of, say, a 50 field monochrome standard to 150 fields per second in order to maintain the original luminance flicker rate. This demands a trebling of the signal bandwidth, other factors being constant.

Noting the discouraging comments of others on the subject of f.s. systems retaining the existing monochrome field rates[3], it was decided to attempt to build and operate such a unit, to work from the

normal broadcast colour transmissions.

A standard PAL decoder[4] provides the three (narrow band) colour difference signals. These are then switched in turn to the grid of a monochrome c.r.t., the change taking place during the field blanking period (see Fig. 1). The luminance (wideband) signal is fed to the c.r.t. cathode as usual, after its passage through a 600ns delay line. This is practically the only major modification needed to the monochrome receiver which is the source of all the signals used. The net result is that the set can be made to display, field sequentially, the black-and-white equivalents of

[1]Herold, E. W., 'Methods suitable for TR color kinescopes,' *R.C.A. Review*, Vol. 12, Sept. 1951, p 445 et seq.

[2]'TV in natural colours demonstrated', *Radio News*, Vol. 10, p. 320, October 1928

[3]Goldmark, P. C., et al., 'Color Television', Part I, *Proc. I.R.E.*, Vol. 30, pp. 162-182, April 1942.

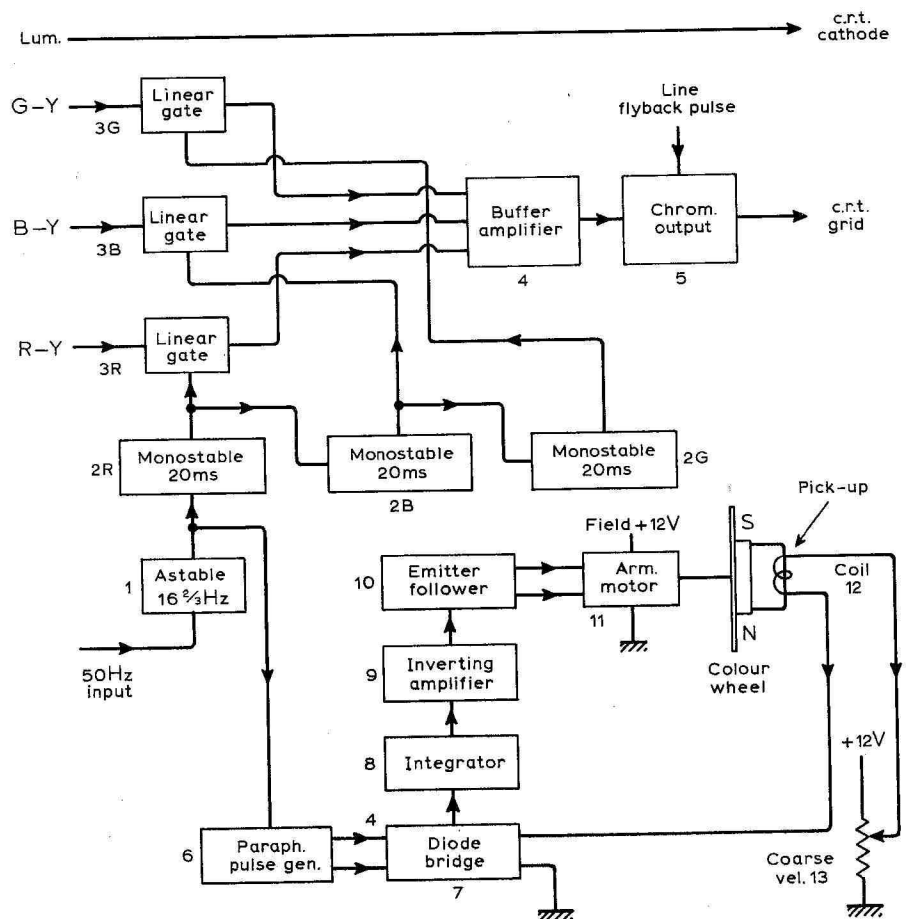[4]See T. D. Towers on principles of colour TV, Jan.-Dec., 1967, *Wireless World*.



*Fig. 1. Block diagram of field-sequential colour receiver equipment. The input is provided by a normal monochrome receiver.*

the red, green and blue images of a colour transmission.

A disc carrying sections of primary red, green and blue filters rotates at 16⅔ r.p.s. in front of the c.r.t. Its rotation is phase locked to the field sync pulses to ensure that when the red picture is being scanned, the red filter is in place, and so on.

It may be noted that two-thirds of the available colour information is wasted in this system, but it should also be recalled that considerably more than two-thirds of the energy imparted to the electron beams in the Shadowmask tube is dissipated as heat in the Shadowmask!

Phase lock of the colour wheel is required, to ensure that the correct filter is in place at the correct time. This is achieved by a simple feedback system using a signal derived from a coil wound on a U-shaped transformer limb, and mounted in front of a bar magnet fixed to the centre of the wheel. The coil output waveform is square in form, with slow sinusoidally changing edges. A four-diode bridge is used to gate through an 8ms portion of this waveform, which has a manually controllable d.c. potential superimposed on it for coarse speed adjustment. The bridge output is integrated, amplified, and with suitable d.c. level adjustments, used to drive the motor armature via a 2N3055 emitter follower.

Assuming phase lock, the gating pulse is placed symmetrically about the mid-point of the positive going edge of the feedback waveform. If the motor speeds up for any reason, the waveform reaches a higher level than it would normally when sampled, and the integrator output moves in a positive direction. Because of the inverting amplifier, the armature voltage is reduced, and the motor slows down. By similar reasoning it can be shown that a reduction in motor speed will also be compensated. Not surprisingly, the system oscillates about its stable position when any velocity transient is applied; settling time from switch-on is about 20 seconds in the prototype, but this is immaterial as it takes the line timebase considerably longer to warm up on the displaying set. Programme switchings, when field sync may be interrupted, tend to upset phase lock, but this effect has not been found troublesome.

## Results

Before embarking on the construction of a PAL decoder, a generator was built to produce the 4f, 2f and f, where f=line frequency, squarewaves needed for the blue, red and green, respectively, signals of the standard colour bars, viz. white, yellow, cyan, green, magenta, red, blue and black. These were applied to the linear gates. The resulting non-composite output was then passed through an existing camera channel, and emerged with a full set of 405 line-standard sync pulses, for ease of application to a monitor.

The resulting wildly flashing vertical stripes, when viewed through a locked colour wheel, became the familiar bars. Colour fidelity, even with the rather crude ex-stage lighting filters in use, was in general excellent, the yellow being the
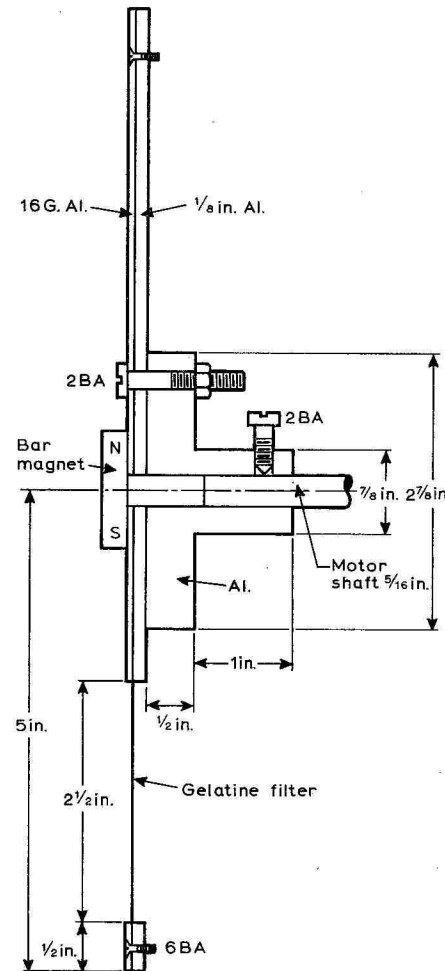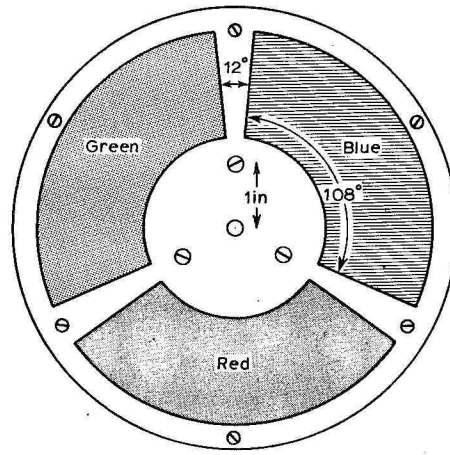


Fig. 2. Front view (a) and section (b) of a simple colour wheel.

least well presented as on all colour sets. The reddish tint obtained when observing a white object through the wheel (due to the red filter having excessive transmission) was neatly compensated by the blueish tint of the c.r.t. phosphor.

Passers-by who ventured unsuspecting into the lab. during this stage of development were invited to peer through the disc, and report the colours seen. Most were correct without prompting, but two insisted they saw blue and red separately on the magenta bar. This only tended to happen at high brightness levels, and is an effect not observed by the author.

Owing to interlacing, successive lines

on the *screen* (not per field), when displaying any but saturated primary colours, will differ in shade. However, since the colour detail resolving power of the eye is poor, the effect could only be seen within about 12in. of the 14in. c.r.t. used. Bearing in mind that these initial tests were on 405 lines, with a 625-line colour picture at normal viewing distance, the effect is unnoticeable.

After the encouraging results obtained with the colour bars, a PAL decoder was built, with slight modifications, notably in elimination of dependence on the line output stage of the receiver: an additional sync separator was added, the line pulses obtained being used to trigger a monostable and produce an accurately timed burst gating pulse. The burst gate itself was in the form of a four-diode bridge, all of which will be discussed more fully next month.

At first the decoder was operated without a delay line; i.e., in the PAL-S mode. Oscilloscope examination of $R-Y$ for the colour bars with careful adjustment of $L_6$ of the May 1969 *W.W.* article enabled results to be obtained which did not differ appreciably line by line. Stability over long periods, however, was not good due to mechanical vibration and thermal changes. Hanover bars were then obtained. Addition of a PAL delay line effected a complete cure.

Adjustment of the $R-Y$, $B-Y$ and $G-Y$ drives to the sequential switches enabled colour pictures to be obtained whose fidelity was indistinguishable from Shadowmask results, with the advantages of full luminance bandwidth (a notch filter has been found unnecessary; some commercial receivers do not include them), and total elimination of the necessity for complex convergence and grey-scale tracking adjustments. With the latter, even if the filters do not give an exact white, there can be no failure, since the same gun is used for all three pictures. Problems will arise, however, if any attempt is made to provide switched compensation for filters of wildy incorrect characteristic.

As mentioned above, field-sequential systems working at low field rates suffer from luminance flicker effects. Another problem is colour fringing, obtained when there are differences between adjacent fields; i.e., when the scene contains movement.

Fortunately, both have proved a far less serious drawback than was expected.

Perception of flicker depends on many factors including background light level, degree of dark adaption and size of the field under consideration. Thus, viewing a f.s. picture under well lit conditions results in the flicker being highly objectionable: the colours are desaturated, and may not be seen at all. This seems to be true whatever the brilliance of the displayed image, which has in any case to be high to overcome the effects of reflected light from c.r.t. screen and colour wheel.

The improvement when pictures are viewed in either total darkness, or very low ambient lighting is considerable, particularly once dark adaption has taken

place. Flicker due to the luminance difference between the red, green and blue images in a black-and-white transmission is negligible, while there is no sensation of colour at all.

In general, flicker in coloured pictures increases with increasing area of colour, its saturation and luminance level, and is greater for the primary colours, particularly green, than the complementaries. The latter is true, since the mark-space ratio with the saturated primary colours is 1:2 (i.e., one field out of three is displayed), while for saturated cyan, magenta and yellow, this ratio is 2:1.

Most programme material does not, however, carry large areas of saturated colour, and the viewer may be unaware of flicker, depending on the content of the programme and its degree of 'viewer involvement'.

A warning is due here: it is probably unwise for anyone susceptible to flicker, as in some cases of epilepsy, to view colour television in this way, as it contains, as well as major components at $16\frac{2}{3}$Hz and 50Hz, smaller components at 25 and $8\frac{1}{3}$Hz due to interlacing. The latter particularly is close to the so-called danger frequency of 7Hz. However, the author, who does not suffer from epilepsy, has used himself as guinea-pig in viewing trials as long as three hours, with no ill effects, apart from a crick in the neck from the difficult viewing position necessary with the prototype: the colour wheel is 10in. in diameter, and close to the eye, while the raster is on a 17in. c.r.t. 4ft away.

The second problem of colour splitting is, of course, only apparent on images carrying movement; it has, however, been found that any movement has to be quite fast before splitting becomes visible, the gesticulations of an orchestral conductor being particularly susceptible. In most cases, though, the subject of attention in a scene is kept stationary on the screen, while the background moves. An example is a horse-race, where the rails can be seen by a conscious effort as red, green and blue bars.

### Possible forms of colour wheel

The prototype colour wheel was a simple affair, and is shown in Fig. 2.

Two aluminium discs were cut out using a woodworker's routing machine. One was $\frac{1}{8}$in. thick, the other 16 gauge. The three cutouts were then made with the same tool, which proved remarkably efficient, a bolt being placed through the centre of the disc and router plate so that the cutting edge of the router was at the required radial distance. The disc was then rotated slowly, leaving the radial arms of the wheel. Straight sections were cut with a hacksaw. Pieces of red, green and blue gelatine filters (as used for stage lighting) were sandwiched between the discs, which were clamped together by the screws through the machined mounting plate, and 6BA screws into holes tapped in the $\frac{1}{8}$in. disc periphery.

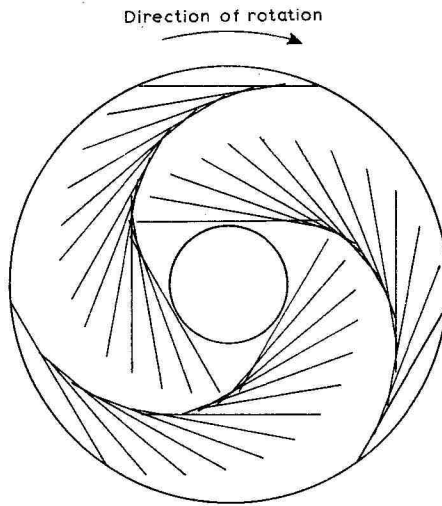The angular position of the magnet in



Direction of rotation

Fig. 3. Spiral colour wheel. The spiral cut-outs are represented by the position of scanning lines seen through the disc at 1/12th field intervals. This figure is drawn for a raster of dimensions 6 × 4.5 units, the central area having a diameter of 4 units.

relation to the pickup coil is adjusted so that the required section of filter moves down the c.r.t. with the field scan carrying that colour, giving a maximum segment of the wheel through which the correct colours can be seen. This implies that the colour picture can only be viewed with one eye through the side of the disc; however, if the viewer moves back about two feet, the right-hand side of the picture can be seen with the left eye and vice versa, with only small (top right, bottom left) areas cut off.

An alternative form of disc uses spiral areas of colour, which follow the field scan down the screen, and enables the colour picture to be seen through the top of the wheel. The spiral wheel can be made slightly smaller than its simple counterpart, thus a specimen for operation directly in front of a 10 or 11 inch c.r.t. is feasible.
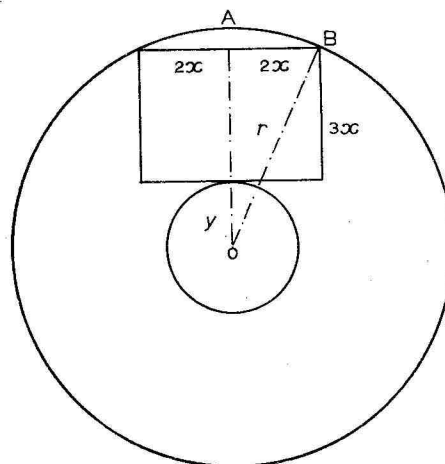


Fig. 4. Disc radius for given raster size. Spiral colour wheel mounted in front of the c.r. tube. Rotation is anticlockwise. The automatic phase-control components are mounted at the rear of the equipment on the colour wheel shaft.

Fig. 3 shows the basic form of the spiral holed wheel, while the calculations for the diameter of it for given raster size are illustrated in Fig. 4.

Let the dimensions of the raster be $4x$ by $3x$, assuming a standard 4:3 aspect ratio. Then in triangle OAB,

$$OB^2 = r^2$$
$$= AB^2 + AO^2$$
$$= 4x^2 + (3x + y)^2$$
$$= 13x^2 + 6xy + y^2, \text{ whence, by taking}$$

the square root, $r$ can be determined.

The dimension $y$ is determined by the physical size of the driving motor, and other factors such as mounting arrangements, but a useful rule for a minimum value is to make $y =$ one-third picture width.

The radius required for this type of disc is clearly less than that for the simple wheel first described:

For a simple wheel, radius$^2 = 18.25x^2 + 8xy + y^2$, by similar reasoning to that above, compared with $r^2 = 13x^2 + 6xy + y^2$ for the spiral disc.
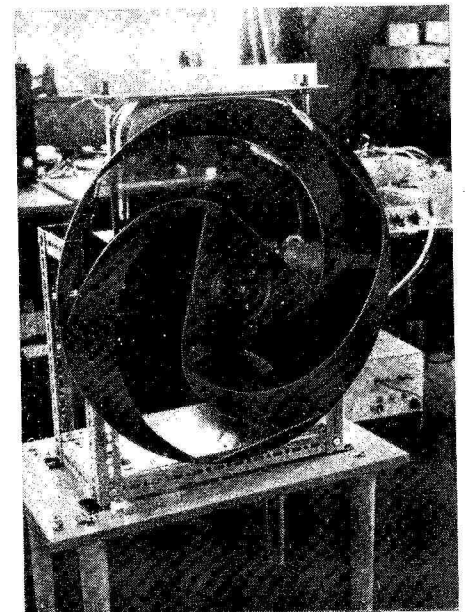


Spirally cut colour drum in front of the cathode-ray tube.

Perspex is an ideal material from which to fabricate a spiral wheel, as the diameter can then be made the minimum possible. It is very easy to work, the routing machine again being ideal for cutting out the two $\frac{1}{8}$th inch thick discs required. The discs are fitted to an identical mounting boss to that used with the simple wheel, but care should be taken with clamping bolts, as the plastic tends to shatter under pressure. Aluminium discs of radius dimension $y$ should be placed on each side of the Perspex at the centre to spread this load. Periphery clamping screws should be countersunk 6BA types, and no longer than necessary, to minimize windage. Again, they should not be overtightened.

In order to obtain a 9.6in × 7.2in (12in diagonal) colour picture, a 23 inch diameter specimen of this type has been manufactured, with successful results. Careful balancing of a wheel of this size is

necessary, and this was carried out by placing it, with a 6in length of shaft through the centre, between two horizontal edges, and adding pieces of lead to the screws through the protective aluminium discs, until the wheel would remain stationary in any angular position.

The following instructions, in conjunction with Fig. 3, can be used to construct the spiral holes. The figure can be conveniently drawn on the protective paper covering the Perspex sheets at purchase.

1. Calculate the desired radius of the disc from the selected values of $x$ and $y$. ($x$ can be determined from the relation $5x=$ diagonal of raster used, since the diagonal and two of the sides of a 4:3 raster make the '3-4-5' triangle of elementary geometry.)

2. Divide the height of the picture into—say—twelve sections of length $a$ (i.e., $a=3x/12$).

3. Divide a 120° segment of the disc into the same number of segments, here twelve of ten degrees each, drawing radial lines.
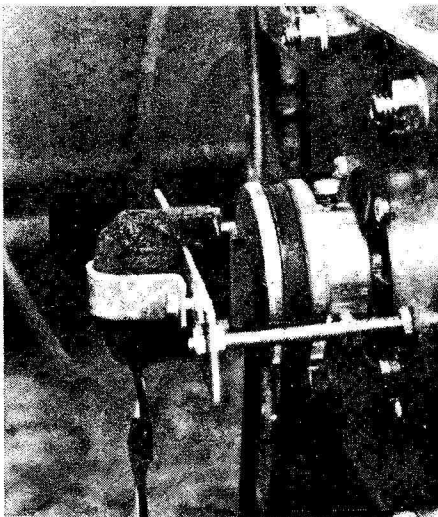
*Synchronizing magnet and pick-up coil.*

4. Draw a line of length $4x$ perpendicular to, and bisected by, the radius pointing towards the top of the paper, at a distance $12a+y$, (i.e., height of picture + $y$), from the centre of the circle.

5. Repeat step (4) with each radial line, moving in an anticlockwise direction, and reducing the distance of the perpendicular from the centre by length $a$ each time, until the 13th radius is reached, when a line distance $y$ from the centre should be drawn.

6. Repeat steps (3) to (5) for the remaining 120° segments of the circle, starting where the innermost perpendicular of the previous spiral was drawn.

This process builds up an envelope of the spiral holes needed, which can be completed freehand. Using a greater number of increments will, of course, increase the accuracy, but tend to clutter the diagram somewhat.

Slices of the coloured filters should be cut to shape, and sandwiched between the discs after removal of all paper but that carrying the design. Small pieces of adhesive tape can be used to secure the filters in position during final assembly, after which all areas of the wheel needed to be opaque should be coated with blackboard paint.

*(To be concluded)*

# Voltage Reference Source

## Constant-current drive with 0.08% stability

*by H. A. Cole\*, M.I.E.R.E.*

Specially constructed zener diodes having very low temperature coefficients (less than ± 0.002%/deg C) are now readily available at moderate cost, and are intended for applications in which a highly stable voltage reference is required. However, unless the operating current of such diodes is maintained within closely defined limits, the advantage of a low temperature coefficient will be lost due to voltage variations occurring across the internal impedance of the diode.

There are many ways in which a constant operating current can be provided for a reference diode but one of the simplest and most effective method is by use of the 'ring-of-two' circuit introduced by Williams in 1966 (references 1). Unfortunately, although this circuit performs extremely well over a wide range of supply voltage variations, its inherent temperature dependence (about −4mV/deg C) makes it unsuitable for use in circuits subjected to wide variations in temperature. The principal cause of its high temperature dependence is variations in $V_{be}$ of the two transistors (typically −2mV/deg C each).

At first sight it might appear feasible to compensate for $dV_{be}/dT$ by selecting zener diodes used in the ring with temperature coefficients identical to those of the transistor $V_{be}$. Unfortunately, although such an arrangement is not impossible, the difficulty of obtaining a zener diode having the desired voltage *and* temperature coefficient (of the desired sign) is considerable. A better solution is to use zener diodes which have a negligible temperature coefficient and then connect an ordinary forward-biased diode in series with each. An arrangement like this lends itself readily to the use of a dual transistor with matching $V_{be}$, as the base-emitter junction of one transistor can be used as the compensation diode for variations in $V_{be}$ of the other transistor. Unfortunately, because of unequal currents in the two junctions, complete compensation cannot be expected. A circuit based on this arrangement, but using transistors with unmatched base-emitter voltages, is now described.

A high-stability reference diode–$D_3$–is supplied with a constant current of
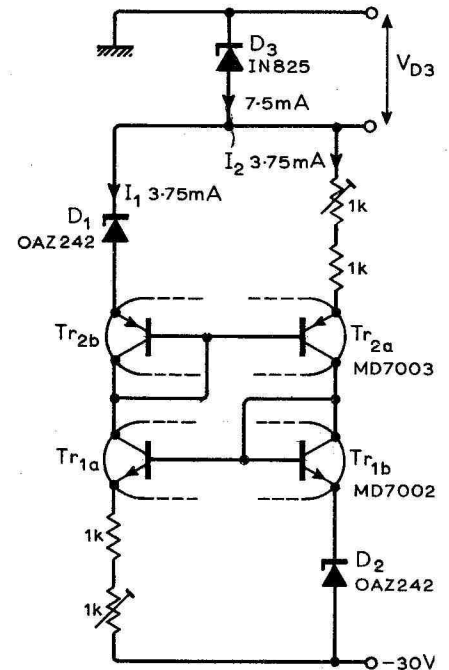
\*A.E.R.E., Harwell

*Fig. 1. To compensate for temperature sensitivity in the transistor $V_{be}$ of the ring-of-two voltage reference, dual transistors with matching $V_{be}$ are used. One base-emitter junction is used as the compensation diode for $V_{be}$ variations of the other transistor. From Fig. 2, circuit is required to give $7.5 \pm 0.2mA$ through $D_3$.*

7.5mA by the remainder of the circuit connected to operate as a ring-of-two, each half providing 3.75mA (Fig. 1). The reference voltage for each half of the ring is formed from the series connection of a zener diode having a very low temperature coefficient (typically +0.5mV/deg C at 3.75mA), and the base-emitter junction of a transistor operated as a forward-biased diode. The overall temperature coefficient of each series connection is about −1.5mV/deg C and provides reasonable temperature compensation for variations in $V_{be}$ of the transistor which it supplies. In Fig. 1, therefore, the zener voltage of $D_1$, plus the $V_{be}$ drop of $Tr_{2b}$ forms a temperature-compensated reference voltage for the transistor $Tr_{2a}$. In a similar way $Tr_{1a}$

is provided with a temperature-compensated reference voltage consisting of the drop across $D_2$ and the $V_{be}$ drop of $Tr_{1b}$.

The current flowing in each half of the ring is adjusted to the desired operating current by the variable resistors. These are wire-wound trimming potentiometers having temperature coefficients of 80 p.p.m. The fixed resistors are metal-film types of 1% tolerance, having temperature coefficients of 50 p.p.m.

At the recommended operating current of 7.5mA, the temperature coefficient of $D_3$ (type 1N825) is at its minimum value, as shown by the curves[2] of Fig. 2. When the operating current ($I_z$) of $D_3$ is increased to 10mA (2.5mA above the recommended value), the temperature coefficient (measured over the range $-55$ to $100°C$) is about $+0.002\%/\text{deg C}$. A similar coefficient, but of opposite sign, is obtained for the same temperature range when $I_z$ is reduced to 5mA (2.5mA below the recommended value). It may be concluded, therefore, that the temperature coefficient of $D_3$ may be considered independent of operating current, and within the manufacturer's specification of $\pm 0.002\%/\text{deg C}$, provided $I_z$ is held constant at any value within $\pm 33\frac{1}{3}\%$ of 7.5mA.

The same cannot be said for the dependence of the zener voltage $V_{D3}$ on the operating current as, from the 25°C curve of Fig. 2, an increase of 2.5mA above the recommended 7.5mA causes a change in $D_3$ of 34mV, an increase of 0.55%. This is due to the dynamic impedance of $D_3$ which is about 12 ohms at 25°C and 7.5mA. A 2.5mA reduction in the recommended 7.5mA (at 25°C) causes a change in $D_3$ of 37mV, a reduction of about 0.6%. Thus

$$\frac{dV_{D3}}{dI_z} \simeq \frac{0.6\%}{2.5\text{mA}} = 0.24\%/\text{mA} \quad (1)$$

referred to $I_z = 7.5\text{mA}$.

Comparing the two coefficients, a 26% increase in $I_z$ from 7.5 to 9.45mA causes $V_{D3}$ to change by 0.47%. On the other hand, a 26% increase in temperature from 25 to 100°C, causes $V_{D3}$ to change by only 0.15%. The $I_z$ coefficient is therefore more than three times greater than the temperature coefficient and in most instances will determine the overall stability of the circuit shown in Fig. 1.

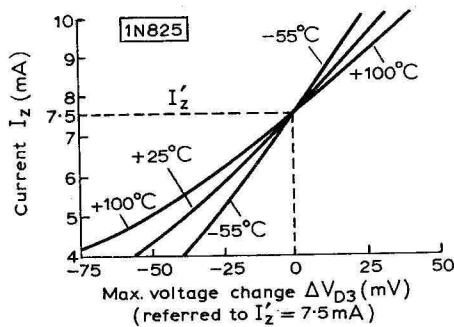To maintain an overall stability of $V_{D3}$ versus $I_z$ no worse than that of $V_{D3}$ versus

Fig. 2. Because the $V_{D3}$ stability with respect to $I_z$ is worse than with temperature–graph shows three times worse–$I_z$ is restricted to $7.5 \pm 0.2\text{mA}$, for a stability of 0.05%.

temperature, the total variation in $I_z$ should produce a change in $V_{D3}$ which is small compared with that produced by a total excursion of temperature within the accepted range. Assuming, for example, a working temperature range of 0 to 50°C, the expected overall stability of $V_{D3}$ versus temperature is $50 \times 0.002 = 0.1\%$.

If the maximum allowable variation of $V_{D3}$ versus $I_z$ is made 0.5%—half that allowed for the total temperature variation—then from expression (1) $I_z$ must be maintained within $\pm 0.2\text{mA}$ of the recommended value, i.e. $I_z$ must lie within the limits 7.3 to 7.7mA. This, therefore, is the current
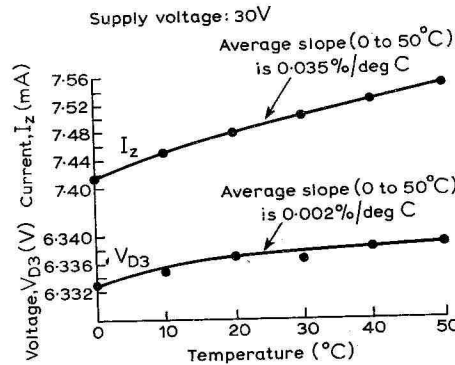
Fig. 3. From these measured temperature stability curves, overall stability is 0.065% referred to the 30°C value.
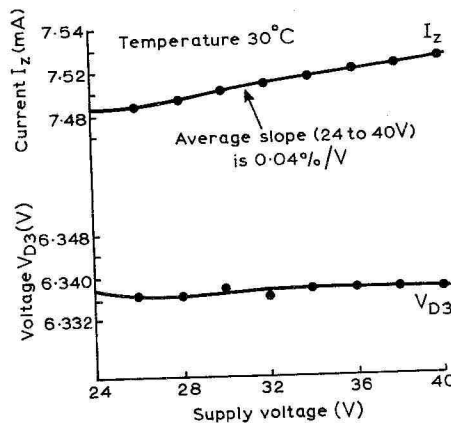
Fig. 4. From these measured voltage stability curves overall stability is 0.016% or 1mV referred to 30V.

stability demanded from the circuit shown in Fig. 1 for an overall stability of $V_{D3}$ of $\pm 0.15\%$.

## Experimental results

The performance of the circuit shown in Fig. 1 was evaluated by measuring the currents in each half ($I_1$ and $I_2$) and $V_{D3}$ for various values of supply voltage ($V_s$) and temperature. Measurements were made using a five-digit voltmeter.

In the first experiment, $I_1$ and $I_2$ were set approximately equal (at 30°C) to 3.75mA ($I_z = 7.5\text{mA}$), with $V_s = 30\text{V}$. The temperature was varied from 0 to 50°C with $V_s$ steady. As can be seen from Fig. 3 $I_z$ in-
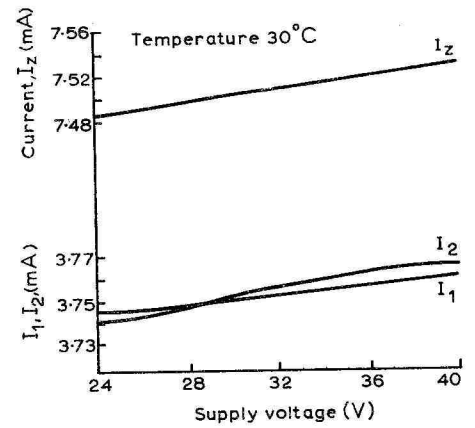
Fig. 5. Relation between currents in the two halves for varying $V_s$ shows very close 'tracking'.

creases almost linearly with temperature, its average slope over the full temperature range being about 0.035%/deg C. The corresponding slope for $V_{D3}$ is about 0.002%/deg C, its overall stability referred to 30°C value being better than 0.065%.

In the second experiment, the temperature was held steady at 30°C and $V_s$ varied over the range 24 to 40V ($I_1 = I_2 = 3.75\text{mA}$ at $V_s = 30\text{V}$). In Fig. 4 the $I_z$ curve has a positive coefficient of about 0.04%/V. The curve for $V_{D3}$ has an overall stability (referred to the 30-V value) of better than 1mV i.e., 0.016%, and within the resolving capability of the voltmeter, this corresponds to a voltage coefficient of less than 0.001%/V.

The relationship between $I_1$, $I_2$ and $I_z$, for variations of $V_s$, is shown in Fig. 5. Currents $I_1$ and $I_2$ track very closely at all points and make almost identical contributions to the total current $I_z$.

## Conclusion

The circuit shown in Fig. 1, when operated at temperatures between 0 and 50°C, and with supply voltages between 24 and 40V, produces a reference voltage which has an overall stability of better than 0.08%.

It is expected that even better results would have been obtained if the dual transistors used in Fig. 1 had matching base-emitter voltages.

## References

1. P. Williams 'Constant-current circuit' (letter). Wireless World, Sept. 1966, p.456.
G. N. E. Pasch, 'Constant-current circuits' (letter). Wireless World, May 1967, p.228.
P. Williams, 'Ring-of-two reference'. Wireless World, July 1967, p.318-22.
2. Motorola Ltd, 'Temperature-compensated reference diodes 1N821-9'. Data sheet DS8007.

# Electronic Building Bricks

## 15. Measuring information

*by James Franklin*

Throughout this series there have been frequent references to 'information' and how it may be represented electrically and processed in electronic systems. By now most readers will have understood that this 'information' is not merely something which we read or hear, but can be a varying physical quantity such as the height of the mercury in a thermometer or an electromotive force coming from a microphone. Within electronic systems information is conveyed as signals, waveforms or electric states. When we design communications or other processing systems it is often necessary to be able to measure this information—or, more precisely, the rate at which the information has to be conveyed. This is because equipment for handling a high rate of information is more difficult to design, and costlier, than equipment for a low information rate (e.g. a closed-circuit television system as against a telephone circuit) so it is uneconomic to provide for a higher information rate than you really need. How, then, are information, and information rate, measured?

Engineers measure information in units called 'bits', which is a contraction of 'binary digits'. Information rate is measured in bits per second (telegraph engineers call them bauds). The binary digit is an element which may have one or other of two distinct states. Represented on paper these could be 'yes' and 'no'; the digits '1' and '0' as in the binary number system;* a black area and a white area; or a hole and the absence of a hole (in punched cards or tape). Represented in electrical form these states could be the 'on' and 'off' states of a switch; two different voltages; two different currents; the presence of a pulse and the absence of a pulse. Such a principle can be applied to any physical variable.

The binary digit is used as a measure of information for two reasons. First, it allows a choice to be made from two entities and thence, by further sub-division, a choice from a whole family of entities. Secondly, in electrical form the two possible states of the binary digit can be represented very clearly and non-ambiguously (e.g. on and off). Fig. 1 shows how a choice may be made of one entity (the letter F) from a family of
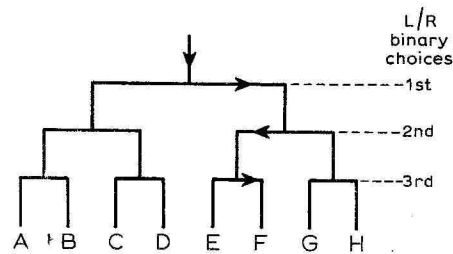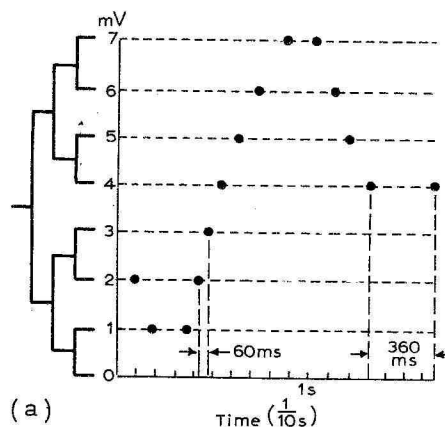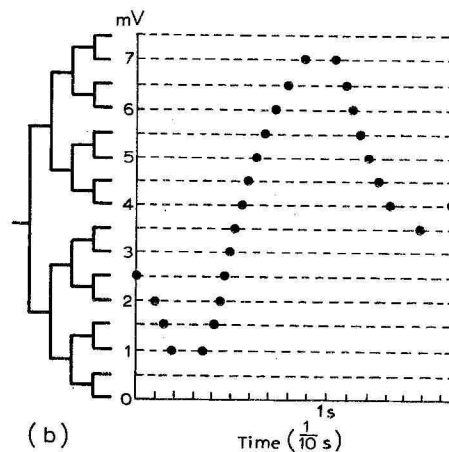


*Fig. 1. Showing how one symbol may be selected from a set of eight by a series of left/right binary choices.*



( a )

(b)

*Fig. 2. The binary choice principle of Fig. 1 applied to values in voltage/time graphs, i.e. signals. At (a) the signal is defined by selection from eight levels of voltage; at (b) by selection from 16 levels.*

entities (an eight-letter alphabet) by a series of three binary choices (left or right). Therefore the number of bits of information contained in the knowledge that one letter has been selected from an alphabet of eight† is 3. If this whole selection is made in, say, a tenth of a second the information rate is 30 bits per second.

Now let us see how this principle can be applied to information in electrical signals. First of all turn Fig. 1 on its side and in place of the eight letters write a family of eight voltages on a scale, then add a horizontal time scale to allow a signal to be represented as a voltage/time graph. The result is Fig. 2(a). We can now select by binary choices any one voltage from a family of eight voltages, and the information contained in the knowledge that this particular voltage has been selected is 3 bits. The signal is not actually drawn in as a continuous voltage/time graph line but is defined approximately by the sequence of points marked where the invisible graph line passes through the individual voltages. If we doubled the number of voltages in the family to 16 the signal would be defined more accurately by more points, as shown in (b), but because more binary choices would be required to allow this, the information content in any one point on the graph would become 4 bits. In theory to define any signal perfectly would require an infinite number of points and voltage levels. In practice it is not all that great; for example, a television signal calls for a minimum of 8 binary choices—8 bits—which means selection from a family of 256 voltage levels.

The information *rate* of the signal in Fig. 2 (a) is determined by the time intervals between the voltage points defining the graph, and here this varies between 60 milliseconds (giving 16.6 bits/second) to 360 milliseconds (giving 2.8 bits/second). In practice the engineer has to allow for the highest information rate necessary for the class of signal he is dealing with. For example, a television signal calls for a maximum information rate of about 11 million bits/second, while a broadcast sound signal needs a maximum rate of about 30,000 bits/second and a telephone signal a maximum rate of 8,000 bits/second.

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* A number system based on the radix 2 instead of the familiar radix 10 of the decimal number system.

† The generalized formula is: number of bits $= \log_2 N$, where $N$ is the number of entities in the family.

# Sampling Oscilloscopes and Sampling Adaptors

## A simple explanation of how sampling is applied to oscillography and the benefits that can be obtained

*by E. B. Callick\* and A. Lawson\**

The design and development of radar, communications equipment, fast computers, counters and timers depends upon accurate display of high-frequency waveforms. Because currently available general-purpose oscilloscopes do not give acceptable performance above 100MHz, special wide bandwidth oscilloscopes have been developed, but their design becomes increasingly complex and expensive as the bandwidth is increased. This is due mainly to the difficulty of designing a cathode-ray tube and deflection system to give adequate brightness and deflection sensitivity. The limit set by the present state of the art is around 250MHz, but within the next few years this may be extended to 500MHz with a corresponding increase in cost.

An alternative way of displaying high-frequency waveforms is called signal sampling which is a means for displaying or recording waveforms which are above the upper frequency limit of the indicating instrument. In a typical case, signals at frequencies up to 1GHz can be displayed using a tube and deflection system with a bandwidth of only 150kHz. The sampling unit can either be part of the oscilloscope or an entirely separate unit.

Unlike a conventional oscilloscope, on which the waveform of the signal to be observed is drawn during a single X-sweep in a time related to the period of the input signal, a sampler builds a replica of the waveform over a period covering many cycles of the input signal. It will be assumed, for the purposes of description, that the input signal is applied to a sampling gate which is opened for a very short time once in each input cycle. Each time the gate is opened the sampler measures the input signal and causes a dot (or sample) to appear on the face of the c.r.t. which represents the amplitude of the input waveform at the time the sample was taken. The sampler has a memory circuit which enables each dot to be displayed until shortly before the next sample is taken. It also provides a scan signal which places each dot at the correct position on the X-axis. The frequency at which the gate opens is made lower than the input frequency, so that each sample represents

\* G. & E. Bradley Ltd.

a different, later part, of the input signal. Thus a replica of its waveform is built from a number of samples taken over a period equal to many cycles of the signal. Because the memory retains a signal representing the amplitude of the sampled waveform, it is necessary only to increase or decrease that signal by an amount representing the increment in signal amplitude between successive samples. This up-dating of the memory is done in a short gating period during and after sampling. It is not essential that a sample be taken during each cycle of the input signal. If the sampling frequency is such that the gate is opened once during every tenth, hundredth or thousandth cycle of the input signal, this will produce a

corresponding increase in the effective bandwidth of the sampling system, but the time taken to build the replica waveform will also increase in the same proportion. This implies that an authentic display will be obtained only when the input signal is time invariant over the period in which image is built up.

Fig. 1 is a simplified block diagram of a typical sampling system. Fig. 2 shows how a replica of one cycle of input signal is produced. To allow time for the trigger circuit to operate, the input signal is

delayed by 50ns before being applied to the sampler. As the trigger and gate generator circuit operate in about 40ns, this gives a 10ns visible delay on the display (i.e. the first sample can be taken 10ns before the signal arrives at the sampler, so allowing the leading edge of pulses to be displayed).

After the initial trigger signal is derived from the input waveform there is a delay of about 40ns before the sampler and memory gates are opened and the first sample is taken. During the 2µs when the memory charges, the display is blanked, the staircase generator advances one step, and the c.r.t. spot moves to the required position where it is displayed until the next sample is taken. The staircase is used for
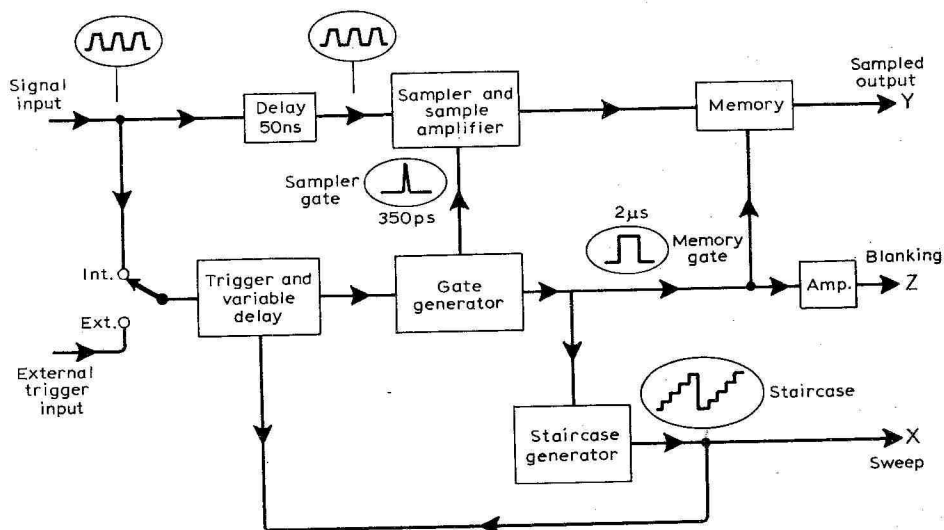


Fig. 1. A typical oscilloscope sampling system.

two purposes; first to position the display spot horizontally during the blanking period and secondly to increase the trigger circuit delay so that successive samples are taken increasingly later after the initial trigger. As this always occurs at the same point on the input signal waveform, the increase in trigger delay with staircase amplitude ensures that successive samples are taken later and later during the input cycle so that the whole of the input signal waveform is sampled as the staircase progresses.
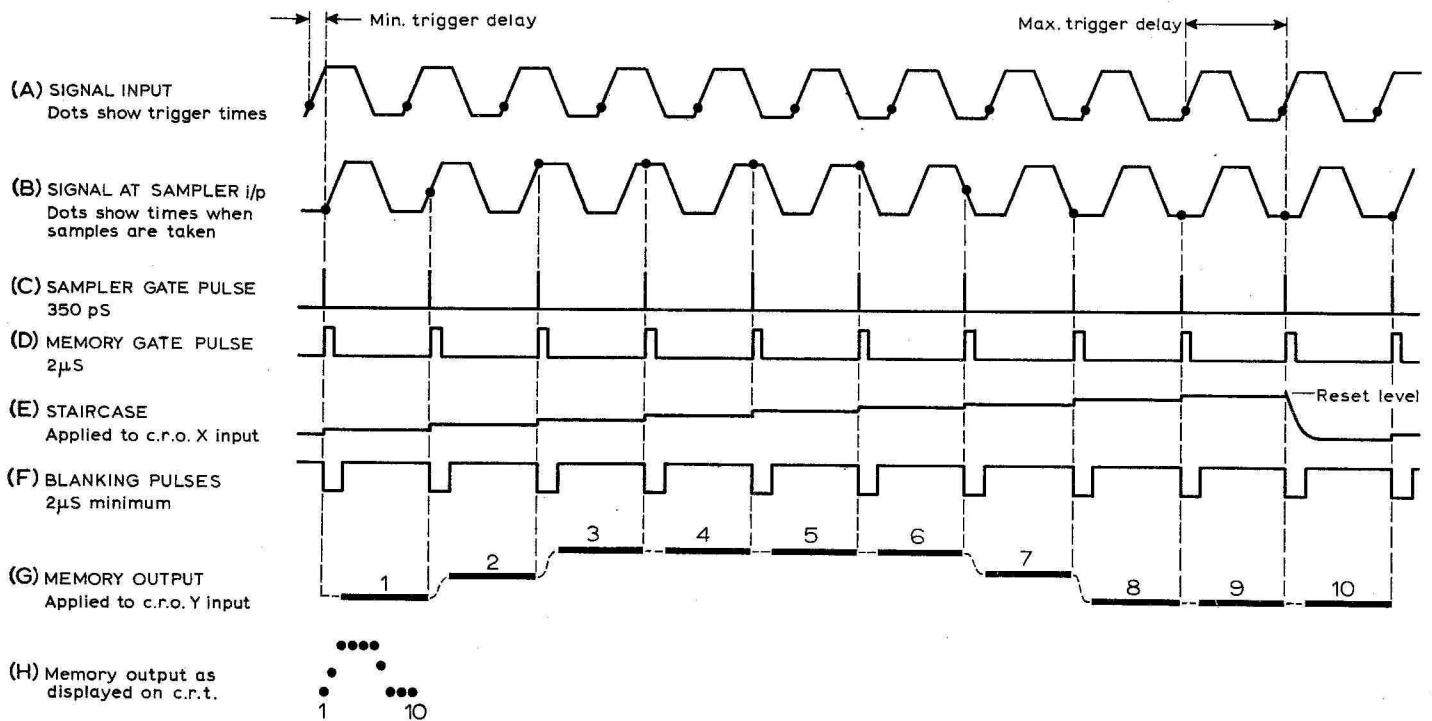
Fig. 2. How one cycle of the input signal is reproduced in a sampling system.

The staircase resets when a fixed level is reached so that a constant amplitude X-scan is obtained. The number of steps per scan can be varied from about 50 to 1000, allowing the display to be built up from any number of dots (samples) in this range.

The effective scan rate of the display is set by adjusting the sensitivity of the trigger variable delay circuit so that staircase steps cause the required incremental delay between samples. The oscilloscope sensitivity is adjusted by varying the gain of the sample amplifier. If a sufficient number of samples is used to build the display, the dots will merge to give a continuous outline as on a conventional oscilloscope.

A typical sampling oscilloscope may have sampler and memory gating periods of 350ps and $2\mu s$ respectively. The minimum time between samples is roughly $30\mu s$. The time taken to build a replica is proportional to the sampling interval, so that this should be kept to the minimum, but this makes design of the gating circuits more complex and expensive. The chosen figure of $30\mu s$ is a working compromise between these two conflicting requirements. When the input signal has a period of $32.35\mu s$ or less, one sample is taken every $32.35\mu s$ so that the time taken for one complete X-scan of 1000 samples is roughly 32ms. For input signals with periods greater than $32.35\mu s$ (frequencies below approximately 30kHz), one sample is taken from each cycle.

At low frequencies this results in a very slow X-scan. For example, an input signal frequency of 1kHz (1ms period) results in an X-scan time of 1 second if 1000 samples are used to build the display. Thus the effectiveness of sampling for visual displays is limited by display flicker for low repetition rate signals unless a long

persistence display tube is used.

The parameters which limit the performance of a sampling system are the signal gating period and the ability of the memory circuit to generate a signal which is at all times representative of the input waveform.

The maximum frequency at which the system will operate is determined by the signal gating period because the sampler output is proportional to the mean signal level during this time. Thus the sampler output will decrease rapidly when the signal period falls below 700ps, and be zero at 350ps. This implies that the frequency response of the system extends well above 1GHz. It is independent of the bandwidth of the indicating oscilloscope provided this is sufficient for it to follow the variation in memory output from sample to sample. With a memory gating period of $2\mu s$ this implies a bandwidth not less than 150kHz. This can be reduced at the expense of brilliance of the trace by extension of the blanking period. In practice, the blanking signal generated usually has a duration slightly longer than the memory gating period, so that acceptable performance can be obtained with oscilloscopes having bandwidths down to 100kHz.

The fidelity of the sampling system is determined by the ability of the memory to be correctly up-dated during its gating period. In simple terms, the memory is a capacitor charged by a control circuit which can deliver a limited current during the gating period.

Accurate representation of the input signal will therefore depend on the difference in amplitude from sample to sample. With a large number of samples per scan this increment will be small, permitting the sampler to build an accurate replica of the input waveform. As

the number of samples is reduced, the increment will become progressively larger, so that ultimately the memory will not be fully up-dated during its gating period. Thus the response of the sampler to a sinewave input will diminish in amplitude as the frequency increases above a critical value, and representation of a fast rising step function be degraded so that the risetime appears longer. The maximum possible number of samples should therefore be used to ensure accurate representation of the input signal. This will be accompanied by a corresponding increase in the time taken to build the replica waveform. If this is unacceptable, the number of samples per scan may be reduced until distortion of the displayed waveform sets a lower limit to the sampling rate. The response of the sampler is also modified by the delay line transmission characteristics, which become a major obstacle at frequencies much above 1GHz.

An understanding of the basic principles of sampling enables a sampling oscilloscope or adaptor to be used as easily and reliably as a conventional oscilloscope. The number of samples per scan used to build a replica of the input signal is typically variable over a range of at least 50 to 1000. This allows the number of samples per scan to be reduced when signals with low repetition rate are examined and so permit building of a replica image in a reasonably short time. Degradation of the waveform which occurs when the number of samples per scan is insufficient to allow an accurate replica to be built may cause inexperienced users to doubt the authenticity of sampled displays. Correct operation is obtained when the maximum possible number of samples per scan is used. Authenticity is then limited by the

intrinsic capability of the instrument.

Against the obvious advantages of sampling both from the operational and cost points of view must be set two inherent properties of sampling systems which may prove to be disadvantageous in some cases. First, 'single shot' operation is not possible, as samples must be taken from many input signal cycles to build a display. Secondly, the scan rate is slow when the input signal repetition rate falls below about 1000Hz. The effect of slow scan rate can be largely overcome by using a c.r.t. with a long-persistence phosphor so that display flicker is reduced. It should be noted that slow scan rate is

an advantage when it is required to record the sampled waveform, as a wide bandwidth recording system is not required.

An oscilloscope or sampling adaptor such as we have considered is ideal for measurements of c.w. and pulsed waveforms in v.h.f. communications and radar equipment. The typical fastest effective sweep rate of 0.1ns/cm enables fast computer and counter logic waveforms to be examined in detail, and time measurements such as signal path delays and semiconductor signal transit times to be made easily and accurately. Circuit faults caused by parasitic

oscillations or ringing due to fast transients often cannot be detected with general purpose oscilloscopes. Such effects are easily located with a sampling oscilloscope which will often bring to light unsuspected design faults.

A sampling oscilloscope or sampling adaptor is therefore a good alternative to a general purpose oscilloscope at frequencies up to 50MHz provided that the input signal repetition rate is above about 100Hz. At higher frequencies its performance is superior to that of expensive special purpose wideband oscilloscopes except when 'single shot' displays are required.

# Sound Synthesizers

A sound synthesizer comprises a system of voltage controlled oscillators and amplifiers, modulating networks, and combining and keying facilities. For a synthesizer to be of value to a composer the sound generated must be fully prescribed by switch positions. Only then can the system be brought under sensible control.

Three new synthesizers have been introduced in the U.K.—two as imports from America, the third home grown.

Tonus of Massachusetts make the ARP 2500 and the smaller 2600 systems, both available from F.W.O. Bauch Ltd, 49 Theobald St, Boreham Wood, Herts.

The 2500 system for all its complexity and versatility avoids 'patchcords' by employing a modular bus-bar system with midget slide switches. The input, output and control of each module is determined by a vertical slide that connects it to any horizontal bus-bar.

In this manner controls can be cascaded and waveform shapes combined in almost any pattern. The 2600 system combines keyboard and sound generators in a neat portable assembly operating from the mains. Bauch are holding a series of lecture-demonstrations and readers can ring 01-953 0091 for details.

From Electronic Music Studios (London) Ltd (49 Deodar Road, S.W.15), the Synthi A attache case synthesizer sells at less than £200 and provides a considerable variety of effects, as may be judged from the photograph.
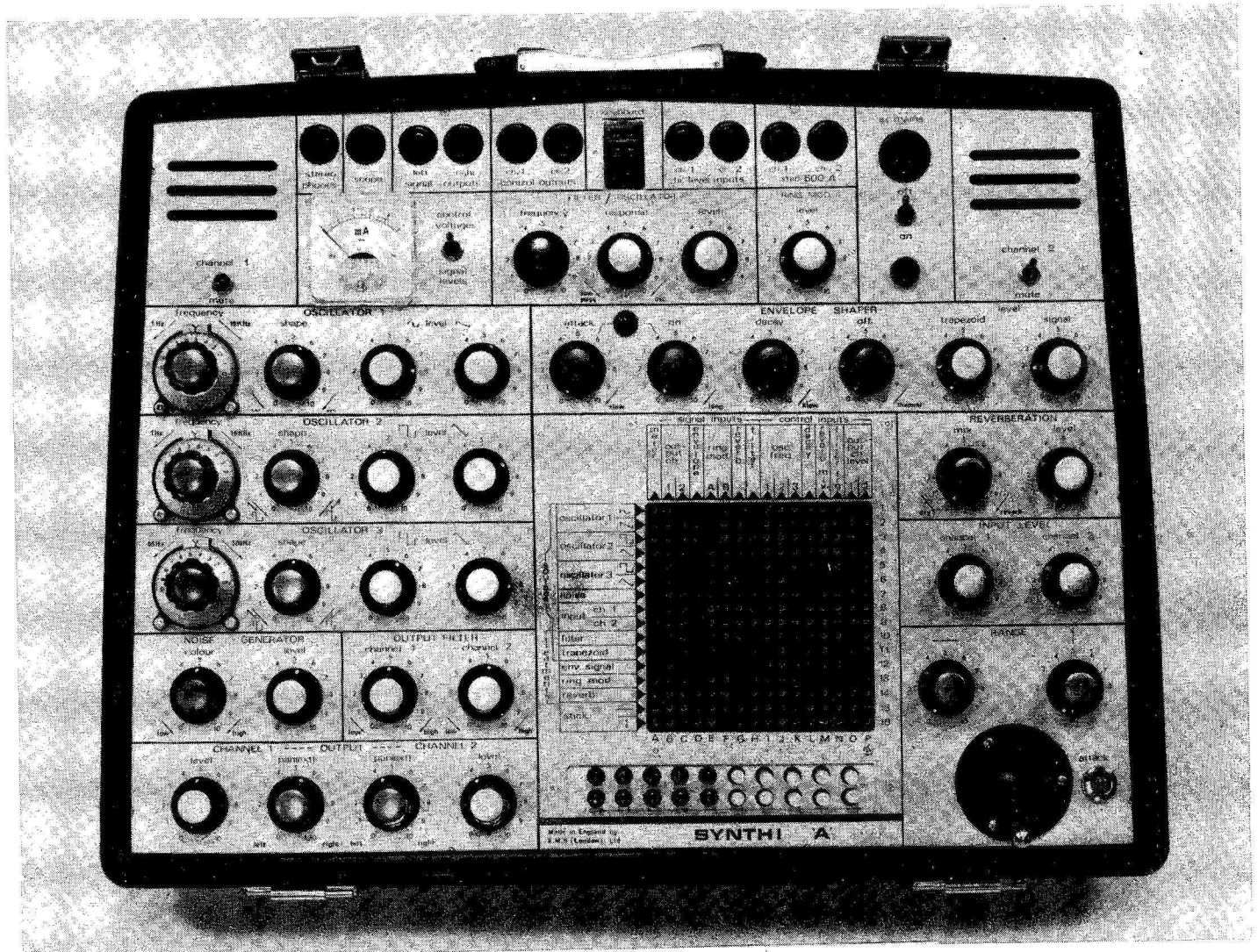
# Elapsed Time Graph for Tape Recording

## A simple method for determining the remaining recording time on partially used tapes

*by B. W. Lingard\**

Any user of a tape recorder will, sooner or later, wish to know the length of recording time still available on partly recorded tracks. If he has been methodical and noted the duration of existing recordings the answer is simple—if not, it is only very approximately obtainable. Graduated "protractors" are available which can be fitted on top of the spool and the time read off. However, the graduations are extremely close over the outer third of the reel and are correct only when the reel hub is of the correct diameter and the tape of the nominal thickness. The digital counter reading which on most recorders is proportional to the number of turns of the left-hand supply spool, has no linear relationship to recording time. A straight line graph is not obtainable even if logarithmic graph paper is available. A graph (curved) can be plotted on linear graph paper, but a different graph will have to be plotted for each reel size and tape thickness—in some cases for different makes of tape, because of the variation in hub diameters and tape thickness. What are the relationships concerned?

A reel of tape when full has $N_T$ turns and an outer radius of $R_2$ inches. If the hub radius is $R_1$ inches it follows that the mean radius is $(R_1 + R_2)/2$ and that the tape length

$$L_T = 2\pi \frac{R_1 + R_2}{2} N_T = \pi N_T (R_1 + R_2) \quad (1)$$

However, if the tape thickness is $T$ inches it is also apparent that

$$N_T = \frac{R_2 - R_1}{T} \text{ and hence } \frac{1}{T} = \frac{N_T}{R_2 - R_1} \quad (2)$$

If $N_1$ turns are supplied from this reel (on the l.h. spool) the radius falls from $R_2$ to $R_2 - N_1 T$, and the length delivered is

$$L_1 = 2\pi \frac{R_2 + R_2 - N_1 T}{2} N_1$$

$$= \pi N_1 (2R_2 - N_1 T) \quad (3)$$

Recorded time is proportional to length so that

$$\text{Time} \propto \frac{2R_2}{T} N_1 - N_1^2 \quad (4)$$

and the relationship is of the form

$$y \propto Ax - x^2.$$

\*Royal Military College of Science, Shrivenham.

Strangely enough a suitable graph can be constructed using a square law graph upside down! Consider, with a square law graph each ordinate is placed at a distance from the l.h. origin proportional to the square of the number of the ordinate i.e. $s \propto x^2$. If such a graph is constructed up to the value $x_T$ and then inverted the ordinates will now be found to be distant from the new l.h. origin by

$$s \propto x_T^2 - (x_T - x)^2 = 2x_T x - x^2 \quad (5)$$

It follows that if a square law graph is constructed and inverted a graph of time (vertical linear scale) against counter reading will plot as a straight line provided that:

$$\frac{2R_2}{T} = 2x_T \text{ or } x_T = \frac{R_2}{T} \quad (6)$$

In practice $T$ varies between makers (for the same nominal thickness of tape) and it is best to substitute from (2)

$$x_T = \frac{R_2 N_T}{R_2 - R_1} \quad (7)$$

However an additional complication arises in that the counter does not usually count turns directly. If $N = kN'$ (where $N'$ is the actual counter reading) (4) above is more properly expressed:

$$\text{Time} \propto \frac{2R_2 k}{T} N'_1 - k^2 N'_1{}^2 \quad (8)$$

and $x_T = R_2/kT$. But also $N_T = kN'_T$ so that

$$x_T = \frac{2R_2 k N'_T}{k(R_2 - R_1)} = \frac{R_2 N'_T}{R_2 - R_1} \quad (9)$$

For one specific tape recorder $x_T$ is found to vary as follows:

|            | S.P. | L.P. | D.P. |       |
| ---------- | ---- | ---- | ---- | ----- |
| 5 in.      | 1450 | 2010 | 2880 |       |
| $5\frac{3}{4}$ in. | 1800 | 2310 | 3650 | $x_T$ |
| 7 in.      | 2050 | 3060 | 4280 |       |

If a standard play tape is not normally used a value of $x_T = 3200$ will be found to give acceptable results.

The graph is constructed as follows:
1. A convenient base to start from is 0–8 with quarter sub-divisions.
2. The vertical graph lines are set out distant from the l.h. origin as follows

| 1·00 | 1·56 | 2·25 | 3·06 | 4·00 etc. |
| ---- | ---- | ---- | ---- | --------- |
| $(1^2)$ | $(1·25^2)$ | $(1·5^2)$ | $(1·75^2)$ | $(2^2)$ |
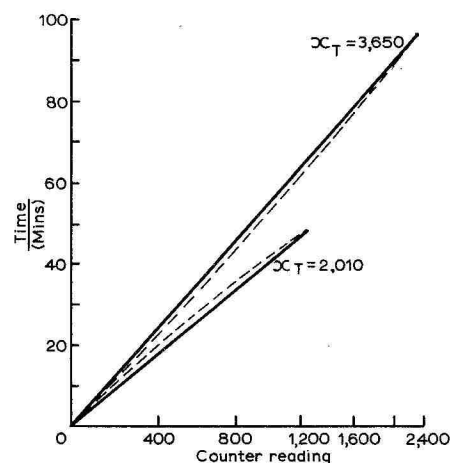


up to 64·00 ($= 8^2$) a suitable scale factor (C) being chosen such that 64C is somewhat less than the width of the page.
3. The graph is then inverted and the ordinates labelled 0, 100, 200, etc. Note that the penultimate ordinate is 2800 and the last 3200.
4. The horizontal lines are evenly spaced and numbered from 0–130 to fill the space available. This will be suitable for $3\frac{3}{4}$ i.p.s. tape speed and the normal range of reels.
5. Finally, for any reel of tape of a given size and time gauge, run the tape off the l.h. spool and note $N'_T$. Plot on the graph a point $N'_T$/nominal time and join it to the labelled origin with a straight line. Generally any time obtained from the graph for a given counter reading will be found to be within 2 minutes of the correct value. Further straight lines can be constructed for other reel sizes and types. It should be noted that at $3\frac{3}{4}$ i.p.s. nominal times for 1800, 1200 and 900 feet are 96, 64 and 48 minutes respectively. Note also that "$x_T$" relates to the length of the base line of the original graph in arbitrary units. It is not the same as $N_T$, nor $N'_T$, and in all cases given $N'_T$ (which is the actual point plotted) will be found to be less than 2400 and hence easily accommodated.

A typical graph is shown in the illustration. Where a specific reel size and tape gauge is always used the graph can be constructed with the correct $x_T$ when full agreement between graph and measured time will be obtained.

# Centimetric Television Broadcasting

## Experimental 12GHz transmissions

*by J. C. G. Gilbert, F.I.E.R.E.*

At the Radio Administrative Conference held in Geneva in 1959 the centimetric band of 11.7 to 12.7GHz was reserved for several services including television broadcasting. The German Post Office Telecommunications Research Institute started an investigation into propagation problems in this band and in 1969 Dr. J. Feldmann, who led this investigation, read a paper at the Montreux Television Symposium on the feasibility of TV broadcasting in Band VI*.

Most of the available channels in the v.h.f. and u.h.f. television bands are already in use in Germany, and as she wishes to increase the number of programmes two possible lines of attack are open. One is the possible use of stationary satellites and the other is to explore the use of centimetric transmissions from ground stations. Research spread over several years followed three main topics: (1) the propagation behaviour of centimetric waves, (2) the technical conditions to be satisfied at the transmitter and (3) the technical problems at the receiver.

Centimetric waves in the order of 2.5cm behave like light waves and are reflected by obstacles and greatly attenuated by roof structures and walls thus making the use of room aerials impracticable. The atmospheric effects of rain and fading measured over a long period indicate that for 1% of the time a loss of 0.4dB per km can be expected, while for 99% the propagation is hardly affected.

The research team decided that the transmitter should have the following objectives: it must be capable of high-quality pictures and that in order to make the system economically viable the receiving equipment should be cheap, simple, require the minimum of maintenance, and that current commercial TV receivers should be readily adaptable. These considerations therefore defined the transmitter as using vestigial sideband amplitude modulation for vision and frequency modulation for sound. For the earlier measurements a low-power transmitter of some 3-4W was used in which both the video and audio signals were combined

* *Wireless World,* July 1969 p.325.

at low level. This system can be used economically up to 100W. For higher power transmitters up to 1kW using multi-cavity klystrons, it has not been found practicable to amplify the vision and sound channels in one tube without introducing cross-modulation. Therefore for higher power transmitters the video and audio channels are kept separate and are combined only at a directional coupler that feeds the transmitting aerial. The specification of the transmitter now in use for experimental transmissions is:

| | |
|---|---|
| Power output | 0.1 to 1kW |
| Modulation | vision—vestigial sideband a.m. |
| | sound—C.C.I.R. standard f.m. |
| Range | 10-15km |
| Signal-to-distortion ratio for cross modulation | 51dB |
| Frequency band | 11.8 to 12.2GHz |
| Transmitter aerial | omni-directional or aerial with sector-shaped pattern and cosec characteristic |
| Signal processing | vision and audio together (low power) or separate for high powers |
| Polarization of far electrical field | vertical |
| Stability of transmitter | better than ± 100Hz per month |

At the receiver the signal is converted into a spare channel in Bands I, III, IV or V. The receiving aerial uses a parabolic reflector which has a gain of 25-35dB for a diameter of 65cm. The side lobe attenuation in the range of ± 10° off the main beam is > 20dB and for the remaining range > 25dB. Between the output from the aerial and the mixer stage is a band limiter to improve the signal-to-noise ratio, and also prevent the local oscillator radiation. The local oscillator frequency is dependent on the receiver channel to be used, and the stability is stated to be better than ± 75kHz per year. The bandwidth of the converter is at least 80MHz which gives a total of eight possible channels.

The most important criterion of the receiver converter is that it should have as low as possible a noise figure and freedom from distortion. The use of a push-pull mixer reduces noise considerably as it
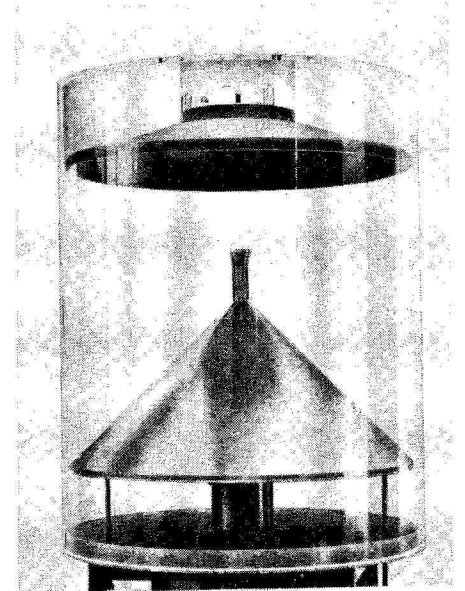


*Omnidirectional transmitting aerial housed in plastic. Signals are reflected from the 'roof' to the cone and radiated.*

suppresses the f.m. noise of the local oscillator. Fortunately the atmospheric and cosmic noise in the 12GHz band is low and with a vertically polarized receiving aerial it remains below 200°K. Where one is dealing with a large communal system it becomes economical to use a parametric pre-amplifier which improves the noise figure, but for single receivers Schottky-barrier diodes are used.

Stability of the local oscillator is required to attain a very high standard, and a simple free-running microwave oscillator may vary several megahertz in an hour. In order to achieve the necessary stability a relatively low-frequency crystal oscillator is employed followed by frequency multiplier stages. Provided that mass produced harmonic crystals are aged they are satisfactory and in order to prevent warming up drift, the power to the crystal oscillator is always connected. A pilot signal is radiated by the transmitter as a reference signal which can be fed to the local oscillator thus maintaining its stability within the required limits. Both the pilot signal and the TV signal are converted to the i.f., amplified and the pilot frequency extracted and fed to a frequency discriminator. The output from the frequency
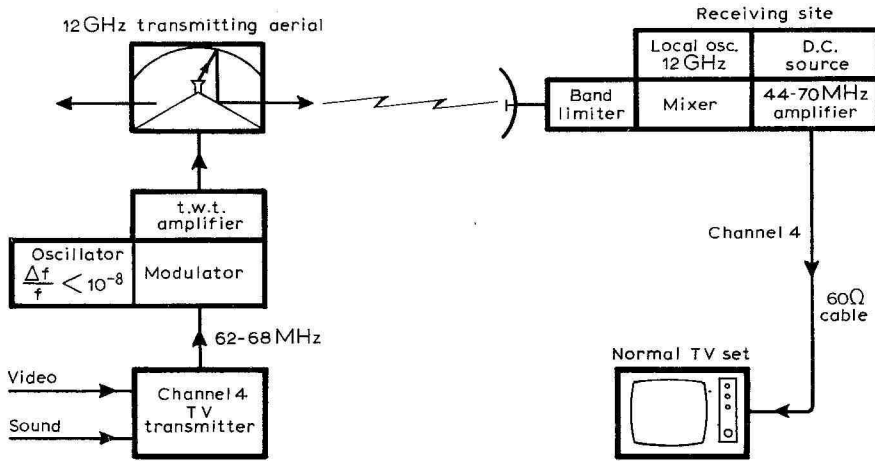
*Block schematic of 12GHz transmitting and receiving system.*

'discriminator then provides a control voltage for the stabilization of the local Gunn oscillator.

Research is continuing with alternative methods of frequency stabilization and two suggested methods use a cavity resonator with an extremely inflexible glass construction or alternatively a cavity resonator using a gas pressure controlled membrane which compensates for changes of temperature.

Considerable effort is being applied to the problems of the receiver installation, which demands accurate siting and positioning of the aerial-converter unit. Wind resistance of a solid paraboloid demands a rigid, guyed mast with means of directing the aerial within 1° to the transmitter. Alternative designs make use of a wire mesh paraboloid using very thin rustproof wire with the crossing points welded and with a stiffening rim. The paraboloid can be mounted either at the top of the mast or in front of it. A rectangular waveguide is used as a feeder and it can be terminated either with a horn or preferably with a circular reflector disc about 3cm in diameter which is supported on a hollow dielectric support. By using miniaturizing techniques the mixer and i.f. amplifier can form part of the waveguide and only the local oscillator is mounted behind the paraboloid.

An alternative form of receiving aerial is a slotted waveguide, and this is acceptable in high field strength areas as its gain is only 15dB compared with a horn-paraboloid combination of 35dB.

During a recent visit to the new German Post Office Research Centre demonstrations were given of reception from three transmitters located some 15km from the receiver. The weather varied from heavy drizzle to rain but the standard of the received picture was of a very high order. We were also given the opportunity of inspecting a mobile field strength van which has a telescopic mounting for the aerial 40 metres high. Also mounted on the top of the mast is a television camera in line with the receiving aerial. Remote control from the interior of the van enables the operator to rotate the mast head and to automatically record the received field

strength from the remote transmitter on an X-Y plotter. Intervening tall buildings are viewed on the television monitor from the mast head camera, and a correlation made with the plotter.

Currently three transmitters are in operation and about 100 receivers placed at strategic positions to assess the variations of received quality with changing atmospheric conditions. It is thought that by mass production methods the cost of the aerial-converter can be as low as £15-£20 plus the cost of the guyed mast.

Demonstrations of the reception of these transmissions will be given during the Radio & Television Exhibition in Berlin from August 27th to September 5th.

Acknowledgement is made to Dr. J. Feldmann and his colleagues at the Fernmeldetechnisches Zentralamt in Berlin for much of the information given in this article.

# Books Received

**D.C. Amplifiers** by B. Mirtes, edited in translation by E. W. Firth. The work is primarily concerned with explaining analysis, design and application of directly-coupled differential operational amplifiers employing semiconductors, and of single-ended drift-corrected op-amps. There is a brief treatment of op-amps using thermionic valves. Other d.c. amplifiers covered include directly-coupled amplifies without feedback, sensitive chopper-type amplifiers, electrometer amplifies, d.c. voltage and current stabilizers, and drift-corrected amplifiers designed to amplify low-level floating voltages. The contents fall into three parts. The first includes a practical and theoretical discussion of electronic devices. The second deals with fundmentals of analysis and design of directly-coupled, amplifying

circuits and systems. The third part discusses directly-coupled, chopper-type and drift-corrected operational amplifiers. There are six pages of bibliography and a twelve-page index. Pp.520. Price £4.50 (cased version only). Iliffe Books, Butterworth & Co. (Publishers) Ltd, 88 Kingsway, London WC2 6AB.

**Selected Papers on Frequency Modulation** edited by Jacob Klapper. This collection is divided into four sections—general f.m. theory and basic experiments, f.m. circuit theory, f.m. threshold reduction, and digital f.m. Armstrong's famous paper "A Method of Reducing Disturbances in Radio Signalling by a System of Frequency Modulation" opens the first section. It was Armstrong who first successfully used f.m., demonstrating its greater immunity to noise interference compared with a.m. systems. The compilation is intended as a "reference work for the practitioner, as a guide for those interested in entering the field, and as a textbook in f.m. principles". Forty further references are given in a bibliography at the end. Pp.417. Price £3.75. The imprint is Dover Publications, Inc., but it is available in the U.K. from Constable and Co. Ltd, 10 Orange Street, London WC2H 7EG.

# Corrections

**Charging.** We apologise to readers and to 'Cathode Ray' for the inclusion by the printers of a wrong diagram for Fig.1 on p.391 of the August issue. Here is (we hope!) the correct diagram of 'the familiar circuit used to study the charging of a capacitor'.
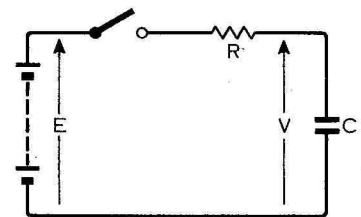


**Stereo Mixer.** In Fig.8(a), Part 1, May issue, $R_5$ is a 'select on test' resistor in the range 200-700$\Omega$ or a 1k$\Omega$ preset adjusted for 16.5V at the emitter of $Tr_3$. The voltage at the emitter of $Tr_5$ in Fig.7 is 16.5V. For Fig.7 66dB s/n ratio is referred to 450$\mu$V on 200 $\Omega$ (not 45$\mu$V as incorrectly stated on p.300 (June issue)). In Part 2 (June) the series resistor to the main balance control, Fig.11, should be 4.7k$\Omega$ not 100$\Omega$, the bass control should be a 100k$\Omega$ lin. not 10k$\Omega$, and the emitter resistor of $Tr_{14}$ 15k$\Omega$. The residual noise level for Fig.11 is $-98$dB, not $-93$dB as quoted in col. 3 p.296. In Fig.13, a 0.22$\mu$F coupling capacitor should be connected between the first 22k$\Omega$ resistor and the input to give d.c. isolation. In Fig.19(a) the reservoir capacitor should be 2000$\mu$F at 50V and the current limit is set by $V_{BE2}/R_1$, not $V_{BC2}/R_1$ as stated in the text col.2 p.298.

**Darlington Output Transistors.** In the protection circuit for use with complementary Darlington output transistors (August issue, p.399) the two complementary transistors were incorrectly shown as MPS1000. They should be MPSA20 (n-p-n) and MPSA70 (p-n-p).

# World of Amateur Radio

## Amateur satellite service

A new 'amateur satellite service' has been defined internationally and amateurs will be able to conduct space communications experiments on 7, 14, 21, 28, 144 (already in use) and 435-438 MHz and 24 GHz bands, including the use of geo-stationary orbits. These, then, are the main changes in the world of amateur radio which will result from decisions made at the I.T.U. World Administrative Radio Conference in Geneva*. This outcome is a considerable improvement on what, at one stage, seemed likely. As reported last month, the delegations from a number of European countries—particularly those from Western European countries most closely associated with the Conference of European Posts and Telecommunications (C.E.P.T)—placed little value on the amateur radio service; indeed in some cases this amounted to active hostility towards amateurs. It was only at the last minute—in the Plenary sessions—that many of these improved facilities (at present amateurs can officially conduct space experiments only in the 144 MHz band) were secured by a reversal of some of the recommendations of the Working Parties. Proposals that amateurs should be permitted to use their 1215, 5650 and 10,500 MHz bands for space experiments were however not accepted.

Many amateurs feel the need to place on record that their proposals received notable support from the official U.K. delegation, led by Don Baptiste, of Minpostel, and from such countries as New Zealand and the United States. The attitude of the C.E.P.T. administrations appears to be in the tradition of earlier I.T.U. conferences and accounts for the serious disadvantages under which amateurs in Region I operate.

In a press interview after the ending of the Conference, Mr. Baptiste is quoted in *The Times* as saying of amateurs: "They provide a laboratory of thousands of enthusiasts all over the world and undoubtedly add to the sum of human knowledge".

The presence at Geneva of amateur advisors—such as Roy Stevens, G2BVN—financed by the national amateur radio societies, undoubtedly helped to reverse some of the adverse recommendations of the working parties.

## British slow-scan TV activity

We have referred several times to the growing interest in the U.K. in international slow-scan television operation on the h.f. bands, in which a picture is sent every 7.2 seconds with narrow-bandwidth. One of the most successful British exponents of this art is H. Jones (G5ZT and G6ABC/T) of Eggbuckland, near Plymouth. He has had many two-way 'television' exchanges with amateur stations all over the world, including over 100 in the period April to June. Pictures have been exchanged with KL7DRŻ (Alaska), VK6ES (Western Australia), KP4GN (Puerto Rico), ZL₁AOY (New Zealand) a number in Italy and Greece and very many in the United States. Many of his contacts represented the first time British s.s.tv. pictures had been exchanged with stations in the countries concerned. His station is a mixture of home-built and commercially manufactured equipment including Trio transmitter and receiver. According to the latest figures, there are now over 200 stations licensed for amateur television, although the number concerned with



*Slow-scan picture received by H. Jones from the United States (W4LAS).*

slow-scan transmission is still quite small. Minpostel is believed to be sympathetic to the view that means should be found to allow amateur double-sideband TV transmissions to continue when the 70-cm band is narrowed.

## More long-delay echoes?

Two years ago ('W.o.A.R.' August 1969), we drew attention to the efforts of a team at the Radioscience Laboratory, Stanford University, California, to enlist amateur co-operation in re-opening the 40-year-old mystery of long-delay echoes of periods up to and sometimes well beyond five seconds. Such echoes were originally reported by Stormer and Van der Pol in the 1920s.

During the past two years a significant number of new instances of apparently authentic echoes of this type have come to light, including several reported by British amateurs, and the number of useful reports is now approaching 100. There have also, it must be said, been a number of reports made in good faith which have later proved to have been the result of elaborate hoaxes. Several possible mechanisms for this strange phenomenon have been postulated, including 'way-out' theories that these echoes may be deliberately induced by space probes coming from outside our solar system, although the Stanford investigators believe that the eventual explanation may prove far less spectacular. The team is still seeking any further details of these rarely occurring (if in fact they do occur) echoes.

## In brief

M. G. Whitaker, G3IGW, has recently worked several South American stations on 1.8 MHz and also ZD8AY in Ascension Island bringing to 50 the number of countries he has worked on 'Top Band'. . . . Eric Trebilock, a long-time keen listener to amateur stations who lives in Australia, has now had over 300 countries confirmed—a remarkable score for a non-transmitting amateur. . . . Peruvian stations have been authorized to use the prefix OB instead of OA this year to mark 150 years of Peruvian independence. . . . Extended range v.h.f. conditions were much in evidence on 144MHz during mid-July with many West European stations received in southern England. . . . The prefix JE is now in use in Japan. . . . The Scottish V.H.F. Convention is to be held at the Carlton Hotel, Edinburgh, on Sunday, October 3 with speakers including Tom Douglas, G3BA, and Geoff Stone, G3FZL, and there will be an exhibition of equipment (details from V. M. Stewart, GM3OWU, 9 Juniper Avenue, Juniper Green, Midlothian EH14 5AJ).

PAT HAWKER, G3VA

# Personalities

**Stephen S. Forte,** B.Sc., Ph.D., F.I.E.E., and **Robert Pace** have been appointed joint managing directors of General Instrument Microelectronics Ltd following the resignation of **G. Brookes.** Dr Forte joined G.I.M. in 1970 as marketing director having previously been with Marconi-Elliott Microelectronics since its formation in 1964 where he held successively the posts of applications engineering manager, manager for custom circuits and, finally, manager of the m.o.s. products division. Dr. Forte spent several years in the Marconi Company R & D Laboratories prior to transferring to M.E.M. Mr. Pace, who holds several of the basic patents issued in the m.o.s. field, has been with General Instrument Corp. (parent company of G.I.M.) since 1965 where he held a number of posts including director of engineering and, latterly, assistant to the general manager of the m.o.s. division. Prior to 1965 Mr. Pace was head of engineering at General Microelectronics. G.I.M. was formed in 1968 to design and manufacture m.o.s. large-scale integrated circuits for the U.K. and E.F.T.A. markets.

**J. Stuart Sansom,** M.I.E.R.E., technical controller of Thames Television (one of the I.T.A. programme contractors for London), is the 1971/2 chairman of council of the Royal Television Society. Mr. Sansom, who is 42, spent two years with the Royal Corps of Signals before joining E.M.I. In 1953 he joined High Definition Films where he worked for four years on telerecording equipment. He then joined Television Wales and the West and in 1959 went to A.B.C. Television where he became chief engineer in 1966.

**Ian C. Macarthur** has been appointed managing director of the Service Division of RCA Ltd which he joined in 1961. Mr. Macarthur, who is 35, was formerly manager of the Service Division's government and project

services. He was most recently responsible for all installation, operation and maintenance projects of the company, including the ballistic missile early warning system, the Suffolk radio research facility, the European Space Research Organization station in Redu, Belgium, and the Skynet S.R.D.E. station at Christchurch. Mr. Macarthur replaces **Warren Werner,** who is returning to the United States to take up a new position in the Service Division's International Marketing Organization.

Semicomps Ltd, of Wembley, Middlesex, have appointed **Tony Manning** as sales manager. Mr. Manning was with Mullard for 13 years where he was commercial product manager for discrete semiconductors. Before he joined Mullard, he had five years' experience as a development engineer in guided weapons with G.E.C. at Stanmore.

**Leonard F. Knott** (43) has joined Minster Automation Ltd, of Wimborne, Dorset, as chief engineer. He joins Minster from Plessey, where he was latterly responsible for the engineering of Ministry contracts in the fields of transmission lines and logic switching for use in data handling. His technical experience includes eight years with the Post Office Engineering Department, national service with the Royal Navy Electrical Branch and fourteen years on telephone switching and remote control systems.

**J. Don Sinclair** was recently appointed managing director of Astro Communication Laboratory (U.K.), of Coventry, the U.K. subsidiary of Aiken Industries Inc. Astro manufacture surveillance and telemetry receivers and computer peripherals. Mr. Sinclair was previously with Litton Industries as vice-president and general manager of Litton Precision Products International Inc., the European sales and

marketing group for electronic components and microwave products. He was at one time a director of Amplivox and also has been head of facsimile communication sales with Muirhead. His engineering background in electronics was in microwave systems development at the Cavendish Laboratory, Cambridge.

Exel Electronics Ltd, who recently moved from Reading to Branksome, Poole, Dorset, have announced the appointment of **Roy S. Bibby** as sales director and **Ray J. Chapman** as production director. Mr. Bibby, who is 40, joined Exel in June 1970 from Coutant Electronics, to set up and develop a marketing team for the company's new range of digital panel meters. He served with the Royal Signals and spent seven years with Advance Electronics digital division before joining Coutant. Mr. Chapman (31) also joined Exel in June 1970 from Coutant Electronics, to act as production manager. He served his apprenticeship with Fairey Aviation and worked as a draughtsman with Dawe Instruments and design engineer with De La Rue Frigistor.

**Daphne F. Jackson,** D.Sc., F.Inst.P., A.R.C.S., reader in nuclear physics in the Department of Physics in the University of Surrey, has been appointed professor and head of the department. Dr. Jackson, who is 34, is believed to be the first woman to be appointed as head of a physics department in any university in the U.K. She took her degree at Imperial College in 1958, and went to the University of Surrey, then Battersea College of Technology, to take her Ph.D. in the field of theoretical nuclear physics. She joined the staff as an assistant lecturer in 1960 and was appointed reader in nuclear physics in 1967. During 1963–64 she visited the University of Washington, Seattle, as research assistant professor and has just accepted an invitation to become visiting professor to the University of Louvain, Belgium.

**Bryn Tinton,** who joined Ericsson Marine U.K. as technical co-ordinator in March, is in charge of the new training programme for ships' radio officers now being provided by Ericsson Marine, at the Norway Trade Centre in Pall Mall, London. Before joining Ericsson he spent five years with Cunard Brocklebank, latterly as senior radio officer. He has held an amateur radio licence for eight years. His call sign is G3SWC.

**K. R. Sturley,** Ph.D., B.Sc., F.I.E.E., who has been professor of communications and head of the Electrical Engineering Depart-

ment of the Ahmadu Bello University, Zaria, Nigeria, for the past three years, is returning to the U.K He has completed his work as chairman of a technical committee of Nigerian telecommunications engineers set up by the Federal Military Government to advise them on the modernization of Nigerian broadcasting. A graduate of Birmingham University Dr. Sturley obtained his doctorate for research in electrothermal storage problems. In 1936 he joined the staff of the Marconi College, Chelmsford, and was assistant principal when he left in 1945 to join the B.B.C. as head of the engineering training department. From 1963 to 1968 he was chief engineer of external broadcasting in the B.B.C.

**Peter Sinclair,** who joined Circaprint Ltd, the printed circuit designers and manufacturers of Maidstone, Kent, last year from Palmer Aero Products Ltd, has been appointed sales manager. Mr. Sinclair, who is 47, was sales manager of the printed circuits division of Palmer Aero Products.

**A. R. Pritchard** has joined English Electric Valve Co. Ltd as sales engineer with responsibilities for power valves, power klystrons and vacuum capacitors. Mr. Pritchard was previously with The Marconi Company for 10 years, latterly as sales engineer in the radio communications division.

Recently announced academic appointments include the following: **David S. Campbell,** D.Sc., technical manager, capacitor division of the Plessey Company, has been appointed to a chair of electrical engineering at Loughborough University of Technology. **K. D. Stephen,** B.Sc., F.I.E.E., senior lecturer in the department of electrical and electronic engineering in the Heriot-Watt University, Edinburgh, has been appointed to the new full-time post of director of television at the University. **Professor J. H. H. Merriman,** C.B., O.B.E., F.I.E.E., senior director, telecommunications development, in the Post Office has been appointed by the I.E.E. to serve for four years on the governing body of the Imperial College of Science, London. The University of Birmingham has appointed **R. Mellitt,** B.Tech., and **A. W. Rudge,** Ph.D., to be lecturers in electronic and electrical engineering. At the Heriot-Watt University **P. H. Etherington,** B.A., (asst. lecturer at Kenya Polytechnic) is to be a lecturer in the department of electrical and electronic engineering, and **J. Helszajn,** M.Sc., Ph.D., is to be a part-time senior research fellow in the department.
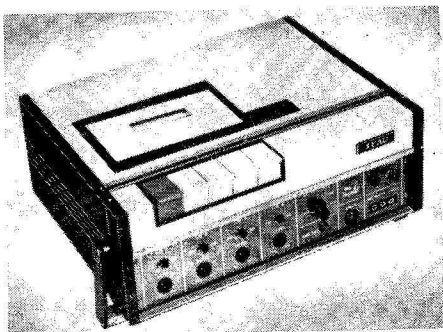
# New Products

## H.F. linear amplifier

Racal-Mobilcal's TA-940 100-watt h.f. linear amplifier has been designed to increase the power output of low- and medium-power h.f. s.s.b. manpacks. Coverage of the h.f. range of 1.6 to 30MHz is provided and continuous 'key-down' operation is possible to full specification – 100 watts output for entire duty cycle. The amplifier will operate with inputs, pre-set internally. between 10mW and 5W. Operation is from a negative earth 28V d.c. power supply. An aerial tuning unit and range of aerial systems are available for use. Racal-Mobilcal Ltd, 464 Basingstock Road, Reading, Berkshire, RG2 0QU.
**WW309 for further details**

## Cassette data recorder

A standard Philips ⅛in tape cassette is used on the TEAC R-70 recorder (marketed in the U.K. by the Industrial Import Division of Dodwell Ltd) to provide simultaneous four-channel recording, using f.m. or a.m., with the additional facility of putting announcements on to channel four, using a microphone. The recording range is 0.1 to 625Hz (f.m.) and 100Hz to 8kHz (a.m.) with a tape speed deviation of ±1% at 4.75cm/s (1.875in/s). Wow and flutter is 0.5% r.m.s., or less, and an 'anti-rolling' tape transport mechanism gives steady tape travel and vibration-resistant operation. The data recorder is not affected by external vibration or dust. Four power sources are available: six dry batteries will provide two hours of recording/playback; an optional rechargeable battery gives four hou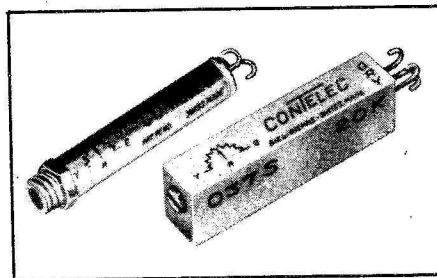rs of continuous operation; an external 11 to 16V d.c. power source can be connected; and a built-in 220V a.c. −10% supply unit used. A 110 and 115V a.c. supply unit is available if required. The input impedance is 100kΩ (f.m.). An optional input filter can be fitted to improve signal-to-noise ratio. The input signal can be monitored from the check terminal and recorded data can be located using the three-digit built-in tape counter. Size is 100 × 340 × 244mm, and weight approximately 6.5kg. Price is £750. Dodwell & Co Ltd, Industrial Import Division, 18 Finsbury Circus, London E.C.2
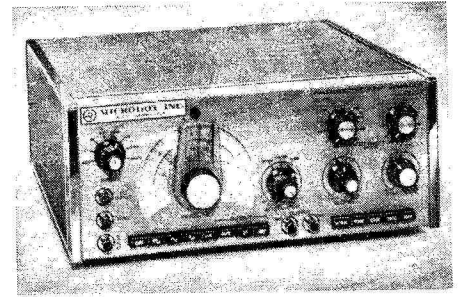**WW310 for further details**

## Wirewound trimmers

Contelec type 025 and 037 wirewound 22-turn trimming potentiometers, are available from Kynmore. Housed in anodized aluminium cases, the units are claimed to be resistant to the effects of humidity and immersion. Type 025 is for panel-mounting, and type 037 is side-mounted. Resistance range is 10Ω to 125kΩ. Both units have a power rating of 1.5W at 85°C. Temperature range is −55 to 170°C. Model 025 is 6.35mm in diameter, and 34mm long. Type 037 measures 6.35 × 9.50 × 31.77mm. Kynmore Engineering Co. Ltd, 19 Buckingham Street, London W.C.2.
**WW320 for further details**

## Waveform generator

Model F220A waveform generator from Microdot Inc.—available in the U.K. from Texscan Instruments—generates sine, square, triangle, ramp and offset sine waveforms over the frequency range 0.005Hz to 3MHz and provides outputs at both 50 and 600 Ω impedances. Triggered, gated or tone-burst outputs can be selected in addition to normal c.w. operation and the generator frequency can be controlled by an external d.c. or wideband a.c. voltage. Output is variable up to a maximum of 32.5V at 600Ω, and Model F220A provides fixed level outputs for each of the waveforms. Accessories available include a power amplifier, signal level monitor and portable power source. Texscan Instruments Ltd, Lord Alexander House, Hemel Hempstead, Herts.
**WW323 for further details**

## Mains input filters

A series of mains input filters from Waycom are primarily designed to offer protection against mains-borne asymmetrical transient voltage spikes. They are suitable for equipment taking up to 4A single phase (3A three phase). The degree of protection offered is such that for a 2kV pulse with rise time of 0.5μs, the transient current flowing will not exceed 20mA, which in typical circuitry means voltage transients of less than 200mV. Waycom Semiconductors Ltd, Wokingham Road, Bracknell, Berks.
**WW 302 for further details**

## Very accurate recording system

A tape recording system manufactured by Gresham Recording Heads is capable of recording and replaying signals in such a way that time intervals are reproduced with an error of less than 0.005%. Noise and distortion are less than 1% in the range 5 Hz-90 kHz. The analogue signal to be recorded is fed into an A-D converter having a sampling rate of $2 \times 10^5$ samples/s. Each sample is then converted into an 8-bit binary word which is fed, in parallel fashion, into 8 channels of a 9 channel digital tape recorder. A timing pulse from a master oscillator having a frequency error less than 1 part in $10^6$ is fed into the 9th channel of the recorder.

The nine channels, each having a data rate of 200 kilobits/s, are then recorded on 0.5in magnetic tape at a speed of 120 i.p.s. A 3200 f.r.p.i. (flux reversals per inch) double-gap recording head has been developed to cope with the high recording accuracy. Read output is 22mV p-p at 150 i.p.s. using an optimum write current of
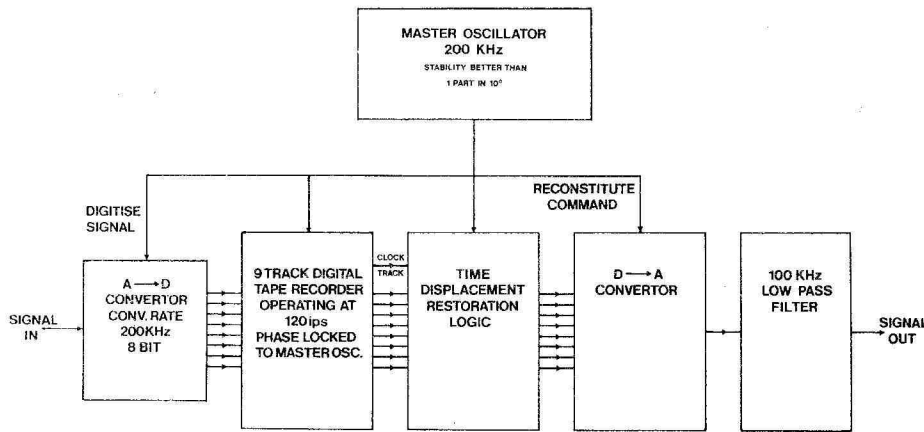
BLOCK DIAGRAM OF SPECIAL PURPOSE RECORDING SYSTEM

$50 \pm$ 10 mA at 1600 f.r.p.i. Crossfeed (write to read) is less than 0.3mV p-p and intertrack crosstalk better than 28dB under worst-case conditions. To reproduce the recording signals, the replayed data is first fed via a time displacement restoration logic unit, and then via a D-A converter to reconstitute the original signal. This converter is followed by a low-pass filter enabling the continual reproduction of the signal within the specified bandwidth to be obtained. The accuracy of the timing of replayed signals is dependent upon the stability of the master oscillator. Gresham Lion Group Ltd, Twickenham Road, Hanworth, Middx.
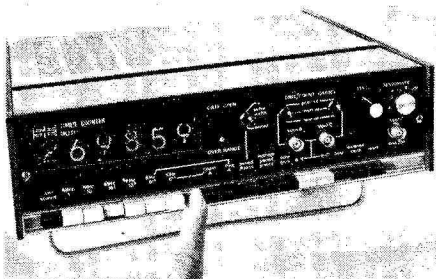**WW325 for further details**

## De-soldering wick

Bradewick de-soldering wick, available from Light Soldering Developments, is impregnated with a flux enabling it to remove molten solder from joints by absorption. It is available in transparent plastic packs in lengths of approximately 1.5m. There are four widths of wick for use with different wattage soldering irons. Price 90p per pack. Light Soldering Developments Ltd, 28 Sydenham Road, Croydon, CR9 2LL.
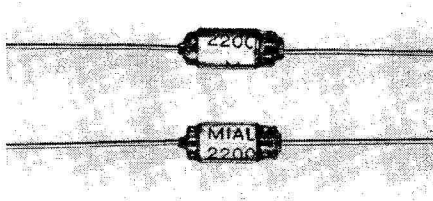**WW329 for further details**

## Versatile counter-timer

The SM 201 universal counter-timer from SE Laboratories measures frequency, period, period average, time interval, count, pulse width and frequency ratio. Single- or double-line gating with positive or

negative transients or contact closure are possible. Display can be stored if required. Full count is 999999, plus over-range indication. Input sensitivity of 10mV and input impedance of $1M\Omega/20pF$ permit the use of high-frequency passive probes. Stability is provided by a crystal-controlled oscillator with a temperature coefficient of less than 1 p.p.m./°C. An external clock can be used. A 3 parts in $10^9$ fast warm-up reference is available as an option. Price £345 approx. SE Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middx.
**WW324 for further details**

## Axial-lead polystyrene capacitors

The Mial 616 range of non-encapsulated polystyrene-dielectric capacitors from Waycom have axial leads of 0.6mm or 0.8mm diameter, depending on capacitor



size. The range of values is 20–100,000 pF in tolerances of $\pm$20, 10, 5 and 2.5% and a voltage of 25–630V. Working temperature is from −40 to 85°C. Waycom Ltd, Wokingham Road, Bracknell, Berks.
**WW321 for further details**

## Quartz crystal filters

Salford Electrical Instruments have introduced a wide range of crystal filters to meet the selectivity requirements of British Post Office specifications for marine and land-based h.f. communications systems. At 100kHz, 1.4MHz and 1.6MHz, a single filter can be supplied which meets both transmitter and receiver requirements. Insertion loss is typically 2dB. At 5.2MHz

four filters are available to meet either transmitter or receiver specifications, including filtering for A3 (a.m.) and A3H (s.s.b. full carrier) modes. Each of the four filters has a volume of 19cc. The filters operate over the temperature range −10 to +55°C within their overall response specification. Salford Electrical Instruments Ltd, Peel Works, Barton Lane, Eccles, Manchester M30 0H1.
**WW327 for further details**

## De-soldering tool

A de-soldering tool, known as the Soldavac, is available from Henri Picard & Frere. Suction in the Soldavac is created by a spring-loaded plunger, contained within the body of the tool. The tool has steadying rests for the fore and middle fingers, and a trigger placed for thumb pressure.
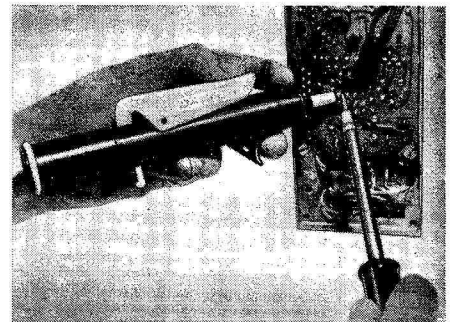


Once fired, it can be re-loaded using the same hand only, either by the action of the thumb or by pressing the plunger tab against the edge of a table. The trigger also acts as a lever for ejecting the nozzle so that the barrel can be emptied and cleaned. Price £1.95. Henri Picard & Frere Ltd, 34/35 Furnival Street, London E.C.4.
**WW319 for further details**

## Double balanced mixers

A range of sub-miniature double-balanced mixers in a low-profile flat-pack configuration is available from Anzac through Wessex Electronics Ltd. Type MD-123 provides conversion loss of 8dB maximum over the range 10 to 3000MHz. Inputs to any two ports will produce the sum and difference frequencies at the third port. The device may be used with local oscillator inputs ranging from 7 to 20dBm. Precision balanced circuits provide
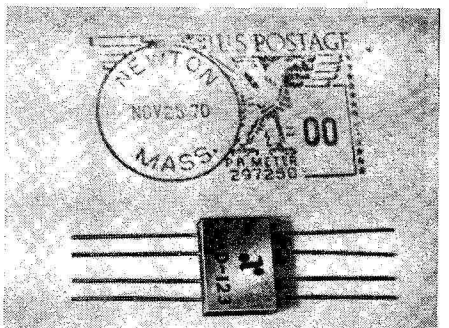
two-tone third-order i.m. ratios of better than 100dB with −30dBm input tones. The full range comprises MD-123 (10-3000MHz), MD-113 (10-1000MHz), MD-125 (0.5-500MHz), and MD-124 (50Hz-200MHz). Wessex Electronics Ltd., Stover Trading Estate, Yate, Bristol BS17 5QP.
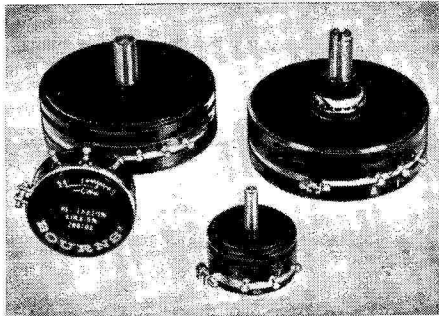**WW301 for further details**

## Linear-law potentiometers

A range of single-turn precision potentiometers—the B-Line from Bourns—employs low temperature coefficient long-life elements. The range is available in diameters of $\frac{7}{8}$in, $1\frac{1}{16}$in and 2in, bushing or servo mount.

**Specification:**

| | |
|---|---|
| resistance value | 20Ω to 100kΩ |
| tolerance | ±10% |
| power rating at 70°C | 1.0W for $\frac{7}{8}$in |
| | 1.25W for $1\frac{1}{16}$in |
| | 2.0W for 2in |
| Max. operating temperature | 125°C |
| output smoothness | 0.1% of v.r. |
| linearity | from ±0.5% to ±0.1% |
| insulation resistance | 1000 MΩ |
| vibration tolerance | 15G |
| shock tolerance | 50G |



Bourns (Trimpot) Ltd, Hodford House, 17/27 High Street, Hounslow, Middx.
**WW318 for further details**

## High-permittivity ceramic

GEC Hirst Research Centre has developed a new class of high permittivity ceramic dielectric for use as a microcircuit substrate at the lower microwave frequencies. The material is a zirconate ceramic of permittivity −35 compared with conventional alumina's −10. Thus smaller size microstrip circuits can be used at u.h.f. or low microwave frequencies. The photograph shows equivalent ring resonators for use at 5GHz deposited on (left) a zirconate ceramic substrate and (right) standard alumina substrate, showing a 2:1 linear size advantage for the zirconate. The new material possesses a very low dielectric loss ($Q$−2000) and a low controllable temperature coefficient of permittivity, so that temperature-stable resonators can be made in a relatively small size. The ceramic may be used to load phase shifters (e.g. for phased array
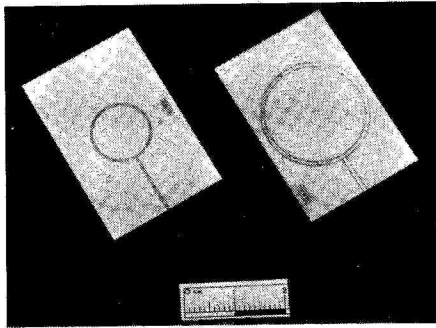


aerials) to give higher performance. GEC Hirst Research Centre, East Lane, Wembley, Middlesex.
**WW313 for further details**

## Low distortion oscillator

Model CR116 oscillator in the NF Instruments Co. range of test instruments, available in the U.K. from Tekmar Electronics, covers 5Hz—540kHz in five ranges. Frequency response is flat ±0.2dB from 20Hz to 50kHz and distortion down to 0.015% between 200Hz and 10kHz. Output level is +16dB maximum (open circuit), +10dB when feeding a 600Ω balanced load. Operation is from the mains and the price is £266.62. A portable version, the CR117CT which employs a NiCd battery, costs £201.96. Tekmar Electronics Ltd, 102 High Street, Harrow-on-the-Hill, Middx.
**WW311 for further details.**

## Wide-range signal generator

Combining the techniques of the frequency sweeper and an a.m./f.m. signal generator the TF2008 from Marconi Instruments covers the range 10kHz to 510MHz. This range is provided in eleven switch-selected bands and the instrument incorporates two primary signal sources—a manually-controlled oscillator and a

voltage-controlled oscillator. When the latter is in use it can be coupled to an internal sweep-drive generator which gives continuous sweep over the whole, or any part, of each tuning band. Narrow-band sweep is possible when the instrument is used as a manually-tuned signal generator. Price £1700. Marconi Instruments Ltd, St. Albans, Herts.
**WW317 for further details**

## Digital indicators

Newtron Indicator Tubes from FR Electronics are 7-segment indicators incorporating directly viewed incandescent filaments allowing viewing angles up to 140°. The units have a normal operating
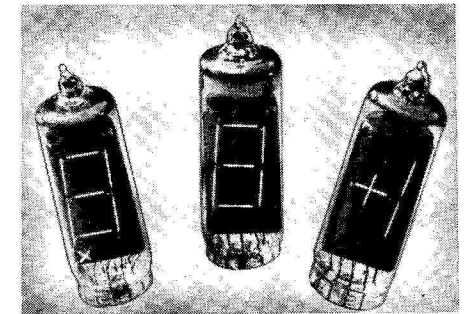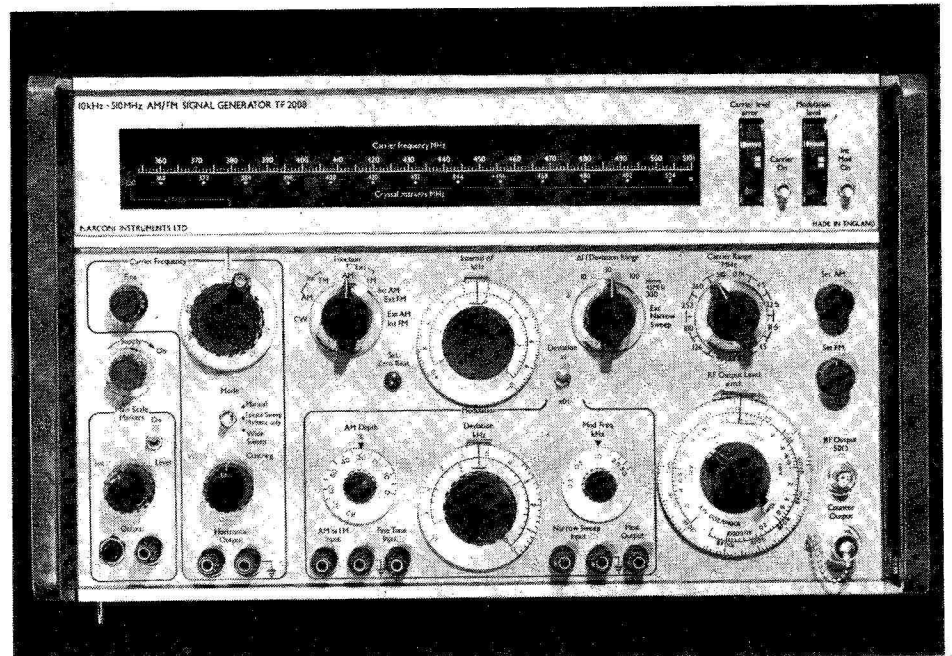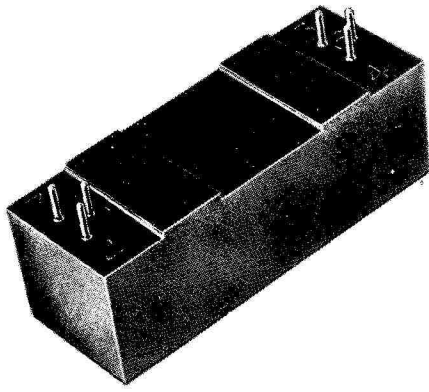


voltage of 5V with a segment current drain of 20mA, and are i.c. compatible. The brightness of the display can be varied to suit all ambient light conditions, permitting viewing even in direct sunlight. FR Electronics, Wimborne, Dorset BH21 2BJ.
**WW312 for further details**

## Sensitive reed relays

Pye TMC has introduced a range of sensitive reed relays with a variety of switching modes encapsulated in a tough, stable, moisture-resistant epoxy resin. The range is particularly designed for high-

speed switching and alarm type contacts. A built-in magnetic shield prevents inter-action between closely stacked relays. The connections are for printed circuit 0.100 matrix and are polarized to prevent wrong assembly. The relay is available in three variants—latching, normally-closed or normally-open. Pye TMC Ltd, Components Division, Roper Road, Canterbury, Kent.
**WW322 for further details**

## Optical position sensor diode

New to the United Detector Technology range of position-sensing Schottky barrier photodiodes is type SC-50 with an active area 1.40 inches square. In common with the other diodes in this series the SC-50 senses the position of a light spot in two dimensions and gives position sensing information independent of the spot size. Resolution and null sensitivity are independent of incident power changes and the null point may be shifted electrically. Position sensitivity is $0.4\mu A/mW/0.001$in at the spectral peak, and non-linearity at 0.05in from the centre is $\pm 1\%$. Techmation Ltd, 58 Edgware Way, Edgware, Middlesex HA8 8JP.
**WW304 for further details**

## Semiconductor tester

A semiconductor tester from Levell, type TB12, measures the characteristics of bipolar transistors, diodes and zener diodes. Leakage currents down to 0.5nA can be determined from 2V to 150V, current gain of transistors checked at collector currents from $1\mu A$ to $100\mu A$, and breakdown voltages up to 100V
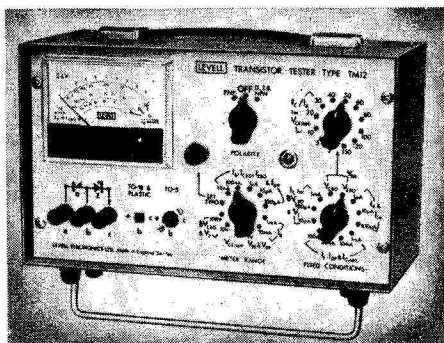


measured at currents of $10\mu A$, $100\mu A$ and $1\mu A$. The collector-to-emitter saturation voltage of a transistor is measured at collector currents of 1, 10, 30 and 100mA for $I_C/I_B$ ratios of 10, 20 and 30. The instrument is powered by a 9V battery and contains a transistor d.c. to d.c. converter to produce 150V. The state of the battery is indicated by a neon panel lamp. Price £65. Levell Electronics Ltd, Park Road, High Barnet, Herts.
**WW308 for further details**

## Discoidal lead-through capacitors

A range of discoidal lead-through capacitors, type DLT/10,000, from Oxley, employ multi-layer construction using a high 'K' ferro-electric ceramic for high capacitance per unit volume. The discoidal construction, which permits a radial current flow in the capacitor electrodes, is
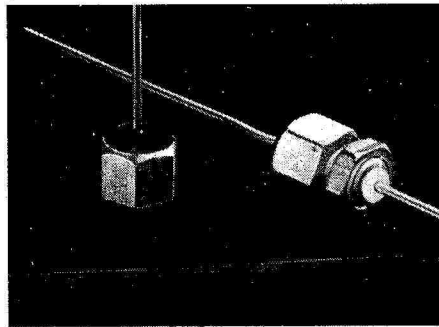


said to result in self inductance considerably smaller than that inherent in a capacitor having a more conventional construction. The component is mounted in a 2BA clearance hole. The body is a 4BA hexagon section, with a gold finish, the lead-through wire being 20 s.w.g. tinned copper.

**Characteristics:**

| | |
|---|---|
| test voltage | 250V d.c. |
| working voltage | 100V d.c. |
| operating temperature range | $-55$ to $+125°C$ |
| capacitance | 10,000pF $\pm 20\%$ or $+80\% -10\%$ |

Oxley Developments Co. Ltd, Priory Park, Ulverston, North Lancs.
**WW314 for further details**

## Temperature-controlled soldering iron

The Oryx 50, from W. Greenwood Electronic, gives simple adjustment for any temperature between 200 and 400°C, without changing the tip. Heat settings are accurate to $\pm 2\%$. Tip temperature variations during soldering are negligible and temperature changes can be made in seconds whilst the iron is on. Oryx 50 operating temperatures can be much lower than with conventional uncontrolled irons. An indicator lamp, controlled by the
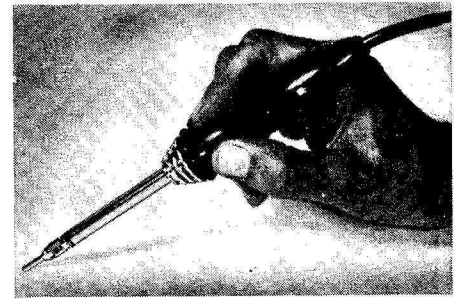


thermostat, is built into the handle. The instrument is fitted with a long iron-coated tip as standard. There is a range of eleven tips in all—long-life or copper/nickel plated to choice. The iron weighs 77 g, is rated at 50W, and heats up in 45 seconds. Operating voltages are 12, 24, 50, 115 or 210/250V a.c. Price with long tip is £3.75. Stand £1.25. W. Greenwood Electronic Ltd, 21 Germain Street, Chesham, Bucks.
**WW306 for further details**

## Digital multimeter

Model 460 self-contained digital multi-meter, from Bach-Simpson, provides 26 ranges including alternating current. A battery pack is built in along with a charger unit which operates automatically when the instrument is mains operated. Polarity and over-range indication are automatic. Ranges (which are measured without the use of external shunts) are as follows:

| | |
|---|---|
| volts a.c./d.c. | $100\mu V–1000V$ |
| amps a.c./d.c. | $100nA–2A$ |
| resistance | $190m\Omega–20M\Omega$ |

The system is protected against overload.
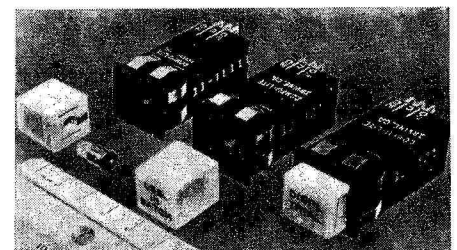


It weighs 3kg (with batteries) and measures $11 \times 24 \times 20$cm (approx.). Price £150. Bach-Simpson Ltd, 331 Uxbridge Road, Rickmansworth, Herts, WD3 2DS.
**WW315 for further details**

## Push-button switches

A range of illuminated multi-pole Compu-Lite Series 11 push-button switches from Guest International Ltd. are designed for front panel fixing. They are
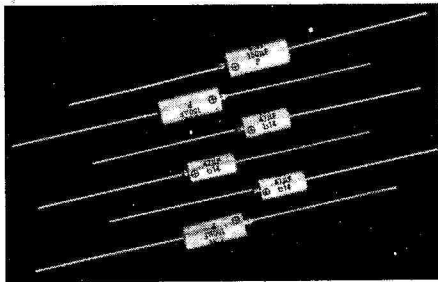
enclosed and sealed and switch up to 5A at 250V. Each switch allows one pole to be switched in before the remaining poles make contact. Gold contacts are available for low-level switching.

A wide range of coloured bezels and screen split or full legends can be supplied, and a number of different switching actions is also available. Maximum depth is only 38mm. Series 11 switches can be made available with AMP-type terminals. Guest International Ltd, Nicholas House, Brigstock Road, Thornton Heath, Surrey. **WW307 for further details**

## Solid tantalum capacitors

A life of 1000 hours operation in the temperature range $-55$ to $125°C$ is guaranteed for a range of metal-cased solid tantalum electrolytic capacitors available from Seatronics (UK). Capaci-



tance tolerance is $\pm$ 20% in the range of 0.35 to 330$\mu$F, at voltages from 6.3 to 50V d.c. Leakage current is 0.002$\mu$A/$\mu$FV max., and tan delta is 0.06 max. The 1000-off price ranges from 17p to 72p each, the latter being the cost of a 50V 22$\mu$F unit. Seatronics (UK) Ltd, 22-25 Finsbury Square, London EC2A 1DT. **WW326 for further details**

## Transistor amplifiers for 3.4 to 4.2GHz

A series of solid-state amplifiers for the 3.4 to 4.2GHz range is announced by Watkins-Johnson Company. The WJ-5102 amplifiers provide a 7dB noise figure, $\pm$ 0.3dB gain flatness, $+$ 25dBm intercept point and 1.2:1 v.s.w.r. Time-delay distortion is small: linear component is $1 \times 10^{-3}$ ns/MHz, parabolic component, $1 \times 10^{-6}$ns/MHz$^2$; residual ripple, 0.2ns peak-to-peak. The design is a microstrip employing chip components. There is a choice of gains from 10dB to 50dB and power output as great as $+$20dBm at the 1dB compression point. These amplifiers are available with or without integrated power supplies. Watkins-Johnson International, Shirley Avenue, Windsor, Berkshire. **WW332 for further details**

## Polyester foil capacitors

Available in a capacitance range of 1000pF to 1$\mu$F, new ISKRA KMFU high-quality polyester foil capacitors from Guest International are non-inductive
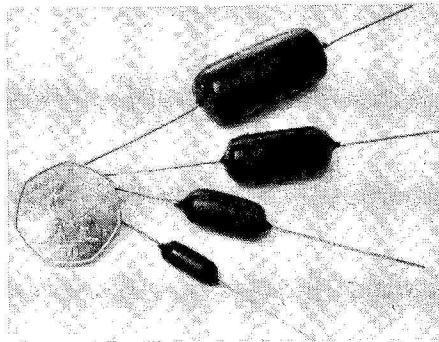


($<$20nH). They can be used at high frequencies and with pulse waveforms. Insulation resistance is very high ($>$30,000M$\Omega$ at 20°C, 100V d.c.) allowing safe use at high temperatures. Ex-stock voltage range is 125, 250, 400, 630 and 1000V d.c. A 1500V d.c. capacitor is available to order. Guest International Ltd, Nicholas House, Brigstock Road, Thornton Heath, Surrey. **WW330 for further details**

## Miniature d.c. motors

The Escap 20 series ironless rotor d.c. motors from Portescap employ self-supporting skew windings, to provide low inertias and short time constants. The motors incorporate gold alloy brushes, precious-metal commutators and self-lubricating sintered bronze bearings. Built-in reduction gearheads with ratios 1:4, 1:15 and 1:59 can be supplied with the motors which offer output powers from
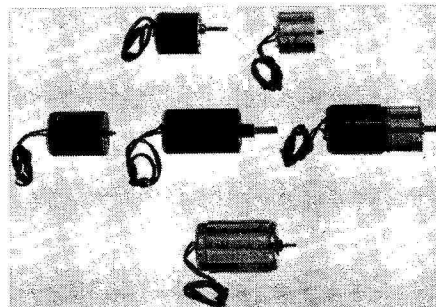


0.15 to 3.1W, starting torques from 6 to 132gcm, and no-load speeds up to 17,300 r.p.m. They measure between 20 and 33mm long by 20mm in diameter, and weigh only 20-65g. Portescap (U.K.) Ltd, 204 Elgar Road, Reading RG2 0DD. **WW 328 for further details**

## Thick-film amplifier/ oscillator modules

Redac have announced a modular select-to-order fixed frequency oscillator and a compatible frequency-selective amplifier. Both modules employ thick-film circuit techniques and require a 12V supply. Oscillator module, type TF002, offers a fixed frequency of operation in the range 100Hz to 1MHz with a tolerance of 1% maintained over 0 to 45°C. Two outputs are provided-1V at 300$\Omega$ and 10mV at 1k$\Omega$. Frequency selective type

TF003 gives a voltage gain of 1000 over 100Hz to 1MHz with input and output impedances of less than 1k$\Omega$ and greater than 100$\Omega$ respectively. Gain bandwidth is 15%. Size 36 $\times$ 36 $\times$ 10mm. Redac Software Ltd, Newtown, Tewkesbury, Gloucester, GL20 8HE. **WW303 for further details.**

## Low thermal e.m.f. reed switch

The MRA-230 reed switch from FR Electronics has a thermal junction e.m.f. of 10$\mu$V/°C. It is of form A construction (contacts normally open) and its miniature
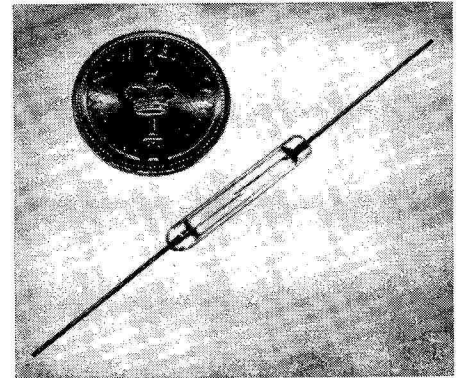


size can be gauged from the photograph. The switch has been designed for use at r.f. up to 30MHz. F R Electronics, Wimborne, Dorset BH21 2BJ. **WW316 for further details**

## Incremental indicator

The Comark incremental indicator type 1211-30 is a battery powered portable instrument which will measure direct voltages from 0 to 30mV with a resolution
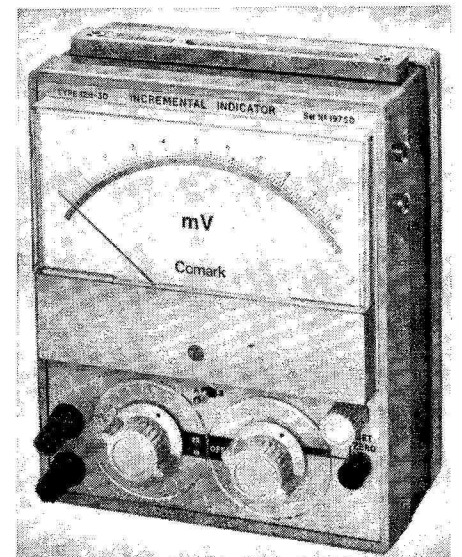


better than 10mV. The instrument has an accurate backing-off source built in, which is used to provide 30 ranges up to 1mV f.s.d. Comark Electronics Ltd, Brookside Avenue, Rustington, Littlehampton, Sussex. **WW305 for further details**

# Literature Received

*For further information on any item include the appropriate WW number on the reader reply card*

## ACTIVE DEVICES

A 22-page product summary brochure covers all the integrated circuits in production at Plessey's Swindon factory. Plessey Microelectronics, Cheney Manor, Swindon, Wilts. ...................... WW401

MCP Electronics Ltd, Alperton, Wembley, Middlesex, HA0 4PE, are distributors of Telefunken semiconductors and have available a short-form catalogue dealing of their diodes, transistors and integrated circuits ..................... WW402

The following data sheets from RCA, Sunbury-on-Thames, Middlesex, describe new high-frequency transistors and microcircuits:

2N6093. 75W (p.e.p.), 2-30MHz, 13dB gain .............................. WW403
TA7993. 2W, 2GHz, 8.2dB gain (3W, 1GHz, 12dB) ............................ WW404
TA7486. 200mW, 1GHz, 12dB gain ... WW405
TA7994. 5W, 2GHz, 7dB gain (13.5W, 1GHz, 11dB) ........................... WW406
TA7702/3. Microwave broadband i.c., 225-400MHz, 16W, 6dB gain at $V_{cc} = 28$V WW407
TA7747. Integrated circuit power combiner/divider 225-400MHz, powers up to 40W, $Z_{in} = Z_{out} = 50\Omega$

AEI Semiconductors, Carholme Rd, Lincoln, have available a revised price list covering their microwave semiconductors, regulator diodes, rectifier diodes, thyristors and triacs, and thyristor/rectifier assemblies ........................... WW409

## PASSIVE COMPONENTS

Toggle, rocker, wafer and slide switches are described in a catalogue from Lorlin Electronic Co. Ltd, Billingshurst, Sussex .............. WW410

Miniature d.c. motors (Escap 26 series) with very low rotor inertia for instrumentation are described in a leaflet from Portescap U.K. Ltd, 204 Elgar Rd, Reading RG2 0DD. Reduction gear-heads with ratios from 1.4:1 to 560:1 are also described WW411

Four 13A sockets with a switch and neon indicator in a box form a mains distribution panel that has been added to the Lecktrokit range. A leaflet describes the panel and explains the Lecktrokit system of construction; A.P.T. Electronic Industries Ltd, Chertsey Rd, Byfleet, Surrey ............ WW412

We have received the following literature from Erie Electronics Ltd, South Denes, Gt. Yarmouth, Norfolk:
Stock catalogue listing a wide range of capacitors, resistors, potentiometers and semiconductors ................................ WW413

Data sheet R/20. Describes a range of $\frac{1}{8}$ and $\frac{1}{4}$W resistors 2.2$\Omega$ to 5.1M$\Omega$ ±5% ..... WW414

Data sheet R/21. Carbon composition resistors, $\frac{1}{4}$ and $\frac{1}{4}$W, 10$\Omega$ to 22M$\Omega$ in ±5, 10 and 20% tolerance ....................... WW415

Saft (U.K.) Ltd, Castle Works, Station Rd, Hampton, Middlesex, have sent us the following data sheets on cadmium nickel batteries and associated equipment:
VR series. Cylindrical 0.5 to 10Ah .... WW416
VB series. 'Button packs' available in sintered plate or plastic sleeve construction, 2.4 to 12V, 90 to 1,750mAH ................. WW417

S1000T. Constant current battery charger for up to 20 cells in series, incorporating a timer; charging rate adjustable from 10mA to 1A ................................. WW418

A range of relays, called the 'GPR300 series', manufactured by Pye/TMC, Components Division, Roper Rd, Canterbury, Kent, are the subject of a leaflet. Various coil and contact combinations are possible ......................... WW419

The plugs and sockets distributed by F. C. Lane Electronics Ltd, Slinfold Lodge, Horsham, Sussex, are described in a short-form catalogue ... WW420

## APPLICATION NOTES

We have received three application notes from Waycom Semiconductors Ltd, Wokingham Rd, Bracknell, Berks, RG12 1ND:
1: 'Pulse transformers for thyristor firing circuits' deals with the theory, makes some recommendations and highlights some pitfalls .... WW421
2: 'Harmonics generated by thyristor controlled circuitry—Part 1'. The nature of the problem is discussed and some general $LC$ suppression methods are given .................. WW422
3: 'Harmonics generated by thyristor controlled circuitry—Part 2'. Deals mostly with the suppression of interference from shunt wound motors from 150kHz to 30MHz ..... WW423

We have received the literature listed below from RCA, Sunbury-on-Thames, Middlesex:
'An h.f. power transistor for linear applications', discusses the 2N6093 and concludes with a 150W, wideband (2-30MHz) linear amplifier design ........................... WW424
ST4700. 'Integrated circuit stereo decoder does everything', describes in detail the phase-locked-loop decoder which was mentioned last month (p. 377) ................ WW425
CA3088E. 'A.M. receiver sub-system and general-purpose array' gives data and circuitry for a new a.m. receiver i.c. .............. WW426
ST4698. 'Advances in f.m. receiver design', describes a new f.m. receiver i.c. which incorporates i.f. stages, detector, a.f.c. output, tuning meter output, a.g.c. output, decoder disable line and facilities for a squelch control ......................... WW427
CA3089E. Data sheet for above ...... WW428
'Recent advances in the design of micropower operational amplifiers', deals with operational amplifiers that have no internal resistors and gives some uses for them ......... WW429
ST3857. 'Microwave power generation using r.f. power transistors' describes the construction of overlay and interdigitated transistors before giving application information ....... WW430
'Power circuits—d.c. to microwave', 448-page book of circuits and explanations, price £1.30

'Gunn diode circuit handbook' is a useful 40-page booklet published by Microwave Associates Ltd, Cradock Rd, Luton, Beds LU4 0JQ ..... WW431

## EQUIPMENT

Addition, subtraction, multiplication, division, 'chain multiplications', calculations with a stored constant, raising to a power and mixed calculations

may all be done with a pocket printing calculator from Computer Ancillaries Ltd, Radio House, Central Trading Estate, Staines, Middlesex. Results are printed on a cassette of paper type—price is £179. A leaflet gives a full description. ..... WW432

Temperature measurement can be made remotely so that environmental conditions are not disturbed using the KT12 infra-red radiation thermometer which is described in a booklet from the Scientific Instruments Division of Guest International Ltd, Nicholas House, Brigstock Rd, Thornton Heath, Surrey ............................. WW433

A new computer, Satellite One, is described in a brochure from Computer Technology Ltd, Eaton Rd, Hemel Hempstead, Herts. ........... WW434

'The complete guide to your digital instrument requirements' is the rather misleading title of a booklet from SE Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex. One would expect (from the title) a complete survey of the whole field of digital instruments when in fact only SE Labs' equipment is mentioned. However, a useful section on using digital instrumentationis included .............. WW435

Details of a comprehensive range of microwave components are given in a 125-page catalogue prepared by Microwave and Electronic Systems Ltd, Lochend Industrial Estate, Newbridge, Midlothian, Scotland ...................... WW436

J Beam Engineering Ltd, Rothersthorpe Cres., Northampton, have published leaflets in English, French and German describing a 450-470MHz glass fibre colinear aerial (type 7058) with a 10dB gain over a half-wave dipole:
English ........................... WW437
French ........................... WW438
German ........................... WW439

An analogue signal converter (type PSC 300), intended for use as an interface unit in instrumentation systems when the process signal and the instrument signal are incompatible, is described in a brochure from Mimic Diagrams and Electronics Ltd, Maxim Rd, Crayford, Kent ......... WW440

An optical colour comparator for setting-up colour television receivers, a colour film assessor and a light-meter calibrated in foot-Lamberts are described in a brochure from Grafikon (Engineers) Ltd, 75 South Western Rd, Twickenham, Middlesex WW441

## PROSPECTUSES (1971-'72)

Department of Telecommunication and Electronics, Norwood Technical College, Knight's Hill, London S.E.27.

Hendon College of Technology, The Burroughs, Hendon NW4 4BT.

Faculty of Engineering, Brighton College of Technology, Pelham St, Brighton BN1 4FA.

Department of Engineering, Twickenham College of Technology, Egerton Rd, Twickenham, Middlesex.

Compendium of Degree Courses 1971, Council for National Academic Awards, 3 Devonshire St, London W1N 2BA.

## GENERAL INFORMATION

Our associated book company, The Butterworth Group, 86 Kingsway, London WC2B 6AB, has produced the following catalogues:
Books on radio ..................... WW442
Selected books on radio and television . WW443
Books on television ................. WW444
Postgraduate books on electronics ..... WW445

A catalogue of electrical and electronic hobbyist books may be obtained from Tab Books, Blue Ridge Summit, Pennsylvania 17214, U.S.A.

An RCA publication 'Beam-lead devices' graphically explains how this type of chip is constructed. RCA, Sunbury-on-Thames, Middlesex .......... WW446

Details of the 3M wildlife recording contest are given in a leaflet available from 3M House, Wigmore St, London W1A 1ET. The first prize is a natural history holiday worth £150.