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## R.C. OSCILLATORS

## 50/70 WATT ALL SILICON AMPLIFIER WITH BUILT-IN 4-WAY MIXER USING F.E.T.s.

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The mixer is arranged for $2-30 / 60 \Omega$ balanced line microphones, $1-\mathrm{HiZ}$ gram input and 1 -auxiliary input followed by bass and treble controls. 100 volt balanced line output or $5 / 15 \Omega$ and 100 volt line.


#### Abstract

50/70 WATT ALL SILICON AMPLIFIER WITH BUILT-IN 5-WAY MIXER USING F.E.T.s This is similar to the 4 -way version but with 5 inputs and bass cut controls on each of the three low impedance balanced line microphone stages, and a high impedance ( 10 meg ) gram stage with bass and treble controls plus the usual line or tape input. All the input stages are protected against overload by back to back low noise, low intermodulation distortion and freedom from radio breakthrough. A voltage stabilised supply is used for the pre-amplifiers making it independent of mains supply fluctuations and another stabilised supply for the driver stages is arranged to cut off when the output is overloaded or over temperature. The output is $75 \%$ efficient and 100 V balanced line or $8-16 \Omega$ output are selected by means of a rear panel switch which has a locking plate indicating the output impedance selected.


100 WATT ALL SILICON AMPLIFIER. A high quality amplifier with 8 ohms 15 ohms or 100 volt line output for A.C. Mains. Protection is given for short and open circuit output over driving and over temperature. Input 0.4 V on 100 K ohms.

THE 100 WATT MIXER AMPLIFIER with specification as above is here combined with a 4 channel F.E.T. mixer, $2-30 / 60 \Omega$ balanced microphone inputs, $1-\mathrm{HiZ}$ gram input and 1 -auxiliary input with tone controls and mounted in a standard robust stove enamelled steel case. A stabilised voltage supply feeds the tone controls and pre amps, compensating for a mains voltage drop of over $25 \%$ and the output transistor biasing compensates for a wide range of voltage and temperature. Also available in rack panel form.

CP50 AMPLIFIER. An all silicon transistor 50 watt amplifier for mains and 12 volt battery operation, charging its own battery and automatically going to battery if mains fail. Protected inputs, and overload and short circuit protected outputs for 8 ohms- 15 ohms and 100 volt line. Bass and treble controls fitted.
Models available with 1 gram and 2 low mic. inputs, 1 gram and 3 low mic. inputs or 4 low mic. inputs.

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20/30 WATT MIXER AMPLIFIER. High fidelity all silicon model with F.E.T. input stages to reduce intermodulation distortion to a fraction of normal transistor input circuits. The response is level 20 to $20,000 \mathrm{cps}$ within 2 dB and over 30 times damping factor. At 20 watts output there is less than $0.2 \%$ intermodulation even over the microphone stage at full gain with the treble and bass controls set level. Standard model 1-low mic. balanced and 1 auxiliary input.


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Precision 6V Regulator 0.C. Supply
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# Wireless World 

Electronics, Television, Radio, Audio

August 1971

Volume 77 Number 1430



The rotating $11-\mathrm{dB}$ helical aerial, shown on the front cover, forms a major part of receiving equipment for picture transmission from weather satellites. It was designed and manufactured by Rohde \& Schwarz, of Munich, whose U.K. agents are Aveley Electric.

## IN OUR NEXT ISSUE

Constructional details of a helical aerial of unusual design covering the $88-170 \mathrm{MHz}$ band.
Swept-frequency audio oscillator in which two decades are covered in one band using a beat-frequency technique.

## ibpa



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## Contents

361 Wasted R\&D?<br>362 Double-trace Oscilloscope Unit-1 by W. T. Cocking<br>365 Announcements<br>366 Ten Practical Source-follower Circuits by J. O. M. Jenkins<br>368 Simple Crosshatch and Dot Generator by A. W. Critchley<br>Square-root Circuit by B. L. Hart \& A. Cheetham<br>News of the Month<br>Letters to the Editor<br>Phase-locked-loop Stereo Decoder I.C.<br>Ceramic Pickup Equalization-2 by B. J. C. Burrows<br>The Diagnosis of Logical Faults (concluded) by R. G. Bennetts<br>Circuit Ideas<br>Electro-optical Gearbox by Jack Dinsdale<br>H.F. Predictions<br>Touch-switch Controller by R. Kreuzer<br>Electronic Building Bricks-14 by James Franklin<br>Charging by 'Cathode Ray'<br>Single-sideband Experimental Broadcasts<br>Telephone Exchanges of the Future<br>Elements of Linear Microcircuits-10 by T. D. Towers<br>Conferences and Exhibitions<br>Complementary Darlington Output Transistors in Audio Amplifiers Automatic Titration Potentiometer by D. R. Bowman<br>Sixty Years Ago<br>World of Amateur Radio<br>Personalities<br>Literature Received<br>New Products<br>Real \& Imaginary by 'Vector'<br>APPOINTMENTS VACANT<br>INDEX TO ADVERTISERS

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# Wireless World 

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## Wasted R \& D

One aspect of research and development costs which we did not consider in our discussion on value for money in $R$ \& $D$ in the June issue is highlighted in a report just issued by the Centre for the Study of Industrial Innovation*. It is called 'On the Shelf and surveys industrial R \& D abandoned for non-technical reasons. The main conclusion of the survey, which analyses 53 shelved projects belonging to 20 companies in the UK, is that the failure to make an adequate market assessment before research and development is the most common reason for firms having to abandon technically viable projects. Eight of the companies or divisions (mostly unnamed) are classified as 'electrical and electronics' and, in fact, a third of the 53 abandoned projects were electronic. Incidentally, the nationality of five of the eight companies is given as U.S., one European and two U.K. Together they employ about 1,500 graduate R \& D staff and about the same number of technicians and support staff. Details are given of case studies (some under disguised names) and together they raise important questions central to the management of $R$ \& $D$. With hindsight, some projects should never have been started; in others development resources were exhausted beyond the point of economic justification.

There are apparently three main points of project rejection, which can occur for technical or non-technical reasons. Projects can fail first to measure up to initial selection criteria for development expenditure; secondly, to measure up to ciiteria of satisfactory development at a periodic progress review; and thirdly, a fully developed prototype can fail to measure up to the conditions required for marketing. Probably most of the projects described in the report were rightly shelved, although we question the attitude of one research director interviewed who stated that the onus is on the individual at the bench to force his own project through; 'He must know how to sell, what to sell, when to sell and who to sell to'. This attitude which, according to the survey is not untypical of research directors and other senior personnel interviewed, lays a tremendous burden on the initiators of a project and may well inhibit progressive thinking.

One of the aims of the study was to assess what steps firms took after the 'shelving' decision to gain some commercial return from the accummulated, but abandoned, know-how. It is this aspect of shelving which we consider is of paramount importance, for it can turn to good effect what would otherwise be wasted R \& D. Of the 53 shelved projects reviewed only six were subsequently economically exploited by three of the 20 companies. Sometimes the resurrection occurred as a result of changed circumstances but only one firm, incidentally Rolls-Royce, had a regular system of project reappraisal. While this aspect of research should not be overlooked by individual companies, it is of paramount importance to organizations undertaking research on behalf of others. In this regard we were particularly interested to learn during a recent visit to the Cranfield Unit for Precision Engineering (see 'Electro-optical Gearbox' in this issue) that written into all its research contracts is a clause permitting the use in any field not competing with the originating company's activities of know-how resulting from research projects.

While it can be said that all $R$ \& $D$ efforts contained an element of fruitless endeavour, there must come a point when this proportion is no longer acceptable as inevitable and it is seen that avoidable wastage has begun to occur.
*The Centre for the Study of Industrial Innovation, of 162 Regent St, London WIR 6DD, was set up last year, with industrial backing, to study the economics of innovation and $R \& D$ in industry. 'On the Shelf' is its first report.

# Dual-trace Oscilloscope Unit-1 

by W. T. Cocking*, f.I.E.E.

In this series of articles the development of a unit which enables two signals to be observed simultaneously on almost any cath-ode-ray oscilloscope will be described and the series will conclude with full details of the final design. In all design work there is compromise and it is necessary to obtain a good balance between conflicting requirements. Sometimes there are several different ways of obtaining a required performance and a designer naturally starts by considering the one which he thinks most likely to be satisfactory. Sometimes, his first choice is a good one; at others, he ends up with something entirely different.
Usually, he says little or nothing about his unsuccessful attempts and only his successful design is presented for all to see. It occurred to the writer that a detailed account of the development, including the unsuccessful arrangements, would be of general interest and might be of some educational value. It is usually true that one learns more from one's mistakes than from one's successes. It might be true that one could learn more from other people's mistakes than from their successes, if one knew about them!

Inf the course of the development, a great variety of problems was met and some were a little unusual. For example, a continuous control of gain was considered desirable and provided by far the most difficult of all problems. In fact, the final choice of circuit was made almost entirely to suit the requirements of gain control.

## Requirements

The first step in design is always to formulate the requirements clearly. The designer does this in the light of his experience of what is practicable. He knows, for instance, that it will probably be difficult to obtain a voltage amplification of 100 times with a bandwidth of 25 MHz from two transistors, He knows, too, that it will probably be easy to obtain a gain of 4 times with a bandwidth of 10 MHz from only one transistor and that it might not be too hard to get a gain of 10 times. The designer has this sort of information available from his past experience but there are always gaps in his knowledge, and then he has to carry out some experiments to see what can be done,

[^2]or else a theoretical analysis. This usually takes longer, but is generally more valuable.
Coming now to the particular (that is, to the dual-trace unit), the first thing is to decide what it must do. Its purpose is to enable two different signals to be displayed so that they can be viewed simultaneously on the screen of the c.r.o. They cannot, of course, be actually present simultaneously, for the tube has only one electron beam. There must be two separate signal channels and an electronic switch to switch the input of the c.r.o. from one to the other and back again repeatedly at an adequately high speed. Persistence of vision coupled with the persistence of the c.r.t. screen makes the traces appear to be present simultaneously.
Both traces are, of course, displayed by the same horizontal deflection of the beam, and so the two signals must be of the same frequency or harmonically related. Also, if the two traces are separated to appear one above the other, the maximum input to the oscilloscope for each signal can only be one half of the normal. The screen cannot be stretched to accommodate two normal size traces!

## Experiment

It is not necessary for the switching frequency to be synchronized to either the signals or to the oscilloscope timebase. Here, for brevity, we are anticipating a little. In reality, at this stage we did not know what would happen, so we rigged up an electronic switch and fed the same sinewave signal to both sides to find out what did happen. This is what we did find. For'signals from about 200 Hz to perhaps 1 MHz the best results are obtained and the operation is easiest when the electronic switch is triggered by the oscilloscope timebase. No spurious effects are then observable, the two signals are displayed alternately on successive sweeps and the switching occurs during flyback. It was found, however, that for the display of higher frequencies, the timebase frequency became too high for the electronic switch to operate properly. It was found, too, that at lower frequencies flicker quickly became intolerable. The cure for both is the same, to use an unsynchronized switch. At low frequencies, the switching frequency is made much higher than the signal frequency. Switching occurs 100
times or more during each signal cycle. If, by accident, the frequencies are integrally related, or there is some unintentional synchronizing action by stray coupling, the traces appear dotted. Flicker is not now any worse than in a normal oscilloscope display. At high frequencies, the switching frequency is made much lower than the signal frequency. One signal is then displayed for ten or more sweeps before the other is switched on, but as long as the switching frequency is above a few hundred Hz one does not notice this.

Unsynchronized operation can be used for all signal frequencies, but peculiar effects occur at certain relations between the signal and switching frequencies. They are in the nature of stroboscopic effects and can be most disturbing. To minimize them the ratio of the frequencies must be very large or small and a fine control of switching frequency is necessary.
In the light of these early experiments it was decided that synchronized operation would be used for most signals, but that an alternative pulse generator would be provided for low- and high-frequency signals. It should be noted also that synchronized operation demands that the oscilloscope has a pulse or sawtooth output available from its timebase.
It was noticed, too, in the experiments that it is impossible to use the internal synchronization of the oscilloscope. With unsynchronized operation of the switch, the timebase invariably locks to the switch frequency and not to the signal.

On its most sensitive range the average oscilloscope needs no more than 1 V peak-to-peak of signal for full screen deflection. Many oscilloscopes need less. It was decided that the dual-trace unit should have an overall gain of unity, with a maximum signal output of 1 V . The oscilloscope used in the development was the Marconi Instruments TF 1330. This is now an oldish model but its performance is quite adequate for most general purposes. It has a $3-\mathrm{dB}$ bandwid th up to 15 MHz and an inputimpedance of $1 \mathrm{M} \Omega$ shunted by 30 pF .

When using the dual-trace unit, the input attenuator of the oscilloscope cannot be employed unless the unit is made capable of handling large signal amplitudes. In any case, the two signals to be observed may have very different amplitudes. It follows that each channel must have its own input


Fig. 1. A passive probe designed to give attenuations of 3.33:1 and $10: 1$ according to the position of the switch and to reduce the capacitance of the cable and oscilloscope in the same ratios.
attenuator. If it were not for one thing, amplification of the signals would be unnecessary. This thing is cable capacitance. A minimum of 3 ft of coaxial cable is needed for the input. If this is ordinary $75-\Omega$ cable its capacitance will be about 60 pF . Special low-capacitance cable can be used, but is less readily available, and even then its capacitance is unlikely to be under 30 pF . The usual practice is to use a passive probe which attenuates the signal to $\frac{1}{10}$ of the input and at the same time reduces the capacitance by the same amount. This is eminently practical, but necessitates an amplifier with a gain of 10 times to make up for the loss.

At this stage we did not know what gain and bandwidth would be practicable. We felt that the minimum requirement was a $3-\mathrm{dB}$ bandwidth of 5 MHz and that it should be as much greater as proved reasonably practicable. We felt it might be hard to get a gain of 10 times with a bandwidth of more than this, and decided that a compromise was desirable. What we initially decided was this. There would be an input probe with an attenuation of $1 / 3 \cdot 33$. With a total cable plus unit input capacitance of 70 pF , this would give a probe input capacitance of $70 / 3 \cdot 33=21 \mathrm{pF}$ about. For the next range, a resistance would be switched in series to give an attenuation of $1 / 10$, making the capacitance 7 pF .

The arrangement is sketched in Fig. 1, where $R_{0}$ and $C_{0}$ are the input resistance and capacitance of the dual-trace unit. The attenuation is

$$
\frac{1}{\alpha}=\frac{R_{0}}{R_{0}+R_{1}}
$$

when the switch is closed and

$$
\frac{1}{\alpha}=\frac{R_{0}}{R_{0}+R_{1}+R_{2}}
$$

when it is open. If $\alpha=3.33$ and $R_{0}=100$ $\mathrm{k} \Omega, R_{1}=233 \mathrm{k} \Omega$ and if $\alpha=10, R_{1}+R_{2}$ $=900 \mathrm{k} \Omega$, whence $R_{2}=667 \mathrm{k} \Omega$. These are non-standard values, but can be obtained from the combination of two or three preferred values. With an amplifier gain of 3.33 times, a 1-V input with S closed will give 1-V output. A $3-\mathrm{V}$ input with the switch open will give $(3 / 10) \times 3.33=0.99 \mathrm{~V}$ $=1 \mathrm{~V}$ output.
The combination of this with one $10: 1$ attenuator in the unit would provide ranges of $1 \mathrm{~V}, 3 \mathrm{~V}, 10 \mathrm{~V}$ and 30 V , which would suffice for many, if not most requirements. The input resistance would be $333 \mathrm{k} \Omega$ on
the 1 V and 10 V ranges and $1 \mathrm{M} \Omega$ on the 3 V and 30 V ranges.
Frequency compensation of the potential divider requires that all time constants be alike. If the cable capacitance is $C_{c}$, this means

$$
\left(C_{0}+C_{c}\right) R_{0}=C_{1} R_{1}=C_{2} R_{2}
$$

and there must be trimmers $C_{1}$ and $C_{2}$ in the probe to enable these capacitances to be adjusted precisely. Easy adjustment requires a square-wave signal of suitable repetition frequency. Adjustment is carried out for a square corner to the signal. If $C_{1}$ or $C_{2}$ in Fig. 1 is too small the square-wave has rounded corners as shown at (a) in Fig. 2, whereas if it is too large there is overshoot as at (c). The correct adjustment gives the square corners (b). If the input signal is a good one, the adjustment is remarkably easy to carry out.

A square-wave generator is not always available, of course, but the switching circuits of the dual-trace unit will, in fact, be operated by a square-wave generator and it was felt that this could be arranged to provide the signal for adjusting the attenuator. At this stage, this was merely noted as a possibility.

At this point it may be advisable to say why $100 \mathrm{k} \Omega$ was selected for $R_{0}$. It is usual for an oscilloscope to have an input resistance of $1 \mathrm{M} \Omega$. This arose originally because this was about the highest stable value which could readily be provided with valve circuits. It is actually on the low side when the c.r.o. is used to investigate valve circuits, and a 10:1 probe is often used to


Fig. 2. With the capacitors $C_{1}$ and $C_{2}$ properly adjusted a square wave is correctly reproduced (b). Too small capacitance gives rounded corners (a) while too much capacitance gives overshoot (c).
bring it up to $10 \mathrm{M} \Omega$ when the signal is large enough.

Most transistor circuits are of a good deal lower impedance and $1 \mathrm{M} \Omega$ is ample for them. It is more important to reduce capacitance than to increase resistance. The use of high value resistors is to be avoided as far as possible because they are more likely to be unstable than lower values and are certainly more affected by surface leakage in damp weather.
It is important that the input resistance $R_{0}$ be substantially defined by a resistor and not by a semiconductor. If $R_{0}$ is $100 \mathrm{k} \Omega$, this means that the input resistance of the first stage should not be less than $5 \mathrm{M} \Omega$ if its effect is to be small. This input resistance is usually highly variable. Of course, if a field-effect transistor is used a much higher input resistance is obtainable, but at this stage we had not decided which would be used and we initially chose values which would suit a bipolar transistor.

## Signal Control

It will be noted at this point that we had tentatively decided on an amplifier gain of 3.33 times because we thought that this should be easy to obtain. We note that the scheme worked out has two disadvantages. One is that, as already mentioned, the input impedance varies with the range. The other is the practical difficulty of including a switch, two resistors and two trimmers in a probe head without making it unwieldy. Further, with one range control on the probe and the other in the instrument, one must remember to note the setting of both to determine the actual range employed.

It would clearly be more convenient for the probe to give constant attenuation for then it need contain only one resistor and capacitor and the input impedance would be the same on all ranges viz. $1 \mathrm{M} \Omega$ and 7 pF . Two attenuators in the instrument would singly and in combination provide ranges of $1,3,10,30 \mathrm{~V}$; the attenuators having ratios of $3 \cdot 33: 1$ and $10: 1$, under the control of a range switch. The possibility of this depends on being able to obtain a stable gain of 10 times from the amplifier with an adequate bandwidth, and at the start we did not know whether this was reasonably practicable. The gain control range required is, of course, unaltered and remains at about $3 \cdot 5: 1$, for it has only to fill in the gaps in the attenuator steps.

Whatever the input stage, protection against overloading is required. Few transistors are rated for more than 6 V reverse base bias and there is always the possibility that the probe will be connected inadvertently to the 240 V supply mains of 340 V peak value or 360 V if $6 \%$ high. Protection is obtained by connecting two diodes back to back across $R_{0}$, as shown in Fig. 3. On the lowest range $R_{1}$ is always in circuit and limits the current to $360 / 233: 1.54 \mathrm{~mA}$. This is the maximum diode current. and few diodes will drop more than 1 V at this current.

The signal amplitude is $0.3 \mathrm{~V} \mathrm{p}-\mathrm{p}$ and we hope that, even without bias, silicon diodes will not conduct on it. The circuit


Fig. 3. This diagram shows the probe of Fig. 1 connected via the cable to a further attenuator of $10: 1$ ratio and diodes arranged to protect the amplifier against overloads.
has now grown to the form of Fig. 3.
One other decision had to be made. This was whether to make provision for a d.c. input. In any case, a series capacitor would be provided for a.c. only. The writer's experience is that a d.c. input is used only rarely and that when it is wanted it often cannot be used, because the same input range cannot be used for d.c. and a.c. together unless the two are comparable in magnitude. The input circuits become complicated if a bipolar transistor is used because of the base supply voltage. It was decided, therefore, to make provision for a.c. inputs only.

The capacitor can be inserted in series with the cable at the output end and the effective resistance is $333 \mathrm{k} \Omega$ on the 1 V and 10 V ranges and $1 \mathrm{M} \Omega$ on the 3 V and 30 V ranges. The drop in response (i.e., the sag) at a time $t$ after the application of a unit step is simply $t / C R$. For a $50-\mathrm{Hz}$ square wave, $t=10 \mathrm{msec}$. If $C=0.5 \mu \mathrm{~F}$ and $R=$ $333 \mathrm{k} \Omega$, the sag is $10^{-2} /\left(5 \times 10^{-7} \times 3.33 \times\right.$ $\left.10^{5}\right)=1 / 16 \cdot 65=0.06=6 \%$. This is as much as should be tolerated and $0.5 \mu \mathrm{~F}$ is the minimum capacitance to be used. For a $1 \mathrm{M} \Omega$ input resistance, a $0.22 \mu \mathrm{~F}$ capacitor can be used to give a sag of $4.5 \%$.

For the initial experiments we did not build the full arrangement of Fig. 3 but used only the simplified system of Fig. 4. The probe must always be screened, of course, and for bench work it proved essential to screen the capacitor to prevent hum pick-up.

At this stage of the proceedings we had solved in principle the input circuit problems and could define the amplifier requirements more closely, which were:

1. To operate into an output load of $1 \mathrm{M} \Omega$ shunted by 55 pF ( 30 pF oscilloscope input capacitance plus 20 pF for 1 ft cable plus 5 pF strays)
2. To provide an output of at least 1 V p-p 3. To give a voltage amplification of 3.33 times (N.B. It was noted that if it should prove possible to obtain an amplification of 10 times this might be adopted and the attenuator system altered).
3. To have a continuous gain control of at least 3.33:1.
4. To be able to handle an input of up to 1 V p-p (so that full output could be obtained with the gain control at minimum). 6. To include a shift control so that the traces could be moved vertically and independently on the screen. A range of $\pm 0.5 \mathrm{~V}$ at the output would be sufficient.
5. The gain and shift controls to have no interaction.
6. The whole amplifier to be stable and easy to set up.
With regard to the last item, it was considered that as this is a piece of test equipment, which will normally be used under laboratory conditions, it would suffice to take the temperature range as $\pm 12.5^{\circ} \mathrm{C}=$ $\pm 22 \cdot 5^{\circ} \mathrm{F}$ about a mean of $65^{\circ} \mathrm{F}$. This covers room temperatures of $42 \cdot 5-87 \cdot 5^{\circ} \mathrm{F}$.
The mean room temperature is thus $18.3^{\circ} \mathrm{C}$. The internal case temperature, which is the ambient of the transistors, is higher than this by what is at present a completely unknown amount, but it will vary with the room temperature and by the same amount. Transistor junctions will be higher than the ambient by an amount depending on their dissipation. Most small transistors have a thermal resistance between junction and case of about $0.5^{\circ} \mathrm{C} / \mathrm{mW}$. Anticipating a little, few, if any, transistors will dissipate more than 20 mW and so their junctions will not be more than $10^{\circ} \mathrm{C}$ above the ambient. No great attention need thus be paid to temperature.

In what follows, we shall assume at first that all junctions are at $25^{\circ} \mathrm{C}$ because this is the figure for which transistor characteristics are usually quoted. Corrections can be applied later. Because of the low power needed in this case, no dangerously high dissipation will occur, and the only important thing to watch is that the case is adequately ventilated. Apart from this the only effect of choosing the wrong design temperature is to change slightly the required bias voltages and as they may in any case have to be adjustable to allow for other


Fig. 4. Simplified probe used in experimental work, and input coupling capacitor to remove d.c.
variations, the result is likely to be trivial.
Before concluding this part, it will be well to say something about the output stage which is controlled by the electronic switch. The arrangement referred to earlier, which was used for some experimental tests, is shown in Fig. 5. The transistors $\operatorname{Tr}_{1}$ are the output transistors of the two signal channels, and they are switched by $\mathrm{Tr}_{2}$ which have square waves applied in opposite phase to their bases; when $\operatorname{Tr}_{2 a}$ conducts $\operatorname{Tr}_{2 b}$ is cut off and vice versa.
When a $T r_{2}$ is cut off the $T r_{1}$ to which it is connected operates as a normal amplifier with collector load $R_{c}$ and emitter resistor $R_{E}$. When a $T r_{2}$ is conductive it drains sufficient current through $R_{E}$ to cut off the $T r_{1}$ to which it is connected. $T r_{1 a}$ and $T r_{1 b}$ have a common load resistor $R_{c}$ and in this way the signals from the two channels are alternatively routed to the oscilloscope.
The oscilloscope input capacitance is about 30 pF and 1 ft of coaxial cable adds 20 pF . With 5 pF for strays, the total capacitance is 55 pF . If $R_{\mathrm{c}}$ is $330 \Omega$, then at 5 MHz , the response is $-20 \log \left[1+\omega^{2} C^{2} R^{2}\right]$

$$
=-10 \log \left[1+0.57^{2}\right]=-1.22 \mathrm{~dB}
$$

At 10 MHz , it is -3.61 dB . This is very reasonable as a starting point.
If $R_{E}=R_{C}$ the gain will be unity, or nearly so.
With a minimum supply of 10.5 V , maximum output demands that $V_{C E}$ be one-half of the supply voltage and so $I_{C}=5.25 /$ $0.66=7.95 \mathrm{~mA}$. The emitter is then 2.625 V above earth and the base about 0.65 V higher, or about 3.3 V . The maximum signal output will then approach 5.2 V p-p. The collector dissipation will be $5.25 \times 7.95$ $=41.8 \mathrm{~mW}$. Each transistor $T r_{1}$ operates for only $50 \%$ of the time, however, so each has a mean current of 4 mA and a mean dissipation of 21 mW in round figures.

Experimentally, it was found unnecessary to operate at quite such a high current and the decision was made to set $V_{B}$ at 2.7 V ,


Fig. 5. This diagram shows the two output stages $\mathrm{Tr}_{1}$ of the two signal channels. These are turned on and off alternately by
transistors $\operatorname{Tr}_{2}$ which are in turn driven on and off by push-pull square waves on their bases.
making $V_{E}=2.05 \mathrm{~V}$, and $I_{C}=6.21 \mathrm{~mA}$. Consequently, $V_{C E}=10.5-4.1=6.4 \mathrm{~V}$ and the dissipation is 39.9 mW . With $V_{c c}=$ 13.5 V , if $V_{E}$ is unaltered the current is unchanged and so $V_{C E}$ rises by 3 V to 9.4 V and the dissipation to 58.4 mW . The maximum mean dissipation is thus 29.2 mW .

Typically, the thermal resistance is $0.5^{\circ} \mathrm{C}$ / mW , and $V_{B E}$ changes by $2 \mathrm{mV} /{ }^{\circ} \mathrm{C}$. Thus $V_{B E}$ falls by $1 \mathrm{mV} / \mathrm{mW}$ for a constant ambient temperature. The change of mean dissipation with $V_{c c}$ is $29.2-21=8.2 \mathrm{~mW}$ and so $V_{B E}$ decreases by 8.2 mV when $V_{c c}$ is at its maximum, and $V_{E}$ rises by the same amount and $V_{C E}$ drops by twice this, or 16.4 mV . The current rise is $0.0082 / 0 \cdot 33=$ 0.0249 mA . At $V_{c c}=13.5 \mathrm{~V}$, therefore, $I_{C}=$ 6.235 mA and $V_{C E}=9.4-0.0164=9.384 \mathrm{~V}$, making $\mathrm{P}_{\mathrm{c}}=58.5 \mathrm{~mW}$. The change is quite negligible.

The normal output is 1 V p-p maximum. It is desirable to design for twice this to ensure a factor of safety; this is $1 \mathrm{~V}_{\mathrm{p}}$. The base of $\operatorname{Tr}_{1}$ swings from 1.7 V to 3.7 V with respect to earth, since the bias is set at 2.7 V . To cut-off $\operatorname{Tr}_{1}$, therefore, $\operatorname{Tr}_{2}$ must draw sufficiert current through $R_{E}$ to bring the emitter of $\operatorname{Tr}_{1}$ at least 3.7 V above earth. The current must thus be at least $3 \cdot 7 / 0.33=11.2 \mathrm{~mA}$. The BC107 transistor has a $V_{E B}$ rating of 6 V maximum. Thus, $V_{E}$ must not exceed $6+1 \cdot 7=7.7 \mathrm{~V}$ and so $I_{\mathrm{c} 2}$ must be under $7 \cdot 7 / 0.33=23.3 \mathrm{~mA}$.

If $T r_{2}$ is saturated with a high supply voltage $(13.5 \mathrm{~V}), \quad V_{C E 2} \approx 0.2 \mathrm{~V}$, and the total resistance must be greater than 13.3/ $23 \cdot 3=0.57 \mathrm{k} .2$. A resistance of more than $570-330=240 \Omega$ must be included in the collector circuit to limit the current. If the current is to exceed 11.2 mA on low supply voltage $(10.5 \mathrm{~V})_{;}$- the resistance must not be greater than $10 \cdot 3 / 11.2=916 \Omega$, so the collector resis:ance must be under 916-330 $=586 \Omega$. This assumes that the base current is negligible, which may not be true under saturated conditions. We thus see that the collector resistance of $\mathrm{Tr}_{2}$ must lie between $240 \Omega$ and $585 \Omega$, and $470 \Omega$ would seem a suitable choice.

With a conventional bistable driving $\operatorname{Tr}_{2}$ at its base, the bistable output will vary from


Fig. 6. The circuit of Fig. 5 redrawn with component-vaues and protective resistors in the base and collector circuits of the switching transistors.
about 0.2 V to perhaps 2 V below $V_{c c}$. It may be less than this, but taking this figure, the maximum will be 11.5 V . The emitter voltage of $T r_{2}$ will be at least 3.7 V , so the effective base-emitter drive will be $11.5-$ $3.7=4.8 \mathrm{~V}$. If we arbitrarily limit the base current to 0.5 mA , a series resistor of $9.6 \mathrm{k} \Omega$ $\approx 10 \mathrm{k} \Omega$ must be placed in series with the base of $T r_{2}$. The resistor can, in fact, be a little less than this because the above figure includes the output resistance of the bistable which is likely to be $1.5-2.5 \mathrm{k} \Omega$.

The output stages and their switching transistors are shown in Fig. 6 with the above calculated circuit values. A final decision, of course, depends on a trial. We might find, for example, that $10 \mathrm{k} \Omega$ base resistors make the switching speed too low and we may have to think again.

So far the supply voltage has been considered but little. It is, however, obvious that with the low signal voltage no high voltage supply is needed. The output stage could, in fact, be designed for a 6-V supply. As will appear later, the amplifier really demands more and the decision was made quite early to adopt a nominal 12 V supply. It was desired to avoid a stabilized supply and so a tolerance of $\pm 1.5 \mathrm{~V}$ on the supply was allowed. It was thought that this would be sufficient to cover a $\pm 6 \%$ mains voltage and component tolerances.

## High-speed cassette duplicator

A tape speed of $1.9 \mathrm{~m} / \mathrm{s}$ ( $75 \mathrm{i} . \mathrm{p} . \mathrm{s}$ ) is used in a new cassette duplicating system, intended for schools, libraries and the like which is being produced by Ampex. The equipment consists of a master unit which plays back the master tape to five slave duplicating units. Each duplicating unit will handle 45 , sixty-minute playing time cassettes, in one hour; all tracks are duplicated at the same time. The sequence of events goes something like this: With a cassette in position a slave unit will carry out the recording in 45 seconds; it then takes 17 seconds to rewind the tape which is done at $3.8 \mathrm{~m} / \mathrm{s}$ ( 150 i.p.s.); finally the cassette is ejected and a new one is automatically loaded, accounting for a further five seconds. The system, which has five slave units, will therefore produce five duplicated cassettes every 67 seconds.

The tape transports employ vacuum servo columns. The tape is pulled out of the cassette into vacuum chambers and against the heads. The result is close tape-to-head contact and precise and gentle tape handling despite the very high tape speed. The bandwidth of the electronics is 320 kHz .

## Announcements

An S-band air surveillance radar, the AR-15, has been introduced by Plessey which replaces the AR-1 introduced in 1965, of which over 100 (valued at approximately $£ 10 \mathrm{M}$ ) are now in service throughout the world. The AR-15 is available in both static and air transportable versions. It uses fully variable polarization. low noise parametric amplifiers, tunable magnetrons, digital moving target indication, background averaging techniques for clutter suppression. and multi p.r.f. integration for best target response.

Ericsson Marine, the newly formed marine communications department of the Ericsson Group, has set up a marine training school for ships' radio officers at the Norway Trade Centre in Pall Mall London. The first three-week course, for Cunard officers, began on 5th July. Initially, courses will be confined to experienced ships' radio officers and electronic technicians to familiarize them with the company's équipment.

A new collective call sign, GZXV, has been allocated to Ericsson Marine. It will be principally used to facilitate 'all ships' calls in the operation of the Ericsson Marine service to shipowners.

A course of eight evening lectures on video recording systems starts at Norwood Technical College, Knight's Hill, London, SE27. OTX. On 19th October. Fee £2.

The Service Division of Marconi Instruments Ltd has been appointed as an approved repair and calibration centre for the Salford Electrical Instruments range of multirange test instruments. Both companies are in the GEC Group.

Eight of Canada's major civil airports are to have Marconi 'bright' radar displays, type S3006, incorporated into their air traffic control systems. The value of the order is in excess of $£ 100,000$.

The Carrier Corporation, of California, has announced an agreement in principle for the acquisition of Reliance Controls Ltd, of Swindon, Wiltshire, formed in the 1930s. Bowmar Instrument Corp., of Fort Wayne. Indiana, at present own $55 \%$ of the Reliance share capital and Booker McConnell Ltd of London, $45 \%$. The transaction will involve approximately $£ 0.25 \mathrm{M}$.

Hamlin Electronics Inc.. reed switch manufacturers of Wisconsin. U.S.A.. have acquired Inter-Market Services Ltd, and re-formed it as Hamlin Electronics Ltd. The new company will market the complete range of Hamlin reed switches and power packs in the U.K. and Scandinavia, as did Inter-Market Services.

Servicing of test gear of all types is offered by a new service introduced by S. C. Murison, 9 Leas Road, Warlingham, Surrey CR3 9LN. (Tel.: 01-820 3830.)

Pye Telecommunications Ltd has appointed the Hallicrafters Company, of Illinois, exclusive U.S.A. distributor of its land mobile radio equipment.

UK Solenoid Ltd, rotary switch and contractor manufacturers, are moving from Hungerford, Berkshire, to 115 London Road, Newbury, Berkshire. (Tel: Newbury 5991. )
T.E.M. Sales Ltd, of Crawley, Sussex, have been appointed distributors for R.E.M. Inc., of California.

# Ten Practical F.E.T. Source-follower Circuits 

by J. O. M. Jenkins*, M.Sc.

Virtually every source-follower configuration can be covered from ten basic circuits, and by considering the related parameters a designer can obtain consistent performance despite inherent device variations. It is true to say that insufficient knowledge and a paucity of written matter has rather inhibited the use of f.e.ts in circuit design. This is regrettable, as the high input impedance and low output impedance of the field-effect device suits it to impedance transformations with bipolars.
There are two basic connections for source followers-with gate feedback and without gate feedback, and for simplicity these are taken separately.

## Biasing without feedback

1. A self-bias arrangement in which the voltage drop across $R_{S}$ biases the gate through $R_{G}$ Since no gate-to-source voltage ( $V_{G S}$ ) can be developed when $I_{D}=0$, the self-bias load line will pass through the origin. Using the 2 N 4339 as a standard for this and the other configurations. the quiescent drain current lies between 0.25 and 0.55 mA when $R_{S}=1 \mathrm{k} \Omega$. Hence the quiescent output voltage lies between 0.25 and 0.55 V .
2. A similar arrangement to the above with a negative supply ( $-V_{S S}$ ) added. This provides an advantage over the first arrangement: namely that the signal voltage can now swing negatively to approximately $-V_{S S}$. The two bias lines shown are for $V_{S S}=-15 \mathrm{~V}$ and $V_{S S}=-1.6 \mathrm{~V}$. In the first case the quiescent output voltage lies between +0.18 and +0.74 V ; in the second between +0.3 V and +0.82 V .
3. Here a current source improves drain-current $\left(I_{D}\right)$ stability, hence the bias load line will be horizontal when $I_{D}=$ constant current. For $I_{D}$ $=0.3 \mathrm{~mA}$ the quiescent output voltage is between +0.15 and 0.7 V .
4. This is similar to 3 , except that the current source is now f.e.t. A which allows constant current, the value of which corresponds to a $V_{G S}=0$ volts. It will be seen that f.e.t. A loses current linearity as its $V_{D S}$ approaches zero, so that this technique can only be used to bias f.e.ts which have a significantly higher pinch-off voltage than the f.e.t. forming the current source.
5. By using a pair of matched f.e.ts, one as a source follower and the other as a current source, the operating drain current ( $I_{D Q}$ ) is set by $R_{S 2}$. In this case ( $1.5 \mathrm{k}_{\Omega}$ ) the drain current can be in the range 0.2 to 0.42 mA (as shown by the intercepts). However, as the f.e.ts are matched $V_{G S I}$ $=V_{G S 2}$; and since $I_{D I}=I_{D 2}$, by making $R_{S I}=R_{S 2}$ the voltage across A-B will equal the voltage across C-D, which in this case is zero. This arrangement exhibits zero or near-zero offset, and if the f.e.ts are temperature matched at the operating $I_{D}$, the arrangement will provide zero or near-zero temperature drift.

## Biasing with feedback

The following circuits appear in the same sequence as before for comparative purposes. In each case $R_{G}$ is returned to a point such that almost unity feedback is provided to the lower end of $R_{G}$ If the value of $R_{S}$ is selected so that $R_{G}$ is returned to zero d.c. volts (except for 6), then the input/output offset is zero. $R_{l}$ is usually much larger than $R_{S}$.
6. This arrangement is suitable for a.c.-coupled circuits, and with $R_{S} \ll R_{I}$ provides near unity feedback. The bias load line is set by the value of $R_{S}$. The output load line, however, is the sum of $R_{S}+R_{I}$. The feedback voltage ( $V_{F B}$ ) at the junction of $R_{J} / R_{I}$ is determined by the intercept of this $R_{S}+R_{I}$ load line with the $V_{G S}$ axis. Quiescent output voltage is $V_{F B}-V_{G S}$.
7. Here $R_{S}$ can be trimmed to provide zero offset. Reference to the graph shows that $R_{S}$ will be between $670_{\Omega}$ and $2.5 \mathrm{k}_{\Omega}$ (and very much less than $R_{J}$ ). The source load line intercepts the $V_{G S}$ axis at $V_{S S}=-V_{G G}=-15 \mathrm{~V}$. Note that this load line is not perfectly flat; it has a slope of $-1 / 50 \mathrm{k}$ because the current source is not perfect, having a finite impedance however high.
8. Here $R_{l}$ is replaced by the ideal current source, and as this has theoretical infinite impedance, the load line is now perfectly flat.
9. By taking the output from the top of $R_{S}$, output impedance is reduced, and $R_{S}$ must be trimmed if the circuit is to operate effectively. The constant-current load line ( $I_{S}=0.3 \mathrm{~mA}$ ) and the effect of a $1 \mathrm{k}_{\Omega}$ source resistor is shown to provide an offset voltage between 0.2 and 0.75 V . The intercept of the $R_{S}$ load line and the $V_{G S}$ axis sets the voltage ( $V_{F B}$ ) at the junction of $R_{S}$ and the current source. For $R_{S}=1 \mathrm{k}_{\Omega}$, $V_{F B}$ will lie between -0.1 V and -0.45 V . Since $V_{F B}$ appears at the gate, it must be zero if the d.c. input impedance of the circuit is to be preserved. This can be done by trimming $R_{S}$ (dotted line) the biasing, then reverting to that of circuit 8 .
10. This is identical to circuit 5 except that feedback is added to raise the input impedance.

[^3]

## Summary

Circuits 1, 4 and 6 can accept only positive and small negative signals, as the source resistors are to ground. All other circuits can handle large positive and negative signals inhibited only by the available supply voltages and device breakdown voltage. Circuits $3,4,5,8,9$ and 10 employ current sources to improve $I_{D}$ stability and improve gain. Of these 4,5 and 10 employ f.e.ts as current sources. Circuits 5,7 and 10 employ a source resistor, $R_{S}$, which may be selected to provide a quiescent output voltage equal to zero. Circuits 5 and 10 use matched f.e.ts. $R_{S}$ is selected to set $I_{D}$ near the specified low-drift operating current. The input-output offset voltage is zero.

## Simple Crosshatch and Dot Generator

## A generator developed from the circuit published in the September 1968 issue which is cheap enough to install permanently in a colour television receiver

by A. W. Critchley*

The crosshatch pattern of white lines has proved to be the best type of pattern to carry out the convergence adjustments on a television receiver, although white dots are sometimes used. Either pattern is possible with the circuit described by means of a changeover switch or link.

The generator has four disadvantages as can be expected with such a simple device: the receiver has to be synchronized by a transmitted programme; the pattern position on the screen depends on the type of pulses feeding the generator; the pattern can occur during some of the
*Television Equipment Division of E.M.I. Electronics Ltd.
flyback time causing a foldover; and the horizontal lines may not be evenly spaced. The latter three disadvantages are not very serious provided that the pattern is stationary and the lines are fewer in number than the normal crosshatch pattern of some twenty-six in each direction.

## Waveforms

The waveform required from the generator consists of two independent sets of pulses representing the vertical and horizontal lines of the crosshatch. Vertical lines are some 200 ns wide with a repetition every $5 u$ s or so, but occurring only during the active, or scanning, line time, of the pic-
ture which is approximately $52 \mu$ s for 625 line systems). Horizontal line pulses last for one such active line and recur once every thirty-two lines or so, also only during the active line-times of the picture. The repetition rates of these horizontal pattern lines are not important provided that they occur only during the picture time and they are steady. The actual number of crosshatch lines is continuously variable in both directions over a three to one range.

V'ertical lines: These are generated by a multivibrator which is permitted to run only during the active lines of the picture


Fig. 1. Block diagram of the crosshatch generator.
as both line and field blanking are applied to prevent any pattern during flyback time. This blanking depends on the widths of the timebase pulses used and varies from receiver to receiver. It is likely that the blanking will not be perfect and some foldover of the pattern is to be expected depending on the receiver.

Horizontal lines: The basic oscillator is a multivibrator which is driven by field flyback pulses. The output square wave is differentiatè to form a pulse of about 64irs duration and is used to open a gate which is also fed with narrow linefrequency pulses. The output of this gate will consist of one narrow line pulse for every period of oscillation which is given the timing of the trailing-edge of the line-flyback driving pulse by differentiation before the gate. This timing is also the start of the active-line-as near as can be obtained by simple means. An R-S bistable is triggered by this single pulse and is thereby turned 'on' at the start of the active line. The 'off' input of the bistable is fed with continuous line driving pulses which start at the end of the active line and finish before the 'on' pulse. The net result is an output from the bistable of one active line once per period of oscillation of the multivibrator.

The effect of varying the oscillator frequency is to cause a 'shuffling' of the horizontal lines as the optimum frequencies are passed through with a relatively
smooth variation in the number of horizontal lines obtained. These lines are always of the correct length.

## Circuit description (Fig. 1)

$C_{1}, D_{1}, D_{2}$ and $R_{1}$ form an excess-voltage protection circuit for the negative-going line-scan input pulses. Integrated circuit la amplifies and clips this signal to give clean rectangular positive-going pulses into i.c. 1b. This pulse is also fed to an attentuating network consisting of $R_{3}$ and $R_{4}$ which together form one timing resistor for the vertical line multivibrator i.c. 3a and 3 c .
$R_{3}$ and $R_{4}$ are virtually in parallel when the input to the network is low during the picture time and the multivibrator then oscillates normally. When the input from i.c. la is high the multivator is prevented from oscillating because the potential at the input of i.c. 3 a is such as to turn it off. $R_{3}$ is really an isolating resistor to remove the shunting effect of the low-impedance output of i.c. 1 a from the timing resistor $R_{4}$, but since the parallel combination of $R_{3}$ and $R_{4}$ is low, then the value of $C_{2}$ is correspondingly higher than $C_{3}$. By this means the oscillator always has the same conditions at the start of every picture line. $C_{2}$ and $C_{3}$, with $R_{5}$ and $R_{16}$ form the rest of the multivibrator.

The output from i.c. 1 b is also used to help to control the starting and stopping of the multivibrator and in fact improves the


Fig. 2 (Upper) Photograph of the printed circuit board shown actual size ( 101 mm in length). (Below) Drawing of the component side of the board. length).(Below) Drawing of the component side of the board.
linearity of the first space in the crosshatch pattern. There is a feed of field scan pulses to i.c. 3a to inhibit the multivibrator during the field flyback time.

The field-scan negative-going pulse is used to drive the horizontal line multivibrator i.c. 1 e and 2 c in the same manner as for the vertical oscillator except that the value of $C_{8}$ has to be kept low because of its physical size. Therefore the input resistor is replaced by a diode to provide automatic isolation of the timing resistor from the gate output.

Both the multivibrators generate approximately square waves and both of them feed differentiating networks. The vertical line network of $C_{4}$ and $R_{6}$ provides a positive-going pulse of some 200 ns width at the input to i.c. 3 b - the negativegoing pulses being ignored by this gate, because they merely turn the gate 'on' harder than it already is whereas the positive-edges turn it 'off' as required.
A similar network of $C_{10}$ and $R_{14}$ with $R_{17}$ generates the positive-going $64 \mu \mathrm{~s}$ pulse at the input to i.c. 2 a . The other input to i.c. 2 a is the positive-going pulse with the timing of the line-scan drive pulse trailing-edge, which is obtained by yet another differentiating network $C_{6}$ and $R_{9}$.

The negative-going output of i.c. 2 a, which is one narrow line pulse for every cycle of oscillation of the multivibrator, feeds the bistable input of i.c. 2 d . The other side of the bistable is fed from i.c. 1 b with cleaned-up negative-going line-scan flyback pulses. Integrated circuit $2 d$ provides the output of positive-going single active lines, or horizontal lines of the pattern, and these are combined with the vertical lines in i.c. 3d, via i.c. 1f, to form a crosshatch of 4 V peak-to-peak positivegoing pulses at i.c. 3d. output. To enable a single-pole switch to be used-or a simple link-for switching to dots-the invertor If has to be used in the feed to i.c. 3d and its input has a low value resistor $R_{7}$ to earth so that when dots are selected the input of i.c. If is virtually earthed and so its output is 'high' and permits i.c. 3d to act as an invertor for the dot signal from i.c. 3b.

The simple multivibrators used in this generator have the very poor stability factor of some $30 \%$ change in the period of oscillation per volt of supply.

## Construction and testing

Construction should present little difficulty if the printed circuit board illustrated in Fig. 2, is employed. Normally the amount of testing required for such a unit is very small especially with integrated circuit construction since the unit either works or it doesn't. However in the case of this crosshatch generator the supply arrangement and the various connections need to be optimized.

The value of $R_{15}$, the zener series resistance should be chosen to allow some 20 mA through the zener diode whilst the complete generator takes 40 mA making a total of 60 mA at 5.1 V .

Next the line pulse resistor $R_{1}$ should be chosen to give between 2.5 and 4 V
peak-to-peak at $C_{1}$. When this is so there should be an output from the generator. with the switch set to crosshatch. which can be fed into the luminance amplifier. $R_{16}$ can then be adjusted to give a suitable number of vertical lines.

For optimum results on the receiver, the colour should be turned off, the brightness increased and the contrast decreased, so that the receiver remains synchronized and the crosshatch appears on top of the picture.

For the best results the output signal should be fed into the luminance amplifier after the detector output amplifier stage, where the video is positive-going for white. $R_{8}$ determines the crosshatch amplitude. Feeding into the amplifier before the sync. separator does cause a slight problem with vertical sync. if the horizontal lines occur just before the field sync. pulse. However adjustment of the number of horizontal lines should prevent trouble in which the receiver 'chases its own tail'.

The field input resistor $R_{10}$ is chosen to give a peak-to-peak reading of 2.5 to 4 V at $C_{7}$. The polarity of $C_{7}$ depends on the input source. If the line \& field pulse sources do not exceed the i.c. supply voltages--at any time--then the protection diodes are not necessary and should be omitted. This should be observed by means of an oscilloscope.

When the field input pulses are correct the output should contain horizontal lines as well as vertical lines, but they will probably be jittering about and $R_{18}$ should therefore be adjusted. On turning this control clockwise the lines will be observed to get wider apart, and fewer in number, in reasonably smooth steps with certain positions of vertical jitter. It should be a simple matter to find several positions where the pattern is stationary.
$R_{17}$ can, now be set so that the horizontal lines are not of double thickness, but at the same time none are omitted. The optimum setting may vary slightly with different settings of $R_{18}$. The setting of $R_{16}$ may also slightly affect the jitter.

If the generator output resistor $R_{8}$ is sufficiently high then the removal of the generator's supply should cause no noticeable effects on the normal picture in which case this is a simple means of switching the crosshatch pattern off. Otherwise the output feed will have to be removed instead of switching off the supply.

The input and output connections may be made with ordinary insulated wire as all feeds are of relatively low impedance. but care should be taken with the run of the output lead due to stray capacitance reducing the amplitude of the vertical


Fig. 3. The prototype.
lines. If this happens then $C^{4}$ should be increased in value slightly.

## Appendix

Operation of crosshatch generator with B.R.C. $\mathbf{3 0 0 0}$ series colour receivers
$R_{1}$ should be $3.3 \mathrm{k} \Omega, R_{10}$ should be $8.2 \mathrm{k} \Omega, R_{8}$ should be $12 \mathrm{k} \Omega, R_{15}$ the zener resistor, is $470 \Omega, 3 \mathrm{~W}$-stood away from the board and $C_{5}$ should be 150 nF .
Line pulse: Chrominance board, Junction of $C_{337}, R_{359}$ and $R_{362}$. Solder the lead to the end of $R_{362}$ nearest the back of the receiver.
Field pulse: Field Scan board. Solder the wire to the top pin of the $R_{427}$ (field hold potentiometer).
Output: I.F. Board. $L_{117} / R_{127}$. Solder the wire to the end of this combination nearest to the front of the receiver (above $V T_{105}$ )-keep the length fairly short.
Earth: Convenient point on the i.f. board. $+30 V$. P.U. board. Solder the lead to the $5 \Omega$ resistor on the top of the lower boardthe end which goes to the positive end, of $W_{620}$. This lead should be taken via a suitably placed on/off switch to the generator.

## Method of operation

Turn off the colour, turn down the contrast, and turn up the brightness a little. The potentiometers should be adjusted for optimum results. Note that the horizontal lines upset the field timebase at certain settings because the crosshatch signal is put into the video chain before the sync. separator and the field timebase tends to chase 'its own tail'.

The Line and Field pulses should be 2.5 to 3.5 V p.p. at the inputs to the i.c's when the generator is switched on and about 2 V when off.

Both these waveforms are fairly wide and thus there is no visible fold-over or flyback.

The pattern is still visible under no-transmission conditions but the video noise masks the crosshatch and renders it unusable.

A worthwhile modification to the receiver would be to replace $R_{423}$ on the field scan board by a $470 \Omega$ potentiometer (from earth) with a $1.8 \mathrm{k} \Omega$ resistor in series. The potentiometer slider is then the field output point. The voltage at this point should be set to be less than 5 V p.p. The input capacitor and diodes on the generator field input can be deleted if this is done. The series resistor should be retained but changed in value to $220 \Omega$ or so, to protect the i.c.-otherwise $D_{5}$ could be retained instead.

A further improvement would be a series-regulator in the supply to the generator instead of the zener arrangement in order to reduce the supply impedance and thereby eliminate the slight tilting of the vertical lines at the right-hand-side of the picture which occurs when the zener supply is used. Each vertical section between horizontal lines is tilted by about a line thickness and whilst the effect does not affect the observation of convergence errors, the pattern does not look good.

## Square-root Circuit

# Using dual silicon-gate m.o.s.f.e.t. to give 1\% accuracy 

B. L. Hart* , B.Sc., M.I.E.R.E., M.I.E.E.E., and A. Cheetham*, M.Sc., M.I.E.R.E.

There are various ways of achieving the square-root operation-for instance the biased diode and multiplier techniques. $\dagger$ However, a simple low-cost approach is made possible by the capability to make an f.e.t. with an accurate square-law transfer characteristic, and of making pairs with their electrical parameters almost identical.

Consider the circuit arrangement shown below, in which the direct-coupled differential amplifier has a d.c. and lowfrequency small-signal voltage gain $A_{v}$, and $T r_{1}$ constitutes two matched p-channel enhancement-mode devices of a dual m.o.s.f.e.t. unit. One of the devices- $\mathrm{Tr}_{1 a}-$ is in the feedback network of the amplifier and passes the input current $I$; the other$T r_{1 b}$-is connected in series with the output of the amplifier and passes a small constant current derived from the interconnection of the integrated bipolar transistor pair $\operatorname{Tr}_{2}$. Transistor $\operatorname{Tr}_{1 b}$ cancels out part of the amplifier output voltage.
As $T r_{1}$ operate with drain-gate straps, each has a voltage-current relationship of the form

$$
I_{S D}=\Psi\left(V_{S G}-V_{T}\right)^{2}
$$

where $I_{S D}$ is the source-drain current, $V_{S G}$ is the source-gate voltage, $V_{T}$ is the threshold voltage, and $\Psi$ is the device constant (a function of material type, doping, geometry). (The order of the subscripts for $I, V$ corresponds to positive values of these quantities for a p-channel enhancement device.).
For simplicity in a first-order approximation assume that $T r_{1}$ have identical $V_{T}$ 's and identical values of $\Psi$. Assuming $A_{v} \gg 1$ and ignoring the input current, feedback action ensures that

$$
\begin{align*}
& I_{S D 1}=I=V_{I} / R=\Psi\left(V_{S G 1}-V_{T}\right)^{2}  \tag{1}\\
& \text { If } \sqrt{I_{S D 2} / \Psi} \ll V_{T} \text { then } V_{S G 2} \approx V_{T} \tag{2}
\end{align*}
$$

But,

$$
\begin{equation*}
V_{O}=\left(V_{S G 1}-V_{S G 2}\right) \tag{3}
\end{equation*}
$$

Using equations (1) and (2) in (3)

$$
\begin{equation*}
V_{o}=\sqrt{V_{I} / \Psi R} \tag{4}
\end{equation*}
$$

For the special case $\Psi R=1$ volt,

$$
\begin{equation*}
V_{O}=\sqrt{V_{I}} . \tag{5}
\end{equation*}
$$

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$\dagger$ C. A. A. Wass. "An introduction to electronic analogue computers". Pergamon: 1956

The successful practical realization of equation (4) depends on the choice of $T r_{1}$.

Now for $V_{I} \approx 0$, the amplifier output voltage is approximately $V_{T}$; thus for maximum range in $V_{o}$ m.o.s.f.e.ts with a low $V_{T}$ are required. This suggests the use of devices made by the silicon gate process. Preliminary measurements indicated a $V_{T}$ $<1.5 \mathrm{~V}$ and a $V_{T}$ matching of a few millivolts for the two devices of the recent silicon-gate dual m.o.s.f.f.et. type ME1202 (Marconi-Elliott Microelectronics) so this was used. The amplifier can be any good quality operational amplifier : a Burr-Brown type 3057/01 was used. Values for $V_{E E}$ and $R_{B}$ were chosen so that $\operatorname{Tr}_{2}$ (SL301-A, Plessey) in the "current mirror" configuration supply a current $I_{S D 2} \approx V_{E} / R_{E} \approx 5 \mu \mathrm{~A}$.
A convenient way of operating the circuit, and the one used for the tests reported here, is to set $V_{I}$ at a point $V_{I}^{*}$ in the middle of the desired input operating range, then adjust $R$ so that a precision digital voltmeter indicates $V_{o}^{\prime}=V_{o}^{*}=\sqrt{V_{I}^{*}}$. This ensures $\Psi R=1$ in equation (4) and hence the validity of equation (5) at the "set" point

A selection of the results obtained with one of the units is given in the table, in which the fourth column records the error $\varepsilon$ calculated from

$$
\varepsilon=\left|\left(V_{O}-\sqrt{V_{I}}\right) / \sqrt{V_{I}}\right| \times 100 \% .
$$

To obtain the readings shown the circuit was set up at $-V_{I}=-4.000 \mathrm{~V}$. For a $20-\mathrm{V}$ input range the maximum departure from


Using a m.o.s.f.e.t. with an accurate squarelaw characteristic in a feedback loop is the basis of this simple square-root circuit.
square-root law behaviour is less than $1 \%$. Other readings (not given) show this to be true also when the circuit is set up at $-V_{I}=-9.000 \mathrm{~V}$.

Test results showing accuracy of square-root circuit

| $-V_{1}$ | $\sqrt{V_{I}}$ | $V_{0}$ | $\varepsilon$ |
| :---: | :---: | :---: | :---: |
| -0.5000 | 0.7071 | 0.6960 | $1.6 \%$ |
| -0.7500 | 0.8660 | 0.8692 | $\uparrow$ |
| -1.000 | 1.000 | 1.009 |  |
| -2.000 | 1.414 | 1.424 |  |
| -4.000 | 2.000 | 2.000 | $1 \%$ |
| -9.000 | 3.000 | 2.998 |  |
| -16.00 | 4.000 | 4.009 |  |
| -20.00 | 4.472 | 4.492 | $\downarrow$ |
| -25.00 | 5.000 | 5.073 | $1.4 \%$ |

- Set-up point

Throughout $V_{I}$ has been taken as a positive quantity -the circuit extracts the square root of the magnitude of an applied negative signal. To find the root of the magnitude of a positive voltage the circuit - must be preceded with a unity-gain inverting amplifier.

## Correction

## Audio sweep generator

F. H. Trist has asked us to make some additions to the circuit of his suggested sweep generator (page 337, July issue). In the v.c.o., a $10 \mathrm{k} \Omega$ resistor should be connected at the junction of the $10-\mu \mathrm{F}$ coupling capacitor with the following resistor and to earth. In the output level amplifier, a $470-\Omega$ resistor should be connected between the negative input of the i.c. and earth. The three level-control resistors in the feedback loop should be reduced by three orders of magnitude. In the frequency-to-voltage converter, a $10-\mathrm{k} \Omega$ resistor should be connected between the negative input of the second i.c. and earth. In this circuit, we apologise for showing the X -output incorrectly connected. It should be taken from the wiper of switch $S_{e}$, and the common connection of the capacitors earthed.

## News of the Month

## Scientific fellowship for authors

A scientific fellowship, worth over $£ 750$, is to be awarded by the Butterworth Group to commemorate 25 years of scientific publishing. The Fellowship, to be presented annually from October, 1972, is designed to allow would-be authors to take time off from their work to write a book. By this means, each year, Butterworths hope to encourage a work on some aspect of a physical or biological science, or its application. Proposals will be judged both on academic merit and relevance to current research.

Candidates should work in a British university or institute or in an industrial laboratory of similar standing. Depending on the amount of work involved, the fellowship will be tenable for a period of three to twelve months, and during this time advances against royalties will be made to cover the loss of normal income. In addition an award of $£ 750$ will be made on acceptance of the manuscript.

The fellowship will be awarded by Butterworth's Scientific Advisory Board whose members are: Professor Sir Harold Thompson, C.B.E., F.R.S., (Department of Physical Chemistry, University of Oxford); Professor D. H. R. Barton, F.R.S., (Department of Chemistry, Imperial College, London); J. A. Charles
(Department of Physics, University of Bristol) and Professor J. L. Harley, F.R.S., (Department of Forestry Science, University of Oxford).

Applications must be submitted by 1st October, 1971 and must be backed by a head of department. It is expected that the fellow will be selected in the same month. Applicants should write for more information and entry forms to The Scientific Publisher, Butterworth Group, 88 Kingsway, London WC2B 6AB.

## Atlantic air traffic control by satellite

Further steps towards using satellite communication links for air traffic control are being taken with the award of a study contract to the Marconi Company by the Department of Trade and Industry. Under the contract Marconi's Radio and Space Communications Division will prepare a detailed analysis of the ground-based parts of a possible aeronautical satellite system for the North Atlantic. This will entail a detailed study of the ground equipment
necessary to relay several different types of information between aircraft and ground via satellite and to determine the best way of putting the study into practice.

Aircraft over the North Atlantic are under the control of oceanic air traffic control centres and the present system is under the jurisdiction of several centres including Gander in Newfoundland, Prestwick in Scotland, New York and Santa Maria (Azores). Aircraft report to these stations using normal h.f. radio, to give position information derived from their own on-board navigational instruments.
Improvements to the system are made continuously to cope with the demand of increasing transatlantic air traffic and it is in anticipation of the time when current methods are no longer effective, that consideration of satellite systems is being undertaken on both sides of the Atlantic.

## Computer telegram system

The Post Office has placed a $£ 3.25 \mathrm{M}$ order with Pye/T.M.C. for a computercontrolled telegram routing system which will replace electro-mechanical systems in 1973 at Cardinal House, Farringdon St, London. It will be the largest system of its type in the world and will be controlling the receipt and dispatch of the 21 million international telegrams handled in Britain every year.

Initially the equipment will receive telegrams for transmission abroad from international area offices throughout the country and will perform all the necessary switching and routing automatically. The same process will apply to telegrams received from abroad which will be automatically routed to the appropriate area office. Eventually the system will convert addresses on incoming telegrams to the telex address (if there is one) so that the message can be immediately sent over the telex network.


## Radar at Heathrow

Marconi Radar Systems has received an order from the Ministry of Defence (Aviation Supply), on behalf of the Department of Trade and Industry, to supply a high-power, 50 cm transmitter/receiver to replace radar equipment at Heathrow Airport which has been in service for twelve years. The new transmitter/receiver (type S2020) is a self-contained $500 \mathrm{~kW} \quad 50 \mathrm{~cm}$ equipment designed for use in coherent moving target indication systems and will be installed towards the end of the year. The power amplifier stage is a three-cavity klystron valve, with a typical life of 30,000 hours.

## Surveillance system for Southampton docks

An extensive surveillance system is to be installed to provide increased safety to shipping using the port of Southampton. The scheme is being carried out by the British Transport Docks Board. Decca Radar and Marconi Communications Systems have been awarded contracts totalling over $£ 0.25 \mathrm{M}$.

Decca Radar are to equip two unmanned radar stations, at Hythe and Calshot, from which data will be transmitted by microwave link to six 400 mm displays in the operations room at the port communications centre. At Calshot and Hythe the radar stations will consist of 7.6 m scanners mounted at a height of 33 m . Remote control of both stations will be effected by microwave link to the port communications centre The six displays to be installed by Decca in the operations room will be able to receive data from either unmanned station (two normally being fed from Hythe and four from Calshot). The Decca computerassisted measurement system will be provided for all six displays, and a Deccaspot system will be available on all pictures received from Calshot. The former system uses a small Honeywell computer to enable rapid and accurate measurements to be made of any point, such as a ship's position, relative to any other point on the display. Deccaspot, a method employing a series of bright spots on the display to depict with great accuracy any permanent feature required. will be used to delineate the centre of the navigation channel from Southampton Docks.

## Desk-top optical mark reader

Interscan Data Systems (U.K.) Ltd. normally associated with complex and expensive, optical character recognition machines, have announced a new low-cost relatively simple document reader. The new reader-there are two versions-can be operated by a company for as little as $\mathfrak{£ 2 , 0 0 0}$ per year. Once loaded the reader will continue to operate all day without attention.
The machine, called o.m.r. (optical mark reader) reads characters on special forms and gives an output in computer compatible code. As long as the characters are put in the correct position on the form they can be machine or hand printed.

The reading head, which is made to mechanically scan the rows of characters, consists of two photodiodes which simultaneously read the upper and lower halves of the characters. Only vertical sections of the characters are sensed, horizontal marks being redundant. The reading head also contains two magnetic proximity sensors which provide clock

Submarine cable repeaters being manufactured in an S.T.C. plant under clinical conditions. Repeaters of this sort will be used on a new $£ 22 M$ transatlantic cable (CANTAT-2) which will run from Widemouth Bay in Cornwall to Halifax in Nova Scotia. The 14 MHz coaxial cable will carry 1840 simultaneous telephone conversations: Repeaters will be fitted at intervals of about six nautical miles. S.T.C. have been awarded the contract by the Post Office and it is calculated that the cost is about $f 6$ per circuit per mile.

pulses, when a character is under the reading head, from castellations machined into a piece of metal mounted parallel to the moving reading head.

Document size can vary from $50 \times$ 100 mm to $216 \times 280 \mathrm{~mm}$ and the reading speed is up to 20 characters per second. The makers say that the equipment costs less than a paper tape station to hire and has ten times the throughput. To another piece of equipment the machines electronically look like a Teletype machine and therefore can be easily interfaced with other data processing equipment or the output can be recorded on a casette tape recorder.

## Motorists' laser warning system

Scientifica and Cook Electronics are working hard to find new applications for the laser. Recently they described a system, which could be used on small airfields, employing a laser to provide a visible glide path to assist landing aircraft.

A nother idea, and apparently a good one, entailed fitting photocell detectors on the nose of aircraft and connecting them to the aircraft's intercom system. The idea being that the control tower staff could contact an aircraft on the airfield very quickly in an emergency using a modulated laser beam regardless of the channel selected on the aircraft's radio.

An extension of this idea has resulted in photocells being fitted to a motor car, the
cells being connected directly to the a.f. stage of the car's radio so that it is possible to transmit warning messages to motorists by using a diffused laser beam directed down the centre of the carriageway. Trials have shown that this idea works well in practice.

## One-plus-one equals party line privacy

One-plus-one is the name given to a new piece of equipment which is to be installed on an experimental basis at 10,000 locations up and down the country by the Post Office. It enables two subscribers to share a line to a telephone exchange with complete privacy and if desired both subscribers can use their telephones at the same time.

A filter is fitted at the point where the line from the exchange divides to go to the individual telephones. One of the telephones operates in the normal manner at audio frequencies and does not require any additional equipment. Two carrier frequencies are used for the second telephone, 40 kHz for send and 64 kHz for receive. Equipment at the exchange and at the subscriber's premises carries out the necessary modulation and demodulation functions. Electronic equipment at the subscriber's end is powered by a small nickel-cadmium battery which is trickle charged over the line from the exchange. The system was designed by G.E.C's Telephone division laboratories at Aycliffe.

## Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

## Ceramic pickup equalization

Without reflecting on other parts of Mr. Burrows' article in the July issue I am appalled at his ability to read out of context.

His quotation from my book 'Pick-ups: Key to Hi-Fi' is given as a myth about 'electrical loading affecting the mechanical operation' of a pickup.

But the quotation clearly mentions correction by electrical means (via element capacity), of a mechanically accomplished equalization. It has nothing to do with damping mechanical resonances at all. It seems this myth belongs to Mr. Burrows. John Walton,
Windsor,
Berks.
I was interested in the excellent article by Mr. Burrows (July issue) basing the mechanical/electro independence of pickups on low energy conversion. This is the first time that I have seen this direction of approach.

However, it is only fair to point out that hosts of manufacturers other than Leak imply by instruction booklet or text or input circuit design that an approximation to velocity characteristics is achieved by connecting a piezo pickup across the relatively low value load of an R.I.A.A.-equalized input.

Surely the point of the exercise is that all quality amplifiers are deliberately equipped with R.I.A.A. low-level inputs so that the advantage can be taken of the optimum performance at the present state of the art provided by the magnetic cartridge?
The lack of simple solid-state, high resistance inputs prior to the f.e.t. obviously made it necessary for manufacturers to suggest a simple artifice to accommodate the minority of 'lower-fi' piezo users. Since the f.e.t. has become more commonplace and less costly, manufacturers who consider that the piezo cartridge is being treated unfairly are yielding designs with an f.e.t. input solely for piezo cartridges or in addition to the usual R.I.A.A.-equalized input. The piezo input is typically $2 \mathrm{M} \Omega$, and with this kind of cartridge mild bass roll-off is not always amiss.
In my judgment it is debatable whether manufacturers would have very much
call for an amplifier with a specifically engineered piezo input possibly requiring adjustment to suit the cartridge used. The hi-fi enthusiast is a magnetic man for various'reasons, and now that magnetic species of surprisingly high quality (in terms of the three main parameters of tracking performance, frequency response and crosstalk) are available for a few pounds the man hitherto piezo prone is turning towards electromagnetic energy for this programme source.

Apart from the obvious lack of true velocity coincidence by running a typical piezo across $47 \mathrm{k} \Omega$ into an R.I.A.A.equalized circuit, the major offence is pre-amplifier overload, since this partnership is not uncommonly practised without input attenuation. Bearing in mind the poor overload margin of such pre-amplifiers it is possible that this rates higher in the poor-piezo-quality stakes than lack of absolute equalizing.
Gordon J. King,
Brixham,
Devon.

## Transformer phase reversal

I am most grateful to your eminent contributor 'Cathode Ray' for throwing his very considerable professional weight behind the campaign for the truth about the transformer (June p.285). Although, as he explained, he had to argue the matter on paper without practical demonstration, I take it that your readers can and will check the experimental fact as to the phase relations between terminal voltages and currents in primary and secondary for themselves (see circuit); or perhaps not, since during the past seven years of teaching the experimental fact (preceded by two years of teaching orthodox phase reversal!) I have invariably found that lecturers will argue heatedly for an hour among themselves but when invited to make a five-minute measurement say they haven't got the time.

Electronics is above all (apart from instrument transformers, where only the direction of a wattmeter deflection is at stake) the field where phase cannot be fumbled. A power engineer, paralleling two 10 MVA transformers and assuming a
phase reversal in both, will come to no harm, being protected no doubt by the same Divinity which looks after children and drunks. But an audio amplifier designer getting his transformer polarities wrong in a feedback amplifier is going to produce fierce oscillations and a damaged loudspeaker. It would be helpful therefore, if manufacturers of interstage transformers who do mark winding starts and finishes,


If the starts and finishes of the windings are not marked they can be quickly established by measuring the inductance of the two windings in series. The connection giving the larger inductance is that in which the finish of one winding is connected to the start of the other winding.
and some others who don't, would tell us what the phase relations in their transformers are, and if manufacturers of feedback amplifiers using interstage transformers and output transformers would say what phase relations they assumed in designing their amplifiers and getting them to work so very satisfactorily. Victor Mayes, Gloucester Technical College, Gloucester.

## Audio sweep generators

While Mr. F. H. Trist is to be congratulated for answering the long-felt need for an audio sweep generator, we feel that his design (July issue) falls short of engineering requirements on several counts.

1. Sweep frequency range. The $10: 1$ frequency change satisfies only a small proportion of the possible uses; in fact only those for investigating narrow-band filters. A $1,000: 1$ change, from 20 Hz to 20 kHz seems a minimum specification for ampli-
fiers, tone controls and filters, and this sort of range is normally offered by commercial designs.
2. An amplitude accuracy of 1 dB is marginally adequate for transducer measurements, and not good enough for amplifier and filter work.
3. A sweep time of four seconds is only suitable for oscilloscopes with c.r.t. phosphors which most users are unlikely to have.
4. Logarithmic scaling of both frequency and amplitude axes in all graphical representation in audio engineering is normal and necessary. To give one example, it is not possible to differentiate between a 26 dB and a 40 dB notch filter on an oscilloscope, if the system responds linearly to amplitude. (A fast enough sweep time to make the use of a normal oscilloscope possible can be achieved only if the sweep is logarithmic.)

It seems to us that most of the drawbacks in Trist's design result from the wrong choice of oscillator. Any bridge-type oscillator is far too "sensitive" (in the sense that Bode gives the term') to achieve a wide frequency change without unreasonably close matching of components. The design considered demands $5 \%$ matching of f.e.t. drain-source resistances for only a $10: 1$ frequency range, and even then an a.g.c. network is needed to compensate for the varying losses in the bridge. Furthermore this a.g.c. system introduces another time constant into the oscillator, which is too long to allow for amplitude correction during the sweep.

Two alternatives to the bridge oscillator suggest themselves, if only to eliminate component selection and complex settingup procedures (Trist's calibrator alone contains fifteen pre-sets); these are the non sinusoidal oscillator ${ }^{2}$ and the two-integrator loop ${ }^{3}$. A switched Miller integrator, of which Trist's ramp generator is an example, can itself be frequency controlled by another ramp generator, producing a swept triangular wave with its amplitude independent of component matching and fixed only by the reference level of the comparator. Provided this triangular waveform is equilateral, a pure sinusoid can be obtained with a simple function generator. A diode network will produce better than $3 \%$ harmonic distortion ${ }^{2}$ and other methods easily better this ${ }^{4}$. Alternatively the two-integrator loop generates sine waves with amplitude fixed by a limiter and tracking errors between the two frequency varying elements produce proportionate errors in frequency only, none in amplitude. Frequency ratios of $1,000: 1$ are easily obtainable, in practice with both the above oscillator types.

We are working on a sweep generator using a two-integrator loop, which we hope to submit for publication shortly. Although our design requires a greater number of i.c. operational amplifiers, it does satisfy requirements $1-4$ above namely a $1,000: 1$ sweep range, good amplitude accuracy, fast sweep rate and logarithmic frequency and amplitude axes. We feel therefore that alternative oscillators to the Wien bridge should be considered by those interested in sweep oscillator design.
A. Falla, R. S. Snell, University of Sussex, Brighton.

1. H. W. Bode: Network analysis and feedback amplifier design (p52), D. Van Nostrand, N.J. 1945. 2. P. J. Kindlman: 'Sound synthesis: a flexible modular approach with i.cs', I.E.E.E. Transactions on Audio, Vol. AU-16 no. 4, Dec. 1968.
2. E. F. Good: 'A two-phase low-frequency oscillator', Electronic Engineering, Apr. 19, '57. 4. 'Triangular-to-sine convertor', Electronics, Vol. 38. no. 5, p96.

## The author replies:

It was with considerable interest that I read Messrs Falla and Snell's comments on my sweep generator. Before answering each point in turn, may I say that all of them occurred to me (unceasingly!) during design stages.
(1). Perhaps I did not state sufficiently clearly that there are four frequency ranges available at the flick of a switch, thus enabling $10-10^{5} \mathrm{~Hz}$ to be covered. This seems to me to be of greater use than squeezing the entire spectrum into some 4 in of c.r.o. display. I do not consider the range quoted by Falla and Snell as adequate; my system allows break points to be studied in detail-you don't gaze at the stars whilst tying your shoelace!
(2). It is doubtful whether a linearity of better than 1 dB is necessary in any audio system. In any case, displaying the input to the network under study will reveal where and by how much the amplitude varies during a sweep.
(3). I should point out that:
(a) Only the lowest range is limited to a 4 second sweep.
(b) Ideally, for normal c.r.o. work, a sweep rate in excess of $25 / \mathrm{sec}$. is required, in order to fool the eye. My system would generate garbage at this speed, even if the a.g.c. could respond fast enough, as there is no control over the starting phase of the oscillator. I assume that the two-lag system proposed will control this function, but sweeping at any rate faster than 0.1 of the minimum oscillator frequency will give little indication of response as frequency will change faster than phase.
(4). The prototype contained a logarithmic operator to provide the display timebase; this was of little practical use, due to the non-linearity of the voltage-resistance characteristic of the tuning devices. At the price- $£ 20$ including case-I don't apologize. As my unit does not attempt to process the signal from the observed network, how could it possibly be expected to provide a logarithmic amplitude display? Perhaps the writers would have me compress the signal to the network!

I do not follow the last sentence of (4). Only an antilogarithmic timebase function could permit faster sweep rates at low frequencies; this would diminish the phase problems detailed above; a logarithmic function must accentuate them.

I do not feel that $5 \%$ matching of two devices is too much to ask for. Falla and Snell mentioned diode shaping an equilateral triangular waveform to produce a low-3\% is low?-distortion sinusoid. If they were to use the switched Miller
integrator proposed, they would require to match a pair of current defining resistors to better than $5 \%$.

It is simply not true that amplitude correction is not applied during a sweep. The sweep frequency is much lower than the minimum oscillator frequency on each range; while the smoothing time-constant all but eliminates ripple from the oscillator, it is small enough to respond to the rampgenerator frequency-the fundamental frequency at which the amplitude attempts to change. I could scarcely claim a maximum deviation of $\pm 0.5 \mathrm{~dB}$ unless this were so.

Non-linear shaping of a triangular waveform can, by definition, never achieve the low-distortion of the Wien bridge. My instrument produced less than $1 \%$ distortion at 10 Hz ; on the upper three ranges no reliable reading could be made using a Marconi distortion factor meter.
F. H. TRist,

Stoke-on-Trent,
Staffs.

## Karnaugh map display

Fig. 5 in the article (published in April) showing the 'equivalent' circuit for the ladder network on Fig. 4 demands some clarification. It may have been tempting to suggest that for a 00 -input (Fig. 5(a)) the value of the equivalent series resistor does not matter very much (its value is not mentioned) as the operational amplifier, with this network connected to its inverting input and with the non-inverting input at ground, will have an 0 V -output whatever the input resistance. However, this resistance together with the feedback resistance determines the amplification of the signal applied to the non-inverting input. The right value is $10 / 3 \mathrm{k} \Omega$. In the same way Fig. 5(d) for an 11 -input is in error: the fact that no current flows through the two paralleled $10 \mathrm{k} \Omega$ resistors with unloaded ladder doesn't imply that the left hand resistors can be neglected when determining the equivalent circuit; the two remaining $10 \mathrm{k} \Omega$ resistors would give an equivalent resistance of $5 \mathrm{k} \Omega$ whereas in reality it should again become $10 / 3 \mathrm{k} \Omega$.

A first inspection of the ladder network shows that if the terminating resistor had not been returned to ground but instead used to feed the operational amplifier, where it sees a virtual earth, then the two voltage sources feeding the ladder would have seen exactly the same load ( $15 \mathrm{k} \Omega$ ) but this doesn't seem to be a necessary requirement. A second inspection shows

that the terminating resistor could have been dispensed with, even when the ladder output had not been connected to a virtual earth, without upsetting the digital-to-analogue conversion: for an unloaded ladder this would have given a $50 \%$ increase in voltage output. The third inspection reveals that two (or three) resistors instead of five (or six) would have done an even better job (Fig.1).

The clamping circuit shown gives considerable voltage loss; if necessary this can be improved upon by replacing the two silicon diodes by one low-voltage zener diode (about 3.3 V ). With a 4 V swing even an oscilloscope having an X-sensitivity as low as $1 \mathrm{~cm} /$ volt would still give a readable image. What happens exactly when one doesn't use a clamping network? From Mr. Crank's observations we may infer that only the clock pulse can be responsible for a double image but this is easier to remedy by using an asymmetric clock signal (small ON/OFF ratio); all other waveform distortions of the type shown will cause some of the 16 centre positions of the Karnaugh map to be shifted only slightly from an 'ideal' orthogonal raster in a reproducible way without provoking a double image. As these shifts are very small their effect will hardly be noticeable. A clamping network is therefore unnecessary!

The output swing being much larger now than in the original version the two operational amplifiers are redundant and the output to drive the 'scope can be taken direct from the digital-to-analogue converter. This results in considerable savings as the major part of the power supply can be dispensed with as well.

Having now only 5 or 6 V available for driving the phase-shift oscillator, its output amplitude is reduced. The two output resistances may thus have to be reduced as well. The larger resistance is required at the collector output and it is therefore this output upon which the $1 / 0$ switch should act in order to minimize the effect of the switching action upon the X -amplitude. It is also preferable to connect the switching transistor in the "inverted mode" in series with a capacitor.

The total savings are impressive: no operational amplifiers instead of two; one battery supply instead of three; a one-pole switch instead of a three-pole one; one electrolytic instead of three; no need for diode/resistor clamping; two transistors instead of four.

The final conclusion is that, without doing any difficult exercise and while retaining some of Mr. Crank's ideas and statements, his simplified logic display aid could have been further simplified.
G. J. NAALJER,

Limeil-Brevannes,
France.

## The author replies:

Perhaps Mr. Naaijer misunderstood the purpose of my equivalent circuits for the ladder networks. The object was to provide a simple explanation of how the square wave outputs of the counter became a stair-
case and to have considered the operational amplifier as well would have only confused the issue. If in my quest for simplicity I have offended the purists I apologize. Perhaps if the offending diagram had been labelled 'simplified circuit' instead of equivalent circuit (with all that this implies), the confusion would not have arisen. I would recommend that readers adopt Mr. Naaijer's digital-to-analogue converter circuit because of the component saving it affords.

I can assure Mr. Naaijer that some form of counter output waveform correction is essential to achieve a 'respectable' display. The zener diode idea was considered during the design but rejected on the grounds that two general-purpose silicon diodes can be purchased at a lower cost than one lowvoltage zener diode. By far the best solution was that proposed by A. W. Critchley in the May issue (p. 257). He suggested using four 'pull-up' resistors connected to the counter outputs.

The question of dispensing with the two operational amplifiers is debatable and depends on the use to which the unit is to be put. The original intention was that the device should be used in schools, I could not see many private constructors building it. In this application the device would very often be required to operate with long leads to the oscilloscope, or perhaps several oscilloscopes might be used, situated at strategic points around the classroom. In these circumstances the low output impedance afforded by the operational amplifier is essential as the visual effects of hum pick-up are particularly unpleasant with this type of display.

No trouble was experienced in the prototype with the $1 / 0$ switch loading the phase-shift oscillator and I can therefore see no point in altering the $1 / 0$ switch if the rest of the circuit is built as published (with the recommended alterations). If Mr. Naaijer's suggestion is adopted it would probably be necessary to redesign the phase-shift oscillator to run on 6 V .

Most of the component savings claimed mean putting up with a high output impedance with the attendant hazard of hum pick-up. Mr. Naaijer's reference to two, instead of four, transistors refers to using the unused exclusive-OR gates as a multivibrator. (This was described 'in A. W. Critchley's letter already published and, therefore, the print was removed from Mr. Naaijer's letter to avoid duplication.) Brian Crank

## Stereo mixer

For readers who wish to build the designs published in the May and June issues, here are some details of suitable components. Capacitors used in equalization, tonecontrol and filter networks should be 5\% components, polystyrene types for values less than $0.01 \mu \mathrm{~F}$, and polycarbonate (e.g. Siemens B32540) above $0.01 \mu \mathrm{~F}$. The 4.7 pF high-frequency compensation capacitors, connected from collector to base of the second transistor in Figs. 3, 8 (a) and (b) are not critical and could be
increased to 10 pF so that polystyrene types can be used. Electrolytic capacitors are from the Mullard C426 and C437 ranges, and non-polarized coupling capacitors are from the Mullard C280 range. Fixed resistors are $5 \% \frac{1}{4} \mathrm{~W}$ carbon film, unless stated otherwise.

The apparently blank statements cocerning residual noise and mixing level made in part 1 (May issue) require explanation as this point was given theoretical treatment in an unpublished part of the manuscript. A noise analysis of the virtualearth mixer Fig. 10, shows the signal to residual noise to be $v_{i} \sqrt{4 . k . T . \Delta f . R . n}$ where $v_{i}$ is the maximum nominal signal at the slider of the channel fader, $R$ is the resistance level of the mixer (i.e. the value of the channel fader or summing resistor) and $n$ is the number of channels. As the maximum output of the pre-mixing circuits is between 8 and 9 V r.m.s. an overload margin of 30 dB requires $v_{i}$ to be around 120 mV after allowing for a 6 dB loss in the channel balance control. If $R$ is $20 \mathrm{k} \Omega$ and $n$ is 5 , then the residual noise level is -84.5 dB on a 30 kHz noise bandwidth. The expression indicates that the residual noise level deteriorates as the number of channels is increased but is improved by a reduction in the resistance level $R$, and by an increase in the signal level at mixing. Both the latter effects also reduce the overload margin, so a compromise has to be found. Alternatively the preset sensitivity control can be moved, for example to the feedback loop, though this presents its own problems of stability.
Hugh Walker,
South Queensferry,
Scotland.

## F.M. stereo tuner

I have found that there have been a small number of tuners produced to my design* which have given signs of instability, and I have been able to reproduce this effect in my own tuners. The trouble is not instability in the normal sense, but gives the impression that it is. The trouble is 'squegging' of the local oscillator, and the cure is the standard one-reduce the base time-constant. I have found that the base capacitor, now 47 pF , is best reduced to 15 or 22 pF , which cures the problem; the only side effect being due to the slight lessening of oscillator amplitude, with a slight reduction in sensitivity. This is of little consequence because of the very high sensitivity and is largely offset by a slight reduction in background noise. After changing the base capacitor to 15 pF in two tuners both of which exhibited the apparent instability, there was no trace of any effects nor could they be provoked by any setting of the tuning or trimming controls. In both tuners the background between stations was very quiet despite a sensitivity for 3 dB limiting below $1 \mu \mathrm{~V}$. L. NELSON-JONES,

## Bournemouth,

Hants.

[^4]
# Phase-locked-loop Stereo Decoder I.C. 

## Build a high-performance decoder with the minimum number of components

It is possible to make a high-performance phase-locked-loop stereo decoder with just sixteen components and a printed circuit board. Only one coil is required and only one adjustment is necessary. The major ccmponent in the decoder is an integrated circuit (CA3090Q), containing 126 transistors, which has just been introduced by R.C.A.

A block diagram of the i.c. is given in Fig. 1. The composite output signal from the discriminator of an f.m. receiver is applied to pir 1 of the i.c. where it is amplified for distribution to other parts of the chip. The phase-locked-loop consists of a voltage controlled oscillator (v.c.o.), two divide-by-two stages and a phase comparator (phase-lock detector). An inductor and a capacitor connected to pins 15 and 16 give the v.c.o. a natural centre frequency of 76 kHz . This 76 kHz signal is divided by four in two cascaded divide-by-two stages to provide a 19 kHz
reference for the phase-lock detector. The phase-lock detector compares the locally generated 19 kHz signal with the incoming 19 kHz pilot tone and provides an output to alter the operating frequency of the v.c.o. if there is any difference. The bandwidth of this loop-which may be likened to a servo system-is determined by an $R C$ network coṇnected to pin 14.
The whole purpose of the loop is to regenerate the 38 kHz sub-carrier which is suppressed at the transmitter before the signal is transmitted. The 38 kHz subcarrier is necessary to demodulate the composite stereo signal and the action of the loop ensures that the regenerated sub-carrier is very closely related in phase to the transmitted 19 kHz pilot tone.

When the v.c.o. is running at exactly the right frequency the output from the phase-lock detector is zero so it is necessary to provide a second detector, to sense the presence of the pilot tone, in order

(A) Composite signal (B) stereo enable signal (C) stereo gating signal (D) Difference signal

Fig. 1. Block diagram of the CA3090 integrated circuit which forms the major part of a phase-locked-lcop stereo decoder.
that the chip can distinguish between a stereo and a mono signal-the pilot tone is not present on a mono signal.

This detector is called the pilot presence detector and it is driven by a second divide-by-two stage operating from the chip's 38 kHz line. The resulting 19 kHz signal is compared with the composite input signal and if a pilot tone is present the pilot presence detector trips a Schmitt trigger. The sensitivity of the pilot presence detector is set by a resistor connected between pins 7 and 8 . With the value shown in Fig. 2, a 4 mV input signal (pin 1) will be sufficient to operate the Schmitt trigger. If greater sensitivity is required the resistor can be replaced with a 4.7 mH coil in series with 15 nF capacitor across pins 7 and 8. The Schmitt trigger will then operate at 3.3 mV (off at 2 mV ) and an improved overload characteristic is obtained as a by-product. An $R C$ combination connected to pin 6 is a filter for the pilot presence detector.

When the Schmitt trigger operates it lights the stereo indicator lamp via an integral driver amplifier and informs the left/right channel detector that a stereo signal is being received and switches the whole chip to stereo operation:

The left/right channel detector uses the 38 kHz sub-carrier (stereo gating signal), generated by the phase-lockedloop, and the composite input signal to produce a stereo difference signal which drives the matrixing circuits. The matrix extracts the left and right channel outputs from the composite input signal in the normal way and after amplification the left and right channel outputs appear at pins 9 and 10.

## Practical notes

The complete circuit diagram is given in Fig. 2 and little need be said about it as the purposes of most of the components have already been described. The capacitors $C_{1}$ and $C_{2}$ provide the necessary deemphasis and the two $10 \mathrm{k} \Omega$ resistors are the collector loads of the 'open ended' channel amplifier output transistors.

The stereo indicator lamp can be a light-emitting diode as shown or a normal filament lamp which may be connected in place of the light emitting diode and $680 \Omega$ series resistor provided that the lamp does


Fig. 2. Additional components required to complete the decoder. For operation in the UK ( $50 \mu \mathrm{~s}$ de-emphasis) change the value of the $7.5 n S$ capacitors to $5 n S$.
not consume more than 14 mA at 12 V . If a higher current lamp is used an outboard driver transistor must be added. The inset shows circuits using either a p-n-p or an n-p-n transistor. The transistor type is not critical provided that it can handle the lamp current. For instance, a $40 \mathrm{~mA}, 12 \mathrm{~V}$, lamp could be used if it were driven by a BC 108 (use the $\mathrm{n}-\mathrm{p}-\mathrm{n}$ circuit in this case). However, the maximum lamp current-whatever the transistor used-should not exceed 100 mA because the drive is limited to 14 mA . Anyway who wants to use a searchlight to indicate that a stereo signal is being received!

The decoder can be built on the printed circuit board shown in Fig. 3 full size, or 'pin-board' construction can be employed. The 2 mH coil can be obtained from Harrogate Radio Ltd., $2 / 3$ Sykes Grove, Harrogate, Yorks., price 15 p including postage, etc. Ask for type 87 BN 135 BX 2 . The prototype used a coil of American origin. The type we have specified in fact contains two coils so for this application use coil pins 3 and 4 only. A slight alteration to the printed circuit board may be necessary. Alternatively use any 2 mH coil which allows a $\pm 25 \%$ adjustment.

When connecting the decoder to the discriminator output of a receiver care should be taken to ensure that the receiver's de-emphasis network is disconnected. The decoder will accept inputs between 40 and 400 mV . If the discriminator


Fig. 3. Prototype printed circuit board layout shown actual size.
of your receiver provides an output higher than this use a potentiometer of about 100 k to reduce the signal. Make sure your receiver has enough bandwidth for stereo operation.

Two methods may be employed to set-up the decoder both of which are extremely simple. If you have access to a digital frequency meter connect it to pin 15 of the i.c. and adjust the core of the 2 mH coil to give 76 kHz . This adjustment is done when there is no input to pin 1.

The second method of adjustment does not require the use of any test equipment. Connect the decoder to a receiver via a $100 \mathrm{k} \Omega$ potentiometer and tune in a stereo broadcast. Start with the core of the 2 mH coil fully out and the potentiometer set to give maximum input to the decoder. Screw in the core of the 2 mH coil until the stereo indicator lamp lights; continue turning the core in the same direction,
counting the turns, until the stereo indicator lamp goes out. Set the core at a point midway between the points where the lamp came on and went off.

Alter the potentiometer setting so as to reduce the input to the decoder and extinguish the stereo indicator lamp. Rock the core of the 2 mH coil about its centre position to see if the indicator lamp lights. If not, slightly increase the potentiometer setting and rock the core again. The correct position for the coil's core is the one that lights the lamp with the minimum input signal.
R.C.A. manufacture two versions of the decoder i.c. One is in a staggered 16-pin dual-in-line package which is used in the illustrated printed circuit board and is called type CA3090Q, the second-type CA 3090 E - is electrically identical and is housed in a conventional 16-pin dual-inline package. The i.c. is available from R.C.A. distributors, price £3.46.

Typical Decoder Specification

| Input impedance | $50 \mathrm{k} \Omega$ |
| :--- | :--- |
| Channel separation | 40 dB |
| Channel balance (mono) | 0.3 dB |
| Mono gain | 6 dB |
| Stereo/mono gain | 0.3 dB |
| Indicator lamp turn-on voltage | 4 mV |
| Capture range (deviation from |  |
| 76 kHz centre frequency) | $\pm 10 \%$ |
| Distortion |  |
| 2 nd harmonic | $0.35 \%$ |
| $3 \mathrm{rd}, 4 \mathrm{th}$ and 5 th harmonic | $0.1 \%$ |
| 19 kHz rejection | 35 dB |
| 38 kHz rejection | 25 dB |
| Input voltage range | 40 to |
| Supply voltage | .12 V | Supply current (lamp off) $50 \mathrm{k} \Omega$

Channel separation 0.3 dB

Mono gain 6dB
Stereo/mono gain $\quad 0.3 \mathrm{~dB}$ Indicator lamp turn-on voltage" 4 mV
Capture range (deviation from Hz centre frequency)
$\pm 10 \%$
Distortion
3rd, 4th and 5th harmonic
19 kHz rejection
$0.1 \%$
35dB
25 dB
12 V

- 22 mA

Operating temperature range $\quad-40$ to $+85^{\circ} \mathrm{C}$ - For improved pilot sensitivity and overload characteristics replace the $150 \Omega$ resistor between pins 7 and 8 with a coil of 4.7 mH in series with a capacitor of $0.015 \mu \mathrm{~F}$.


Fig. 4. Photograph of the prototype. Because this is a demonstration model built by R.C.A. some of the components shown in Fig. 2 are not included.

## Ceramic Pickup Equalization

## 2-Practical low-impedance circuits

by B. J. C. Burrows, B.Sc.

This article gives full circuit details of an economy and a high-performance preamplifier which use a new design principle to provide optimum performance from stereo and mono ceramic cartridges.
Many ceramic cartridges are capable of a very high standard of performance-but this is seldom realized in practice. This is because conventional pre-amplifiers cannot cope satisfactorily with the wide range of electrical parameters encountered in different makes of ceramic cartridge.

The two factors that cause the problems in pre-amplifiers for piezo-electric cartridges are (i), self capacitance, and (ii), the degree of built-in mechanical equalization. In conventionally designed circuits using high-value load resistances ( $1-2 \mathrm{MS}$ ), the pickup self-capacitance has a profound effect on low-frequency performance and hence on the rumble performance. Fig. 1 shows curves of output voltage against frequency for two well known pickups when operated into a conventional preamplifier with $2 \mathrm{M} \Omega$ input impedance. These show that the overall frequency response is far fromflat.

Typical pickups vary in capacitance from 200 pF to greater than 1500 pF , and with manufacturing tolerances plus the uncertain nature of the lead capacitance an overall variation of 180 pF to $>2000 \mathrm{pF}$ is possible. To obtain good l.f. performance with 180 pF needs a loading resistance of $18 \mathrm{M} \Omega$ (not $1-\mathrm{M} \Omega$ as commonly provided). If $18 \mathrm{M} \Omega$ were used with a pickup of 2000 pF the bass turnover frequency would be 4.5 Hz ! This of course would result in very objectionable rumble and l.f.
(a) 9TAHC into $2 \mathrm{M} \Omega$ and 100 pF load
(b) scul into $2 M \Omega$ load


Fig. 1. Voltage/frequency curves of two well-known ceramic cartridges when used with a conventionally-designed pre-amp , with $R_{\text {in }}=2 M \Omega$, and a flat frequency response.
arm resonance $\dagger$ problems.
Conventional pre-amplifier designs do not allow for built-in mechanical equalization which varies from one pickup to another, and unfortunately the usual type of tone controls are not suitable for providing the necessary correction.

We can draw up a list of performance characteristics which an ideal pre-amplifier should possess:
(1) l.f. performance independent of cartridge capacitance;
(2) accurate rumble filtering independent of cartridge capacitance;
(3) means of correcting for variability in mechanical equalization (i.e. some form of 'tone balance' control).
(4) ability to cope with pickups of widely differing output voltages.
To these may be added: low noise, low distortion, good overload capability, builtin tone controls, etc.
Economy pre-amplifier
The complete circuit of the economy design is given in Fig. 2 for a positive h.t.
$\dagger$ See Appendix II.


Table of values for $C_{1}, C_{2} \& R_{1}$ in economy circuit.

| Cartridge type | $C_{1}$ | $C_{2}$ | $R_{1}$ <br> (optimum value) |
| :--- | :---: | :---: | :---: |
| Decca Deram    <br> $\left.\begin{array}{l}\text { Goldring CS91E }\end{array}\right\}$ $3.3 n \mathrm{nF}$ $0.1 \mu \mathrm{~F}$ $18-27 \mathrm{k} \Omega$ <br> Goldring CS90 <br> $\left.\begin{array}{l}\text { Sonotone 9TAHC } \\ \text { Connoisseur SCU1 }\end{array}\right\}$ 3.3 nF $0.1 \mu \mathrm{~F}$ $56 \mathrm{k} \Omega$ <br> $\left.\begin{array}{l}\text { B.S.R.SC5M } \\ \text { Acos GP94/1 } \\ \text { Garrard KS40A }\end{array}\right\}$ $3.3 n \mathrm{~F}$ $0.1 \mu \mathrm{~F}$ $26 \mathrm{k} \Omega$ | $10 \mathrm{kF} \Omega$ | low output |  |



$$
\text { It } \begin{aligned}
\mathrm{R}_{B} \times \mathrm{C}_{\mathrm{B}}= & 318 \mu \mathrm{se} \text { then for a flat } \\
& \text { }
\end{aligned}
$$

Fig. 3. First-stage design of equalization circuit.


Fig. 4 Operation of tone-balance control, $R_{A}$ in Fig. 3.


$$
\begin{aligned}
& \text { (1) Chosign formulae for } Q=1 \\
& \text { (2) Make } R_{C} \text { several times } R_{C} \\
& \text { (3) } C_{B}=\frac{1}{2 \pi f_{1} R_{B}} \\
& \text { (4) } C_{C}=\frac{1}{2 \pi R_{C}}\left(\frac{1}{f_{0}}-\frac{1}{f_{1}}\right) \\
& \text { (5) } R_{D}=R_{B}\left(\frac{\left(C_{C} R_{C}+C_{B} R_{B}\right)^{2}}{C_{C} R_{C} C_{B} R_{B}}-1\right)
\end{aligned}
$$

Fig. 5. Baxandall bass lift-and-cut circuit.
rail system. A negative h.t. rail version is given in Appendix I. For normal use connect $A$ to $A^{\prime}$ and $B$ to $B^{\prime}$ and use full circuit. For ultra-economy operation with any of the pickups except the Deram or CS91E, the second stage may be omitted by connecting $A$ direct to $B^{\prime}$ and omitting the intervening circuitry associated with $\operatorname{Tr}_{2}$. Thus a very good, yet simple, gramophone amplifier may be built by using only $T r_{1}$ and $T r_{3}$ directly connected into an amplifier with 100 mV sensitivity for full output.

## Design principles of equalization stage

Last month the merits of the shunt feedback (or virtual earth) amplifier were mentioned as being very suitable for ceramic pickup equalization. Further, it was shown that loading the pickup with a low impedance had no effect on its internal e.m.f. In the present design, then, the effects of the variability in capacitance have been eliminated by swamping the pickup in every case with a shunting capacitor of 3.3 nF or more. An input resistor of $75 \mathrm{k} \Omega$ then gives an input time constant of $318 \mu \mathrm{~s}$ (equivalent to 500 Hz ); to match this, the feedback circuit has a time constant of $318 \mu \mathrm{~s}$ also (see Fig. 3); the complete circuit has a flat frequency response:

$$
\frac{V_{O}}{E}=\text { constant }=\frac{R_{B}}{R_{A}}=\frac{C_{P}+C_{A}}{C_{B}}
$$

If any one of the components suffixed $A$ or $B$ is made variable, a 'tone balance' type of control is achieved in a much simpler manner than circuits described previously ${ }^{1}$. $R_{A}$ is the best one to vary and provides
performance variation as in Fig. 4. The value of $R_{A}$ to give an overall flat frequency response is termed $R_{0}$. In practice only values of $R_{A}$ between $R_{0}$ and $R_{0} / 4$ are needed to fully correct all ceramic pickups for their lack of complete mechanical equalization, e.g. the Sonotone 9TAHC pickup needs $R_{A}=R_{0} / 1.8$ and the Connoisseur SCUI needs $R_{A}=R_{0} / 4$.

With an infinite gain amplier in Fig. 3, overall gain is flat down to d.c. theoretically. This is no use in audio work because of rumble and the l.f. arm resonance. Some form of rumble filtering is essential and may be built into the equalization stage by using the circuit due to P. J. Baxandall ${ }^{2}$. The essence of this circuit is in Fig. 5, and its performance in Fig. 6.



Fig. 6. Performance of circuit of Fig. 5 with $f_{0}=50 \mathrm{~Hz}$ and $f_{1}=500 \mathrm{~Hz}$.


If a further high-pass $R C$ filter is added,

$$
f_{0}=\frac{1}{2 \pi R_{C}}
$$

where a flat response to nearly 50 Hz is achieved with a rapid turnover to a slope of $18 \mathrm{~dB} /$ octave to attenuate rumble. Finally, with $R_{A}$ adjustable, the tone balance facility is still retained as with the basic circuit of Fig. 3. It is common to design rumble filters with cut-cff frequencies much lower than 50 Hz ; but, to achieve adequate attenuation at $25 \mathrm{~Hz}-\mathrm{a}$ common frequency of the l.f. arm resonance-a high value of $f_{0}$ is required. The actual circuit of Fig. 2 achieves -28 dE at 15 Hz and -15 dB at 25 Hz . In practice this is very satisfactory.

The economy-design pre-amplifier closely matches the theoretical performance of Figs. 4 and 6 and provides excellent bass, good balance and excellent freedom from rumble. As shown in the table relating to the main circuit, the only circuit changes needed to accommodate different pickups are for surbing those with a very high output voltage with a capacitive divider. In connection with the table of values given for the input capacitors it is very important to stress that the values given must be used as specified and that the manufacturers' recommendations regarding load impedance and equalization must be totally ignored. This circuit has been specifically designed to take care of all the loading, matching and equalization factors and no further components are needed.

| High-performance pre-amplifier specification |  |
| :---: | :---: |
| rated outpat | 500 mV r.m.s. |
| harmonic distortion | $0.02 \%$ at rated output |
|  | -60 dB all inputs <br> -80 dB for tuner and aux |
|  | inputs |
| hum | negligible with good layout |
| overload capacity | 23dB over whole audio range, infinite for tuner and |
|  |  |
| sensitivity | tuner 250 mV |
|  | aux 250 mV |
|  | disc magnetic 3 mV |
|  | disc ceramic 20 mV |
|  | tape 4 mV |
|  | mic 10 mV |
| input impedance | tuner, aux $60-100 \mathrm{~K} \Omega$ disc magnetic $47 \mathrm{~K} \Omega$ |
|  | disc ceramic frequency |
|  | dependent |
|  | tape, mic $47 \mathrm{~K} \Omega$ |
| disc equalization | magnetic--RIAA to within |
|  | $\pm 1 \mathrm{~dB}$ ceramic-can be |
|  | $\begin{array}{lcll} \text { adjusted } & \text { to } & \text { give } & \text { fat } \\ \text { response } & \pm \frac{1}{2} \text { dB } & \text { I.f. } \end{array}$ |
|  | response independent of |
|  | cartridge capacitance |
| tape equalization | $7 \frac{1}{2} \mathrm{i}$. p.s. with $R_{F B}=39 \mathrm{~K} \Omega$ |
|  | 1 5i.p.s. with $R_{F B}=18 \mathrm{~K} \Omega$ |
|  | 3亲i.p.s. with $R_{F B}=82 \mathrm{~K} \Omega$ |
| rumble filter | modified design giving |
|  | higher cut off frequency: response at 25 Hz is |
|  | $-15 \mathrm{~dB}$ |
| low-pass filter | switched, flat or cut off at any frequency from 4 to |
|  | 11 KHz (see Ref. 7) |
| tone controls | Baxandall type |
|  | treble $\pm 16 \mathrm{~dB}$ at extreme |
|  | bass $\pm 20 \mathrm{~dB}$ at extreme |
| current consumption | 7 mA |



Fig. 8. Measured voltage/frequency curve for a 9TAHC operating into an 'economy design' circuit with $R_{A}=R_{0} / 1.8$. The curve for the SCU1 would be just as flat, but with $R_{4}=R_{0} / 4$.

The economy circuit as described fulfils all the design criteria enumerated earlier except for the slight inconvenience of changing two capacitors if pickups of widely differing output voltages are exchanged. The noise performance is very good with all the cartridges listed apart from two (the CS91E and Deram) with which it is satisfactory for everything but the most exacting requirements.

## High performance pre-amplifier

This is based on the Bailey ${ }^{3}$ design of 1966 but with all the subsequent modifications to improve the filter ${ }^{4}$ and tone control ${ }^{5}$ circuits, plus the addition of a complete ceramic-pickup equalizing circuit achieving the same performance with ceramic cartridges as the economy pre-amplifier. The complete circuit is given in Fig. 7, which also incorporates one further modification to raise the cut-off frequency of the rumble filter in accordance with the design philosophy discussed in Appendix II. Equalization for magnetic pickups has been retained and is selected by the input selector switch. The 'set level' control needs a mention. To avoid overloading the input stage, adjust the set level control with any particular


Fig. 9. Economy circuit arranged for negative h.t. rail. For values of $C_{1}, C_{2}$, and $R_{1}$, see table earlier.
cartridge to give comfortable listening level with the main volume control at about half of its maximum rotation. This control need be only a preset with screwdriver slot adjustable from the back of the preamplifier. The tone balance could be the same, or it could be brought out as a front panel control, or as a skeleton pot mounted internally or even a 'select-on-test' fixed resistor.

On paper, the specification of the high performance pre-amplifier looks most impressive, but subjectively the economy version is very good indeed, and both represent a considerable improvement on conventional designs in that reproducible low-frequency performance, effective rumble filtering independent of pickup capacitance, and a simple means of correcting for partial mechanical equalization have been incorporated. Fig. 8 in conjunction with Fig. 1 gives a comparison of the performance of the Sonotone 9TAHC and Connoisseur SCU1 using conventional loading ( $2 \mathrm{M} \Omega$ plus flat amplifier), compared with the measured results on the author's 9 TAHC using the economy circuit.
The calculated performance of the Connoisseur SCU1 with $R_{A}=R_{0} / 4$ is a straight line coincident with the 0 dB line on Fig. 8, although in practice there would be a variation of up to $\pm 1 \mathrm{~dB}$ about the 0 dB line.

Modifications to provide a similar standard of performance with the Dinsdale Mark I and Mark II pre-amplifier circuits were incorporated in a previous article ${ }^{6}$.

## Appendix I

Alteration of economy circuit for negative h.t. rail operation, e.g. from a germaniumtransistor amplifier like the Dinsdale Mark I or II, is basically to return all elec-
trolytic capacitors to the positive potential rail, viz. the earth line (see Fig. 9). There are no modifications to circuit values apart from the voltage rating of the electrolytics.

## Appendix II

Arm resonance (l.f.) is the tendency toward damped oscillation at a low frequency and is exhibited by most pickup arms. It has the effect of greatly increasing the cartridge output voltage at or near the resonant frequency. The frequency, $f_{l j}$, is normally in the range $10-25 \mathrm{~Hz}$, so its effect is to greatly increase rumble. The frequency of the oscillation is:

$$
f_{l f}=\frac{1}{2 \pi \sqrt{\mathrm{MC}}} \mathrm{~Hz}
$$

$M$ is the mass of cartridge plus effective mass of arm measured at cartridge.
$C$ is the compliance of stylus cantilever suspension. With $M$ in grams, $C$ is in $\mathrm{cm} /$ dyne.
With modern high compliance cartridges it is desirable to keep $M$ very low-hence lightweight headshells-to make $f_{l /}$ as high as possible. Generally speaking the lower the frequency of resonance the higher the $Q$, and vice versa. But a higher resonant frequency is more trouble electrically. A low-frequency high- $Q$ resonance causes mechanical difficulties-the pickup tends to leave the record surface when excited. A resonance at 25 Hz is acceptable mechanically if the $Q$ is low enough and its electrical effects can be removed with a steep slope filter. Below this resonant frequency the cartridge output voltage falls off very sharply indeed ( 24 dB /octave) thus providing the required severe attenuation of sub-audio frequencies.

With regard to pre-amplifier design, the point to note is that the highest amplitude rumble components will be at, or near, the
1.f. arm resonance. A filter in the preamplifier should ideally provide 12 dB or more of attenuation at 25 Hz , yet not interfere with I.f. audio response. A cut off frequency of 50 Hz with slope approaching $18 \mathrm{~dB} /$ octave is a very good compromise since it causes very little error inthe R.I.A.A. equalization, yet gives -15 dB at 25 Hz and -25 dB at 15 Hz .

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# The Diagnosis of Logical Faults 

## Conclusion

by R. G. Bennetts*, B.Sc., M.Sc.

One of the problems that the designer and user of logical systems is confronted with is that of testing the logical functioning of the circuits within the system. The procedure is usually split into two main processes-namely a simple go/no go test followed by, in the event of a no go decision, a more thorough analysis to determine the location of the fault. The former is known as fault detection whereas the full detection and location process is termed diagnosis. It is the purpose of this series of two articles to illustrate, through the use of examples, some of the techniques that have been developed to assist in determining the necessary tests and to comment on their advantages and disadvantages. The first part of this article appeared last month and concludes this month with a discussion of Boolean difference and partitioning techniques.

## 3: Boolean difference

Before describing how the Boolean difference can be used to determine a detection test set. it is instructive to define the term "Boolean difference" and show how it may be derived.
Consider a Boolean function $z$ given by:

$$
\begin{gathered}
z=f\left(x_{1} x_{2} \ldots x_{i} \ldots x_{n}\right) \\
x_{1} \rightarrow x_{n}=\text { primary inputs }
\end{gathered}
$$

If $x_{i}$ is in error, then a new function $z_{x_{i}}$ is defined by:

$$
z_{x_{i}}=g\left(x_{1} x_{2} \ldots \bar{x}_{i} \ldots x_{n}\right)
$$

i.e., $z_{x_{i}}$ is formed by replacing $x_{i}\left(\bar{x}_{i}\right)$ in $z$ with $\bar{x}_{i}\left(x_{i}\right)$. The Bookean difference.

$$
\frac{d z}{d x_{i}}
$$

is defined:

$$
\begin{aligned}
\frac{d z}{d x_{i}} & =Z \oplus Z_{x_{i}} \\
& =h\left(x_{1} x_{2} \ldots x_{n}\right)
\end{aligned}
$$

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where $\oplus$ denotes the Boolean exclusive-OR operator.

As an example, we will derive the Boolean difference expression for the example circuit with primary input " $C_{3}$ " as " $x_{i}$ ". (This was given in Fig. 4 last month and is repeated here.)

From Fig. 4,

$$
\begin{aligned}
z= & \bar{a} b+\bar{a} \bar{c}+a b \\
= & \bar{C}_{1} \bar{C}_{2}+\bar{C}_{1} \bar{C}_{3}+C_{1} C_{2} \\
Z_{C 3}= & \bar{C}_{1} \bar{C}_{2}+\overline{C_{1}} C_{3}+C_{1} C_{2} \\
\frac{d z}{d C_{3}}= & \left(\bar{C}_{1} \bar{C}_{2}+\bar{C}_{1} \bar{C}_{3}+C_{1} C_{2}\right) \oplus \\
& \left.\bar{C}_{1} \bar{C}_{2}+\bar{C}_{1} C_{3}+C_{1} C_{2}\right)
\end{aligned}
$$

There are mathematical rules for manipulating such expressions but for a small number of input variables, the Karnaugh map (K-map) can be used quite easily and also serves toillustrate very clearly the actual exclusive-OR operation. The procedure is to map $Z$ into one K -map, $Z_{c 3}$ into another and by comparing similar locations to derive the mapping of $\frac{d z}{d C_{3}}$ by inserting a 1 if there is a difference in the values at the two locations, otherwise blank. The method is illustrated in Fig. 8.
Returning to the theory, let us examine the significance of the Boolean difference expression. If there is a fault in the value of
$x_{i}$, then the function that the faulty network will realize will be that defined by $Z_{x_{i}}$. Under these conditions, the faulty output will differ from the true output only for those terms that make $\frac{d z}{d x_{i}}=1$, i.e. $\frac{d z}{d x_{i}}$ defines the full set of inputs (tests) that will cause an incorrect and hence observable output if there is a fault in the logical value of $x_{i}$. Note that so far we have not defined whether $x_{i}$ is $\mathrm{s}-\mathrm{a}-1$ or $\mathrm{s}-\mathrm{a}-0$ - only that it is logically incorrect. It therefore remains to partition the set of tests defined by $\frac{d z}{d x_{i}}$ into those pertaining to $x_{i} \mathrm{~s}-\mathrm{a}-1$ and $x_{i} \mathrm{~s}-\mathrm{a}-0$. This is achieved by splitting the list of all tests into those containing $x_{i}$ and those containing $\bar{x}_{i}$. The former will demand a 1 on $x_{i}$ and therefore test for $x_{i} \mathrm{~s}-\mathrm{a}-0$ and the latter conversely will test for $x_{i} \mathrm{~s}-\mathrm{a}-1$.

Thus, for $\frac{d z}{d C_{3}}$ in Fig. 8 :

$$
\frac{d z}{d C_{3}}=\overline{C_{1}} C_{2} C_{3}+\overline{C_{1}} C_{2} \overline{C_{3}}
$$

and the $\overline{C_{1}} C_{2} C_{3}\left(t_{3}\right)$ term defines the test for $C_{3} / 0\left(f_{5}\right)$ and $\overline{C_{1}} C_{2} \overline{C_{3}}\left(t_{2}\right)$ defines the test $C_{3} / 1\left(f_{6}\right)$. These can be confirmed from the detection matrix $G_{D}$ of Fig. 6 (last month). Note that for each fault, there is only one test and hence $t_{2}$ and $t_{3}$ are both essential.
As another example, we will consider how the Boolean difference can be used to determine the tests for a fault on one of the lines that is not a primary input. $C_{4}$ say.
As above, we have:

$$
Z=\overline{C_{1}} \bar{C}_{2}+\overline{C_{1}} \bar{C}_{3}+C_{1} C_{2}
$$

and $C_{4}=C_{1}+C_{2}$

$$
=\overline{\overline{C_{1}} \overline{C_{2}}} \text { (by De Morgan's theorem) }
$$

by substitution $Z=\overline{C_{4}}+\overline{C_{1}} \overline{C_{3}}+C_{1} C_{2}$


Fig. 4. The circuit example; reproduced from last month's issue.

Fig. 8. Karnaugh maps for deriving $d z / d C_{3}$.

$$
\text { and } Z_{C 4^{\prime}}=C^{4}+\overline{C_{1}} \overline{C_{3}}+C_{1} C_{2}
$$

By using four variable K-maps, the Boolean difference $\frac{d z}{d C_{4}}$ is found to be given by: $\frac{d z}{d C_{4}}=C_{1} \overline{C_{2}} C_{3}+C_{1} \bar{C}_{2} \overline{C_{3}}+\overline{C_{1}} C_{2} C_{3}+$

$$
\overline{C_{1}} \overline{C_{2}} C_{3}
$$

Now, since $C_{4}=C_{1}+C_{2}$, the only time it will be 0 will be when both $C_{1}$ and $C_{2}$ are 0 . Thus in order to detect for $C_{4} / 1$, the input must contain the terms $\bar{C}_{1} \bar{C}_{2}$. All other combinations of $C_{1} C_{2}$ will detect $C_{4} / 0$. From this we see that only $\overline{C_{1}} \overline{C_{2}} C_{3}\left(t_{1}\right)$ will detect $C_{4} / 1$ whereas $C_{1} \overline{C_{2}} C_{3}\left(t_{5}\right), C_{1} \overline{C_{2}} \overline{C_{3}}\left(t_{4}\right)$ or $\overline{C_{1}} C_{2} C_{3}\left(t_{3}\right)$ will serve for $C_{4} / 0$, and again the fault matrix $G_{D}$ confirms this.

The Boolean difference tends to be limited to circuits having a relatively small number of input variables, but it can be expressed as a fairly rigid algorithm and would seem quite suitable for implementation in a computer program. Its main advantage is in spotting essential tests and once these are known, the path sensitizing procedure (discussed last month) for evaluating all other faults detected by that test can be used. Using these two techniques together can result in an efficient procedure for deriving an optimal test sequence.

At present, the technique is restricted to combinational networks, but successful excursions into the area of sequential net works have been reported though this aspect is still very much in its infancy.

## 4: Partitioning

As has been indicated previously, the partitioning technique is more applicable to multi-flow testing procedures and this calls for certain criteria to be used. Before considering these criteria in detail, let us consider the basic technique itself.

The circuit under test is usually simulated in order to arrive at the test set for detection and/or location and the simulated model can be converted from its no-fault version $f_{0}$ to any of $n$ previously defined faulty versions $f_{1} \rightarrow f_{n}$. (In the case of our example, $\left.f_{1} \rightarrow f_{16}\right)$. A test is then applied to all versions of the circuit and this will effect a partition based on the value at the output. The members of each equivalence class, as it is called, indicate that the output is the same and further tests are required to increase the degree of resolution until either $f_{0}$ is identified alone (fault detection) or all versions are işolated (fault location).

The value of this procedure lies in its ability to try different tests and ascertain which one is best for the job in hand. This implies the use of criteria and we will consider initially the use of the checkout criterion for fault detection only. Again, we will illustrate this through use of the circuit example.

Fault detection using the checkout criterion: The initial equivalence class for the example circuit is $f_{0} \rightarrow f_{16}$ inclusive and we require to isolate $f_{0}$ as quickly as possible by means of a set of test inputs. This amounts to determining which test separates the largest

| Test | No. of foulty circuits detected |  |  |  |
| :---: | :---: | :---: | :---: | :---: |
|  | $N_{1}$ | $N_{2}$ | $N_{3}$ | $N_{4}$ |
| $t_{0}$ | 3 | 2 | 0 | - |
| $t_{1}$ | 5 | $(4)$ | - | - |
| $t_{2}$ | 4 | 4 | $(2)$ | - |
| $t_{3}$ | $(8)$ | - | - | - |
| $t_{4}$ | 7 | 2 | 1 | 1 |
| $t_{5}$ | 7 | 2 | 1 | 1 |
| $t_{6}$ | 3 | 2 | 1 | 1 |
| $t_{7}$ | 4 | 3 | 2 | $(2)$ |

Fig. 9. Assignment of checkout weighting and selection of best tests.
number of faulty circuits from the good circuit-this being the checkout criterion. If we look at the detection matrix $G_{D}$ of Fig. 6 (last month) we can list the number of detectable circuits for each test and this is shown in column $N_{1}$ of Fig. 9.

Obviously, $t_{3}$ is first choice and this will create a partition $P_{1}$ defined by the two equivalence classes $P_{1}{ }^{1}$ and $P_{1}{ }^{2}$ where:

$$
\begin{aligned}
& P_{1}^{1}=\left\{f_{0} f_{1} f_{4} f_{6} f_{8} f_{10} f_{11} f_{13} f_{15}\right\} \\
& P_{1}^{2}=\left\{f_{2} f_{3} f_{5} f_{7} f_{9} f_{12} f_{14} f_{16}\right\}
\end{aligned}
$$

The exercise must now be repeated on the equivalence class containing $f_{0}$ and the test weightings are shown in column $N_{2}$ of Fig. 9. This can be derived from the detection matrix $G_{D}$ by removing all those columns in $P_{1}^{2}$ and then counting the number of detectable faults on the remaining columns. When this is completed, there is a choice between $t_{1}$ and $t_{2}$ and we shall arbitrarily choose $t_{1}$. This creates the partition $P_{2}$ given by:

$$
\begin{aligned}
& P_{2}^{1}=\left\{f_{0} f_{1} f_{6} f_{10} f_{11}\right\} \\
& P_{2}^{2}=\left\{f_{4} f_{8} f_{13} f_{15}\right\}
\end{aligned}
$$

The procedure is again repeated until eventually at partition $P_{i}$ (in this case $i=4$ ), $f_{0}$ is isolated from all other versions and the full detection set can be defined. The remainder of the calculation are shown in columns $N_{3}$ and $N_{4}$ and the partition sequence is shown pictorially in Fig. 10.

Fault location using the information gain or distinguishability criteria: The prime object for fault location is to continue partitioning of every equivalence class until each version $f_{0} \rightarrow f_{16}$ has been completely isolated as
far as possible (obviously indistinguishable fault-sets are not subject to any further partitioning). To assist this process, two criteria have been proposed-information gain and distinguishability.

The information gain criterion is similar in concept to the entropy function used in information theory. Initially there is uncertainty as to which of the $f_{0} \rightarrow f_{16}$ versions of the circuit exists and the application of a particular test will remove some of this uncertainty, i.e. will result in a gain in information. This can be expressed mathematically as a function of the particular test and again a table similar to that of Fig. 9 would be created enabling the correct test selection to be made.

The alternative criterion is the distinguishability criterion. This is derived in the following manner: for a particular equivalence class, one wishes to select the test that distinguishes between the greatest number of circuits. This amounts to determining how many pairs of circuits within the same class are distinguishable using test $t_{i}$, $0 ₹ i ₹ n$ for $n$ tests. This criterion is more applicable to multi-output circuits in which the partitioning is to some other radix rather than two (binary) and it too can be expressed mathematically. Since the example circuit has only one output, the partition is simple binary as shown in Fig. 10.

Both criteria are somewhat complex in their evaluation and the usual process is to derive the full detection partition using the relatively simple checkout criterion; determine the degree of diagnostic resolution that is already available and then use the more complex criteria to increase the resolution to its maximum. If this is applied to the partition of Fig. 10, it is found that only one further test need be specified in order to achieve maximum diagnostic resolution. The full partition is shown in Fig. 11 and the addition of $t_{4}$ enables partitioning of $\left\{f_{1}, f_{11}\right\},\left\{f_{4} f_{8}\right\}$ and $\left\{f_{5} f_{7} f_{9} f_{12} f_{14} f_{16}\right\}$. The remaining classes of $\left\{f_{7} f_{9} f_{12} f_{14} f_{16}\right\}$ and $\left\{f_{6} f_{10}\right\}$ are indistinguishable fault sets and consequently cannot be further partitioned without the use of extra access such as test points.

The sequence of test dictated by the partition is $t_{3} t_{1} t_{2} t_{7} t_{4}$ and one aspect of this approach is that not only can the fault be located by analysis of the output sequence corresponding to the test set, but that it is now possible to specify a test for a particular fault. This is a common requirement when trouble-shooting new designs.


Fig. 10. Partition showing detection test set.

## Concluding remarks

I have introduced the general problems associated with the diagnosis of faults in logical systems and described four of the techniques that have been developed to assist in determining a satisfactory diagnostic test sequence. The techniques themselves tend to be restrictive but it has been indicated how they may be combined in an attempt to broaden their overall coverage. The real problem however has been shown to be in diagnosing faults occurring in sequential circuits, and although some of the techniques can be applied, they are not really satisfactory. Other approaches are currently being studied, the most important of which is based on an analysis of the state table for a sequential circuit. (The state table is used to formally describe the behaviour of a sequential circuit-much in the same way as a truth table does for a combinational circuit. Every configuration of the sequential circuit is defined by a state variable and there are procedural techniques for deriving the actual circuit, in terms of its connections and gates, from the initial state table description).
One major advantage with state table analysis is that a check can be made on the table at the initial design phase to ascertain the diagnosability of the sequential circuit it describes and if necessary, apply modifications to make it fully diagnosable. This is a departure from previous diagnostic philosophy in that it is now possible to make the diagnosis requirement an initial design restraint and not something that is determined after the circuit has been designed. State table analysis does rely however on being able to formulate the state table for the sequential circuit and in the case of the intuitive design, this represents quite a problem. If however a switching theory approach has been adopted in designing the circuit, then the state table is already known and this in itself is sufficient justification for using switching theory in logical design.

In this paper, we have confined ourselves only to considering faults that can occur in logical circuits. The successful diagnosis of faults at full system level, a digital computer say, is a much greater problem and the "diagnosis is a design restraint" requirement becomes even more important. The current approach is to devise a hierarchical set of tests such that if an overall system fault is detected, a more detailed set of tests can be applied that will theoretically converge onto the fault. This can sometimes be somewhat haphazard and really what is required is a fundamentally new approach to the system design process such that diagnostic capability is a design parameter: not only at circuit level, but also at full system level.

One final comment. The advent of m.s.i. and l.s.i. has caused a shift in emphasis in diagnosis requirements in that in general one only requires fault location to the smallest replaceable unit and if this is a full circuit or a sub-system itself, i.e. an 1.s.i. chip, this tends to ease the locational extensions of detection techniques, such as the fault matrix, since the faults on the same chip can be grouped together and treated


Fig. 11. Partition showing detection and location test set.
"en bloc". It does however bring us back to the overall system test problems and serves to reinforce the comments about system check-out techniques.

## References

Since 1960, there has been a profusion of papers dealing with fault detection and location and the most recent bibliography ( 86 referenced papers) is including in the review ${ }^{1}$ written by myself and D. W. Lewin. This paper also summarizes the main techniques and has pertinent comments on the effect of diagnosis requirements on computer system engineering, the requirements of digital systems in terms of diagnosis and functional testing and diagnosis of 1.s.i.
We have recently seen the publication of the first book ${ }^{2}$ to be entirely devoted to this problem and this in itself is indicative of the importance that is now attached to fault diagnosis.

In terms of the actual techniques, the paper by Kautz ${ }^{3}$ is a well written and lucid account of the fault matrix approach and similar comments may be made about the paper by Sellers et al ${ }^{4}$ dealing with the Boolean difference.

The most famous implementations of path sensitization is the D-algorithm of Roth ${ }^{5}$ and its subsequent modification ${ }^{6}$. Both papers are somewhat heavy going due to the "calculus of D-cubes" that he defines and uses to implement the concept and the contents of the first paper is well covered in ${ }^{2}$. The basic D-algorithm and its extensions have been employed by IBM to prepare diagnostic routines for their System/360
range of computers.
The technique of partitioning has been programmed by Seshu ${ }^{7.8}$ and the suite of programs, known as the Sequential Analyser, has been in use for many years now.

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## Circuit Ideas

## F.E.T. voltage regulator

The regulator described here provides fairly good performance with a minimum number of components. The basic circuit is shown below (top). Any change in output voltage caused by a change in load resistance alters the gate-source voltage of the f.e.t. via $R_{1}$ and $R_{2}$. This causes a compensating change in drain current. The stabilization ratio is excellent ( $\approx 1000$ ) but the output resistance is very high $R_{O}>1 / Y_{F S}>500 \Omega$ ) and the output current is low. To overcome these defects, the lower circuit can be used. The output resistance is greatly reduced and the stabilization ratio is still high. The maximum output current is limited by the allowable dissipation in the final transistor. Resistor $R_{3}$ is chosen to produce a quiescent current of a few mA in $\operatorname{Tr}_{3}$. An experimental set-up using the values shown produced a change of less than 0.1 V when the load current was altered from 0 to 60 mA at 5 V output. The effect of temperature on the output voltage has not been investigated but it could probably be minimized by appropriate choice of the drain current of the f.e.t.
C. R.MASSON,

Edinburgh.

transistor to detector. The effective loading imposed by the detector on the resonator (which should be six times the impedance at the 'ring') can be taken as one quarter of the net d.c. load resistance.
G. W. Short,

South Croydon.

## D.C. motor controller

Fine control of a d.c. motor can be obtained using an op-amp and a tachogenerator. The op-amp is used as a voltage sensitive switch. In the circuit shown, when the output of the generator is less than the preset reference voltage the switching transistor will bottom and full power will be delivered to the motor. Switching will take place within one or two millivolts


## Reversed operation of 'Transfilter'

Piezo-electric overtone resonators (e.g. Brush Clevite "Transfilters") are normally used as interstage couplings ini.f. amplifiers. where the requirement is to match the relatively high output impedance at the collector of one stage to the relatively low base input impedance of the next. This is accomplished by connecting the 'dot' of the resonator to the collector and the 'ring' to the base. In the final i.f. stage shown above the impedances run the other way, and the resonator is used 'backwards' to couple the transistor to a high-impedance detector. This arrangement gives a useful voltage step-up (about 2.5 times) from
of the reference voltage. A dual power supply is required, but need only be zener stabilized. This system allows for infinitely variable drive without mechanical complication. For a record deck, for example, the speeds can be set by the simple switching of a voltage divider. The op-amp switches to within a volt or two of the supply rails, and by using a double emitter follower large motors can be controlled. The reference voltage may be set by thermistors, light-dependent resistors etc. The experimental arrangement shown used an RCA 3047A op-amp, and a 0.25 W 6 V motor as generator giving about 4 V at 13000 r.p.m.
N. G. A. Boreham, Newton Abbot, Devon.


## Electro-optical Gearbox

## using moire fringe technique

by J. Dinsdale*, м.A.

Mechanical gearboxes are generally used either for transmitting rotary power from one shaft to another, where the emphasis is on the torque ratio, or for controlling the angular velocity of one shaft with respect to another, where the emphasis is on the velocity ratio. In both cases the performance of practical gearboxes falls short of the ideal due to variable friction losses, backlash, and non-uniform velocity transmission caused by errors in the form and pitch of individual gears.

Backlash and friction effects are to some extent interdependent; in general, attempts to reduce backlash generally lead to significant increases in friction losses and the degree of backlash may be critically dependent on the working temperature of the gearbox, itself a function of friction losses.

The accuracy with which motion may be tranismitted clearly depends on the precision of the form of each tooth of the gears within the gearbox. A continuous linear transmission is desirable first to maintain linearity of motion of the member being controlled by the gearbox, and second, to minimize vibrations which can be set up by a non-linear transmission. The high-frequency vibrations caused by a typical geartrain can lead to rapid deterioration of bearings and, more seriously, to the early onset of fatigue failure. This latter effect is of particular significance in aero engines, a and much work has been devoted in recent years to improving the accuracy of gears used in aero engines and machine tools.

In the light of these deficiencies, the design of a "gearless" transmission system for controlling the angular velocity of a shaft with respect to another was investigated, the system to possess the following properties

- variable speed ratio from 1:999 to 999:1, numerator and denominator to be integral
- minimum backlash-less than 20 arc seconds over full working range
- input shaft speed range from zero to $2000 \mathrm{rev} / \mathrm{min}$ (nominal)
- output shaft speed range from zero to $200 \mathrm{rev} / \mathrm{min}$ (nominal)
- bi-directional motion
- output shaft power to be $\frac{1}{4}$ h.p.

[^5](nominal) approx. 200 watts

- transmission errors not to exceed 20 arc seconds at any speed or load up to the specified maxima.
The system $\dagger$ developed consists essentially of two shafts: an input or driven shaft, and an output or driver shaft on which is mounted an electric motor (Fig. 1). Both of the shafts are fitted with incremental shaft encoders of very high resolution. Each encoder consists of a glass disc on which has been photographed or etched a uniform pattern or grating of alternate opaque and transparent radial lines. It would not be possible to detect individual lines at this spacing by normal electro-optical means, but if such a grating is mounted in close proximity to a further small piece of similar grating (the reference grating) and the pair illuminated by white light, moire fringes appear as a series of broad light and dark bands normal to the grating lines.

The breadth and pitch of the fringes depend on the angle between the lines on the main grating and the reference. Because each moire fringe is formed over

+ Patents applied for
a relatively large area (say one sq. cm) by the integration of a large number of lines on the grating, any small pitch errors or blemishes on the grating tend to average out to give an extremely accurate fringe spacing. In fact, a grating will still give observable fringes even when $95 \%$ of its lines have been mutilated or even obliterated.

When the main grating is moved with respect to the reference, the fringes move at an equivalent rate; i.e. if the movement of the main grating with respect to the reference is at a rate of 1000 lines per second, then the fringes will move at 1000 fringes per second. The fringes are of such a size that they can easily be detected by a suitable photo-detector arrangement to give a sinusoidal signal whose frequency is proportional to the rate of angular rotation of the grating, and the number of lines on the grating. Typical gratings may have from 10,800 lines to 72,000 lines, giving angular resolutions of 2 arc minutes and 18 arc seconds respectively.

The reading head for moire fringes normally consists of a number of photo-sensitive devices and a light source


Fig. 1. In the electro-optical gearbox sinusoidal signals whose frequency is proportional to the rate of angular rotation of input and output shafts are fed to separate batch counters to set the gearbox ratio. A phase comparator provides an error signal proportional to their phase difference which controls the torque-motor driving the output shaft.
placed on either side of the small reference grating (Fig. 2). By incorporating multiple photo-sensitive devices, two signals at phase quadrature can be produced, and subsequent circuitry can determine the direction of movement of the grating.

Two diametrically opposed reading heads are normally used at each grating, and the reference and quadrature signals fed to "eccentricity logic" circuits which combine the signals in such a way as to reduce the effects of any eccentricity in mounting of the grating. In addition the signals may be interpolated electronically by a factor of up to 20 times, to increase the resolution of the system. By this means, a 72,000 -line grating can give an effective resolution of 0.9 arc second.


Fig. 2. In practice the encoder gratings are too fine to read directly, and a stationary reference grating is used to produce moire fringes which move at the same rate as the shaft but are formed over a larger area. In practice two signals-in quadrature-are needed to establish direction of rotation and to reduce the effects of any eccentricity in the mounting.

The signals from each eccentricity logic or pulse multiplier circuit are squared to give a train of pulses whose frequency is exactly proportional to the speed of rotation of the shaft, with every small fluctuation shown immediately as a corresponding variation in pulse frequency.

These pulse trains are now fed to manually set batch counting circuits, which may be arranged to give an integral batch size from 1 to 999 (or higher if need be). It is by means of these batch counters that the gearbox ratio is set, a ratio which may be altered manually at any time, even while the shafts are rotating. The outputs from the batch counters are input to a pulse-phase comparator, which
produces an error signal proportional to the instantaneous phase difference between the two pulse trains, and the phase error signal is converted to analogue form, amplified and used to feed the torque-motor driving the output shaft, thus closing the negative feedback loop.

The system is so arranged that the output shaft tries to rotate at a speed which gives pulse trains of equal frequency at the comparator. In this condition the output of the comparator appears as a square wave of unity mark/space ratio, which when integrated gives zero error. Any deviation from the correct shaft speed is detected initially as a small change in the mark/space ratio at the comparator output, equivalent to a fraction of a fringe spacing. This means that the maximum error in the transmission can be reduced to a fraction of a fringe over a speed range from zero to several hundred revolutions per minute. It must be emphasized that this is a "phase servo"-not a velocity servo.

In addition to the basic system as described some additional features ensure that the specification is maintained. Local tacho-generator feed-back around the torque motor ensures system stability at very low speeds. "Direction logic" ensures that the direction of rotation of the output shaft is always the same as that of the input shaft. (Of course, a switch can be used to reverse this direction if desired.) A counting system built into the comparator ensures that any gross errors built up during vicious acceleration and deceleration will ultimately be corrected by the system and not lost.

Velocity lag error, an inherent characteristic of position servos, is eliminated. It is explained simply by saying that if the output shaft were running at, say, $100 \mathrm{rev} / \mathrm{min}$ and providing, say, 100 watts to an external load, there will be zero signal output from the comparator when the system operates with zero error, zero current either into or out of the power amplifier and hence no power to drive the load. In other words, some inherent error must exist to drive the system. Velocity lag error is reduced by feeding forward part of the demand signal directly to the power amplifier via a frequency-to-voltage conversion circuit.

The principal motor-driving signal is always supplied by the input demand, and the error circuit is used solely to correct any deviations from the ideal performance.

The electronic gearbox has many obvious applications, wherever a precise drive between two shafts is required with the absolute minimum of backlash and transmission errors. The technique is already being applied in the machine tool industry, and it is expected that many more situations will arise where the extreme precision and smooth transmission properties of this system, and especially its potentially high reliability and freedom from wear, will make it more attractive than its mechanical counterpart.

# H.F. Predictions —August 

The charts show median standard MUF. optimum traffic frequency (FOT) and lowest usable frequency for reception in the UK.

LUFs are calculated by Cable and Wireless Ltd. for point-to-point telegraph circuits. Curves for domestic broadcast reception would be almost identical but for the amateur service would be typically 5 MHz higher at mid-day. The variable effectiveness of low-power services is caused by day-to-day changes in the ionosphere which are on the increase at this time of the year.


# Touch-switch Controller 

by R. Kreuzer

This article describes the operation and construction of three units, a touch switch, a variable d.c. memory and a thyristor power control unit. These units can be used separately in other equipment or together as described here for controlling a.c. power. If used as a lamp dimmer the longer one keeps a finger on the touch switch the brighter the lamps will become.

## Touch switch

The touch switch is a simple high-gain, high input impedance non-linear amplifier. The f.e.t., $T r_{1}$, provides a high input impedance and some voltage gain. The potentiometer $R_{2}$ in the f.e.t's source is the sensitivity control which sets the bias for $T r_{2}$. It should be adjusted so that $T r_{2}$ is just turned on with no input signal to $\operatorname{Tr}_{1}$. When a finger is placed on the touch plate a minute a.c. voltage appears across $R_{1}$ via $C_{1}$ because of the capacitive coupling between the mains cable and the operator. This voltage is amplified by $T r_{1}$ and $T r_{2}$ and a 50 Hz square wave appears across $R_{4}$.

## Memory unit

The square waves across $R_{4}$ charge the capacitor $C_{3}$ via $R_{5}$ and $T r_{3}$ so that $T r_{5}$ (connected as a source follower) provides an output voltage across $R_{8}$. A transistor, $\operatorname{Tr}_{3}$, is used instead of a diode to prevent $C_{3}$ discharging because its base-to-collector reverse resistance is much higher than that of an ordinary silicon diode. However, if the 'diode' is too perfect $C_{3}$ may charge up slowly due to leakage current from $\operatorname{Tr}_{5 \& 6}$. This is unlikely to occur in practice but if it
does a 'less perfect diode' must be used since it is essential that $C_{3}$ should be able to discharge very slowly. The unijunction transistor $\operatorname{Tr}_{4}$ discharges $C_{3}$ when the voltage across $C_{3}$ reaches the emitter trigger voltage of $T r_{4}$; thus enabling the switch to be turned off. The zener diode $D_{1}$ is used to bias $T r_{5}$ so that with approximately 0.5 V on its gate the voltage across the resistor $R_{8}$ is approximately $2 \mathrm{~V}\left(R_{8}=2.5 \mathrm{k} \Omega\right)$. This voltage can be varied by adjusting $R_{8}$. It is essential that when $C_{3}$ has been discharged by $T r_{4}$ the voltage across $R_{8}$ should not be more than 2 V . If this can be achieved only by using very low values of $R_{8}$ then a different voltage zener diode should be used.

## Thyristor power controller

The voltage across $R_{8}$ charges $C_{4}$ via $R_{9}$. At 10 ms intervals $C_{4}$ is discharged by $\operatorname{Tr}_{6}$ because this transistor is operated directly from the rectified mains and, therefore, its emitter junction becomes forward biased when the mains voltage falls to zero. When an input signal is applied the voltage across $R_{8}$ increases, $C_{4}$ charges to the emitter trigger voltage of $T r_{6}$ and $T r_{6}$ produces an output pulse; the thyristor is triggered on. With a high voltage across $R_{8}$, say 4 V , the thyristor is triggered on early in the mains cycle and maximum power is supplied to the load.

The power taken by the touch switch and
the memory is supplied by $R_{12}, D_{3}$ and $C_{5}$ running from the rectified mains. The maximum current taken by the two units is 5.5 mA at 10 V . Diodes $D_{4}$ to $D_{7}$ ensure that control is provided over both positive and negative half cycles of the mains supply High-frequency noise generated by the thyristor is suppressed by $C_{6}$.

## Construction

The method of construction used is up to the individual since it is not particularly critical. The prototype switch was assembled on two $50 \times 50 \mathrm{~mm}$ printed circuit boards one being mounted on top of the other behind the faceplate. The touch plate was a piece of copper foil $25 \times 12 \mathrm{~mm}$ glued to the front of the faceplate and covered by a thin sheet of plastic. The following points should be noted:

The wiring from the touch plate to the gate of $T r_{1}$ should not be longer than 50 mm otherwise feedback from the power supply and cabling to the switch may occur.

All wiring to $C_{3}$ and $R_{5}$ should be as short as possible and must be self-supporting to minimize leakage current.

Resistor $R_{12}$ should be adequately ventilated as it runs hot.
The mains on/off switch should not be omitted. The circuits can then be isolated from the mains for safety reasons.

To test the unit connect a $200 \Omega$ resistor across $R_{12}$, a $4.7 \mathrm{k}^{\prime}$, resistor across $R_{11}$, connect a 12 V a.c. supply to the input of the diode bridge $D_{4}$ to $D_{7}$ and use a 12 V lamp as a load. The unit can then be set up without the danger of getting an electric shock. Remember to remove these additional components before connecting the unit directly to the mains supply. Apart from the diodes $D_{1}$ and $D_{2}$ and $T r_{3}$ the other component values are not critical. Although the prototype employed a 1.5 A thyristor higher current devices may be used. The complete device can be used for dimming lights, controlling heaters or other electrically operated equipment.


## Electronic Building Bricks

## 14. The comparator and subtractor

by James Franklin

In processing information in electronic systems we sometimes wish to compare the value of one electrical quantity with another, decide which is the bigger and which is the smaller, and perhaps measure the difference between the two. This may be needed, for example, in self-adjusting systems-say a power supply stabilizer or an electronic temperature controlleror for the control of switching operations.

Measuring the difference between two quantities is another way of saying subtraction. As such it is an arithmetical process which can be performed electronically by analogue or digital computing methods.

A familiar mechanical analogue of the comparator is the kitchen scales or the laboratory balance. One weight is compared with another and if there is any difference between them the balance arm


Fig. 1. Two batteries connected in series opposition give an overall voltage that is the difference between the individual, battery voltages.


Fig. 2. The subtraction principle of Fig. 1 applied to two voltages which are varying with time.
swings one way or the other (though there is no measurement of the actual difference). The essential principle of the balance, that one weight offsets the effect of the other, can be applied to electrical quantities. We utilize the adding methods shown in Figs 1 and 3 of Part $12^{*}$, but reverse one of the e.m.f. or signal sources so that it opposes, instead of assists, the other. This gives the effect of adding a minus quantity-which of course is the same as subtraction.

For example if we use the method of adding voltages by series connection (shown in Part 12 as Fig. 1 (a)), to adapt this for subtraction we reverse the connections of one of the batteries-say the 6 -volt one, as at Fig. 1. The 6 -volt battery now opposes the effect of the 9 -volt battery because, as an e.m.f. source, it is acting to move electrons in the opposite direction to that in which the 9 -volt e.m.f. source is moving them. The e.m.f. of the 9 -volt battery is offset to the extent of 6 volts and so the net e.m.f. is 3 volts. Thus the subtraction $9-6=3$ has been performed.

This principle can be applied to the subtraction of one continuously varying e.m.f.-a signal-from another. The connections of one of the signal sources are reversed - shown symbolically in Fig. 2 by "Signal source B" being printed upside-down-and then the varying e.m.f. of source $B$, instead of assisting that of source A opposes it. At each instant the effect of the e.m.f. of source $B$ on electron movement is subtracted from the effect of the e.m.f. of source A. This is illustrated graphically in Fig. 3, where the voltage scale for $v_{A}$ is drawn upwards from zero (as in Fig. 2 of Part 12) but the scale for $v_{B}$ is drawn downwards from zero, by convention, so that graph $v_{B}$ becomes a "mirror image" of what it was in Part 12. Values of $v_{B}$, are subtracted from corresponding values of $v_{A}$, giving a set of difference values which are plotted as the graph $v_{A}-v_{B}$ So $v_{A}-v_{B}$ is the varying voltage, or signal, formed by continuously subtracting $v_{B}$.from $v_{A}$.

For subtraction of signals represented by varying currents, again the principle is to use the adding circuit of Part 12 (Fig. 3) but reverse the connections of one of the signal sources so that its e.m.f. acts to move electrons in the opposite direction.

Fig. 4 illustrates this for subtracting $i_{C}$ from $i_{A}$ and $i_{B}$ instead of adding it to them. Electron flow in the common path is the result of the combined e.m.fs of the three sources. In this path there is an aggregate movement of free electrons in one direction resulting from sources A and B assisting each other, but also an aggregate free-electron movement in the opposite direction resulting from the oppositely acting source C. Since number of electrons moved in a given time is electron flow rate, which is current, the net current in the common path is $i_{A}$ plus $i_{B}$ diminished by $i_{C}$ or $i_{A}+i_{B}-i_{C}$. Thus the signal $i_{C}$ is subtracted from the signals $i_{A}$ and $i_{B}$.

Digital subtraction can be performed by for example, a binary computing method or by an incremental system such as a reversible counter. In the last-mentioned, one sequence of events (e.g. electrical pulses). accumulates a total count in the normal way, while another sequence of events causes the counter to work backwards and so diminish (subtract from) this total count.
*Correction. The Electronic Building Bricks article in the May issue, "Adding quantities and numbers", should have been shown as Part 12.


Fig. 3. Graphical illustration of what happens in Fig. 2 over a period of time. At any instant the voltage in the solid-line graph is the result of subtracting $v_{B}$ from $v_{A}$.


Fig. 4. Principle of subtraction with currents. Current in the common flow path due to source $C$ is flowing in the opposite direction from that due to sources $A$ and $B$.

## Charging

# A further look at the CR coupling 

by Cathode Ray

In reviewing basic theory since 1911 for the 60th birthday issue of Wireless World I mentioned that during the second World War I was shocked to find radar instructors teaching that when (say) a positive-going input signal was applied to a CR couping the output also went positive because of the charging of $C$. In actual fact (as I went on to say) any charging or discharging of C appears only as distortion of the signal at the output. I included also the words 'of course', by way of apology to readers for wasting their time by explaining where the quoted teaching was wrong. Wasn't it too obvious in these enlightened days?

Apparently rot, for I soon got a letter to say that it was $I$, not the instructors, who was wrong. Touched though I was by this loyalty to a fine body of men, I felt that this evidence that my own experience of them was not unique called for some more detailed exposition of the point in question, in case the fallacy lingered on in a bigger way than I had suspected. I admit that some trainees might have misunderstood what their instructors taught about this. I will go farther and declare that many trainees did misunderstand what their instructors taught about this and about many other things. So not all that they taught in 1941 should be judged by what their trainees thought they said. And even if some of them were wrong on this point of circuit theory, we won the war so what the hell?

No one is likely to argue that uncertainty on the part of some radar mechs about the precise mode of functioning of interstage couplings in pulse amplifiers was responsible for a major loss of effectiveness in Britain's wartime radar defences, but I will and do hold that anybody who wants to be clever with electronic circuits ought not to have a fundamental misconception about how capacitors function in such circuits. So let's make sure.

The vital fact to be remembered is that the potential difference between the ${ }^{\prime}$ plates of a capacitor cannot change instantaneously, but only as a gradual process due to current flowing in or out.

This follows from the basic equation for capacitance, as important for it as 'Ohm's law' for resistance:

$$
\begin{equation*}
V=\frac{Q}{C} \tag{I}
\end{equation*}
$$

in which $C$ is any capacitance (in farads), $Q$ the electric charge stored in it (in coulombs) and $V$ the p.d. between its plates (in volts). We usually think in terms of current (amps) rather than coulombs, so we also have to remember that $Q=I t$
which means that the charge $Q$ in equation (1) is equal to the amount of current $I$ (in amps) that has been flowing into $C$, and $t$ is the time in seconds during which it has been flowing. (To make things simple we are assuming $I$ is constant.) Putting (1) and (2) together, therefore, we see that the voltage across a capacitor cannot change unless the capacitor receives a proportionate charge, and that takes time. If time were not allowed, $t$ would be zero, so for any charge at all $I$ would have to be infinitely large, which is impossible.

Fig. 1 shows the classic capacitorcharging experiment. Before the switch is closed the capacitor C is uncharged, so in the basic equation (1) $Q=0$, so $V=0$. The moment the switch is closed the voltage $E$ is applied across C and R in series. No time has elapsed since it was closed, so $t=0$, so $Q=0$, so $V=0$ still. So the whole of $E$ appears across R. That


Fig. 1 The familiar circuit used to study the charging of a capacitor.


Fig. 2 The familiar (exponential) charging curve; a graph of voltage against time.
means that a current (call it $I$ ) is flowing through R, and 'Ohm's law' tells us it is equal to $E / R$. That same current is flowing into C , charging it. After one second, $t=1$, so equation (2) tells us that $Q=I$. And we already know that $I=E / R$, so $Q=E / R$, so $V=E / C R$. The capacitor voltage is rising at the rate of $E / C R$ volts per second.

But not quite. By the end of the first second the voltage across $R$ is no longer $E$; it is $E-V$. So the charging current is less than it was, so the rate of charging is less. The nearer the capacitor voltage gets to $E$, the less voltage is left to drive current through R and the slower the charging continues. This is shown by the familiar rate-of-charging curve, Fig. 2. Theoretically the capacitor never quite gets charged to the full voltage applied, $E$, but the deficiency soon becomes negligible.

To continue this lesson in elementary theory we draw the dotted line sloping upwards in Fig. 2 to show how the capacitor would charge if the starting rate could somehow be maintained. The instant at which C would be charged to the applied voltage $E$ is indicated by the point at which the sloping line reaches the $E$ level. Dropping a vertical dotted line from there to the time scale shows (or would do if the scale were graduated in seconds) how long this would take. As our scale is not graduated we will call the answer $T$.

From what we already know we can find a general formula for $T$. Combining equations (1) and (2) by substituting It for $Q$ in (1) we get

$$
V=\frac{I t}{C}
$$

At the end of our imaginary uniform-rate charge, $V=E, \quad t=T$, and $I=E / R$. So

$$
E=\frac{E T}{C R}
$$

and for that to be true

$$
T=C R
$$

I'm quite sure that all the radar instructors included this result in their repertoire, whether or not they proved it in the above or any other way. $T$, the time a capacitance $C$ would take to charge to the applied voltage through a resistance $R$ if the starting rate could be maintained, is called the time constant of the series combination of $C$ and $R$. If they are in farads and ohms (or more conveniently in microfarads and megohms) $T$ will be in seconds.

Because it refers to a mode of charging that doesn't exist in normal practice you might consider all this a waste of time. But as we noted earlier one cannot say how long a capacitor takes to charge in the real practical Fig. 1 way, because theoretically it always takes an infinitely long time, and that is not a very helpful piece of information. The only thing left, then, is to decide on how charged is 'charged'; $99 \%$, say?

The mere suggestion may bring before you a vision of endless committee meetings all over the world trying to agree on a percentage to use as an international standard. Happily, there is no need for this. It turns out that the actual charging curve in Fig. 2 has a fixed shape, so that the time taken to charge to any given percentage of
'full' is an easily calculated factor of $T$, which is so simply equal to $C R$. The simplest possible factor is of course 1 , and it happens that $C R$ is the time taken to charge to $63.2 \%$ of 'full', as shown in Fig. 2. That looks like rather short measure. $99 \%$ requires an odd factor, so I suggest a choice of either $4 C R$ (for $98.17 \%$ ) or $5 C R$ (for $99.33 \%$ ).

The radar instructors probably mentioned the name of the curve of this particular shape (the exponential curve) but they may well have decided it was unnecessary (for the purpose of fitting people to keep radar equipment working) to go into the mathematics of the thing. I too am saying it is unnecessary for our present purpose, and anyone who really wants to know can find it in almost any of the textbooks on electricity (with or without magnetism). The only vital point to carry away just now is that some idea of how long in seconds $C \mu \mathrm{~F}$ takes to charge through $R \mathrm{M} \Omega$ is given by multiplying $C$ by $R$, and that charging is practically complete in 4 or 5 times $C R$.

Now we have got the basic principles straight we can apply them to a circuit of the type which might have given rise to the lecture on CR time constants. It is a circuit in which a square wave developed in the output of one stage has to be passed on to the input of another stage for amplifying, blanking, gating or whatever. Fig. 3(a) shows the relevant part of such a system. Valves are shown, because they were used in wartime radar and because in many cases the input' of the second stage had such a high resistance that R was not appreciably shunted by it. Fig. 3(b) is a transistor equivalent for the benefit of those to whom valves are devices that used to be used, too long ago to be worth trying to understand. But an allowance will have to be made for the shunting of $R$.
The square input waveform is shown in Fig. 3(a), and the object is to reproduce it, with as little distortion as possible, at the

(b)

Fig. 3 The part of a circuit in which the theory developed in Figs. 1 and 2 is useful: (a) the valve version considered, and (b) its transistor equivalent.


Fig. 4 The solid-line square wave is the input to $C R$ (shown in Fig. 3), less any continuous voltage bias, and the dotted line is the output at the junction of $C$ and $R$.
input to the next stage-i.e., the junction of C and R . Of course if direct coupling is used $C$ and $R$ are not needed and distortion does not arise, but with valve circuits especially it is usually necessary to maintain a fixed p.d. between the two stages by means of C , to keep the electrode working voltages right. When considering signal voltages this fixed p.d. can be ignored. So in the signal-voltage/ time graph (Fig. 4) we can assume both the input voltage (applied across C and R ) and the resulting output voltage (across R ) start from zero level.

Up to the point on the time scale marked $a$ the input signal voltage remains at zero, and so does the output, so there is no voltage across the capacitor, so (as equation (1) tells us) it must be totally uncharged. But at $a$ the input suddenly goes $E$ volts positive. (Of course it can't do tinis absolutely instantaneously, but let us suppose that compared with the time ad the rise time is negligible.) This is the point at which I have heard instructors go on to say 'so $C$ charges, making the output (which is the input to the next stage) positive'. But I have, I hope, by now convinced even the most instructorloving reader that it just isn't possible for C to charge appreciably during the rise time, and the fact that the output follows the input and goes positive to the same extent is actually evidence of it. In other words, C does this part of its job by not charging. For as long as it stays uncharged, both sides of it are at the same potential and the output is an exact undistorted copy of the input waveform. The ideal, then, is for C never to be charged, at all.

Let us now consider the state of affairs from $b$ to $c$. Because the input, $E$ volts, is applied across C and R, and the voltage across C alone (at $b$ ) is zero, the whole $E$ comes across R , causing a current to flow through it. Assuming (as we did) that the second valve takes no grid current, all the current has to go into C , beginning to charge it at a rate of $E / C R$ volts per second. The voltage now rising across C is no longer available for R as output voltage. So the output voltage falls. How much it falls in the period $b c$ depends on the time constant, $C R$. If the output is to be undistorted, it mustn't fall at all; which means that $C R$ must be infinite. It can be made very nearly
so by removing $R$ altogether, leaving a gap. But then the grid potential would be at the mercy of stray circuit leakages. To ensure that it starts definitely from zero (or any other designed voltage) R must be used, but its resistance should be made not lower that is needed to anchor the grid to zero volts. Provided that $C$ also is made large enough, the drop in output signal voltage, represented in Fig. 4 by $c c^{\prime}$, can be kept small, as shown. Incidentally, because the voltage across R is nearly constant, the rate of charge is nearly constant and $b c^{\prime}$ is nearly a straight line.
At $c$ the input returns abruptly to zero volts (d), and as the p.d. across C cannot change so quickly the grid side of C drops by the same voltage ( $E$ ). As it started from $c^{\prime}$, less than $+E$ volts, it now goes slightly negative, $d^{\prime}$. This negative voltage, $d d^{\prime}$, to which C became charged during the period $b c^{\prime}$, is now applied to R , through which the charge leaks away during the period $d^{\prime} e^{\prime}$. Because the voltage is so small the rate of discharge is very small and $d^{\prime} e^{\prime}$ is practically horizontal. So when the input goes positive again, from $e$ to $f$, the output at $f$ is practically the same as at $c^{\prime}$. It therefore starts its decline during the next positive half-cycle from a lower voltage than it did in the first.

## Effect of d.c. barrier

So long then as the output half-cycles continue to be more positive than negative, the different rates of charge and discharge bring them gradually more nearly equal, as shown by the dotted waveform in Fig. 5. In the end, whatever the input waveform, the output will arrange itself so that the time $\times$ voltage area below the line is equal to that above the line. The line, of course, represents the level to which the output is anchored by R ; in this case zero volts. This phenomenon, which we have been examining in detail, results inevitably from the fact that a capacitor is a barrier to d.c. So a signal that starts (as in Fig. 4) all above the line, or more one side of the line than the other, inevitably adjusts itself so that this d.c. component disappears and the output is wholly alternating. The less the time constant $C R$ the faster it adjusts-and the more distortion it introduces.
If the signal frequency is very low, so that C has a long time in which to discharge during each half-cycle, a very long time constant is needed to avoid appreciably distorting a square wave. And the system takes a very long time to readjust to a change of input amplitude. This problem arises in oscilloscopes where capacitance couplings are used in the deflection ampli-


Fig. 5 How the voltage/time graph started in Fig. 4 continues.


Fig. 6 Here for comparison with Fig. 4 is what happens when the time constant is only a fraction of one half cycle.
fiers. It is so tedious waiting for them to settle down that nowadays designers almost always provide direct-coupled amplifiers.

The d.c.-losing effect can be prevented by suitably connecting a rectifier in the circuit, creating a 'd.c. restorer'-but that is another story.

The only other thing I think I need mention-and it will be familiar to radar trainees past and present-is that a CR coupling is often used not to pass on the original undistorted form but to introduce deliberate distortion. The commonest application is for changing square waves into brief pulses. For this purpose the time constant is made much less, so that instead of a gradual charge such as $b c^{\prime}$ in Fig. 4 the capacitor charges practically completely within the half-cycle, as in Fig. 6. When the end of the square-wave half-cycle comes (cd) the output going negativewards by the same amount ( $c^{\prime} d^{\prime}$ ) yields equal negative and positive half-cycles from the start. The negative ones can then be removed by a rectifier and the positive ones clipped by another, to give a train of pulses.

Note that (whatever the instructor said) C charges from $b$ to $c^{\prime}$ and discharges from $d^{\prime}$ to $e^{\prime}$, in Fig. 4 and in Fig. 6.

I used to find that even fellows who could state Kirchhoff's voltage law quite correctly when asked for it seemed to forget all about it when considering the CR type of circuit. One form of the law says that the sum of the voltages across the components in a series circuit is equal to the voltage applied. Now in Figs. 4 and 6 the voltage applied is represented by the height above zero of the 'input' waveform: alternately $E$ and O. The Voltage across R ('output') is represented by the height of the dotted line, so the voltage across C (due to its charge) must, by Kirchhoff's law, be the vertical difference between the two. Looking at the matter this way, one can be in no doubt about when and how much the capacitor is charging and discharging.
The essential thing is to grasp the message of Figs. 1 and 2. Then a correct view of the action of any CR circuit is (to coin a phrase) a piece of cake.

## Single-sideband Experimental Broadcasts

For some years there have been discussions on the possibility of utilizing the medium-wave sound broadcasting band more effectively by means of single-sideband transmissions. At first sight it seems attractive in view of the fact that s.s.b. is now so well established in h.f. communications. But there are complications in reception, the main one being that the simple envelope detector found in conventional sound receivers inevitably leads to excessive distortion and must be replaced by a product detector, in which case. for tuning, a local oscillator of high stability, among other things, is required. In Britain the broadcasting authorities don't seem very enthusiastic about s.s.b. but in Germany there is considerable interest-measured by the fact that the Deutschlandfunk broadcasting organization has been putting out experimental s.s.b. transmissions from its station at Mainflingen. near Frankfurt.
The broadcasts took place in the early hours of the morning. after close-down of normal broadcasting. on 1475 kHz . At least one group of British radio research people was willing to stay up in order to study and listen to the transmissions. This was a radio section of the Department of Electrical and Electronic Engineering at University College Swansea, headed by Dr. R. C. V. Macario (author of an article on an s.s.b. receiver module in the July
issue). Some results of their monitoring are shown in the accompanying frequency spectra. Fig. 1 is a 200 kHz wide part of the m.f. spectrum showing the s.s.b. transmission at 1475 kHz , in relation to the permanent a.m. transmission from the Mainflingen broadcast ing station on 1538 kHz and to Radio Luxembourg on 1439 kHz . More detail can be seen in Fig. 3, which is 50 kHz wide. The upper sideband of the s.s.b. transmission can be seen as an asymmetrical distribution of energy in contrast to the symmetrical distributions. like church spires. of the a.m. stations on each side of it. In Fig. 4 the frequency scale is 20 kHz wide and shows the upper sideband in even greater detail.

The carrier alone of the s.s.b. transmission was suppressed 20 dB below the peak sideband levels, and is shown in Fig. 2, on a frequency scale 20 kHz wide.
The spectra were displayed on a HewlettPackard spectrum analyser, model 8552A 8553 L , with a stored display. A simple roof wire aerial was used. Recordings of the transmissions were made via various receiving systems, but it is interesting to note that direct conversion was possible since the lower sideband of the transmission was relatively free of interference.


# Telephone Exchanges 

## of the Future

A new type of telephone exchange is in operation at a GEC-Marconi establishment (at Writtle in Essex) which does not use any electromechanical switches or in fact any moving parts. The system is called Martex and is typical of the sort of exchange which is to be built in the future.

The system is a modular range of equipment which covers all aspects of switching and transmitting telephone traffic, and some types of data communications. The complete system is based on the use of digital switching and computer techniques to switch information in digital form.

Equipment employing pulse code modulation, a particular form of digital speech transmission, is now being used increasingly by the Post Office, to increase the capacity of existing telephone lines. Each telephone channel is converted into a stream of digital pulses which provide a complete representation of the original voice signals. These signals can be reconstituted into normal speech with rather less loss of quality and fidelity than would be experienced by a conventional telephone transmission line.

The great advantage of the digital network of transmission is that the spaces between consecutive pulse groups from a single voice input are arranged to be sufficiently large for a number of other pulse streams, from other telephone circuits, to be fitted into these spaces. If this is done in an ordered fashion, then a number of separate telephone inputs can be fed simultaneously along the same transmission line, and separated at the far end into the original voice signals.

This method of combining channels is known as time division multiplexing, t.d.m. It has the advantage that signals in this form make better use of the digital switching equipment.

At the start of a call, the first event in the complete sequence will be the lifting of the receiver, which will initiate a demand for a signal path into the exchange system. This will be established through a local concentrator system, which will allocate a particular time slot in one of the digital input circuits of the exchange. In the exchange system, a register will be connected to the appropriate line, through the digital switch returning 'dial tone' to the calling subscriber. The subscriber will
then dial a code, using either a conventional rotary dial, or push-buttons. When the register has accepted the complete information, the control computer in the exchange will examine the contents of this register. Using information from a magnetic drum store, it will generate control signals to produce the appropriate switching functions in the exchange, together with additional switching instruction codes for onward transmission to a subsequent exchange, depending on the routing of the call. These latter will be assembled in the memory of the 'sender', and transmitted through the system when the switching operation is complete.
On arrival at the exchange, each speech channel will have been converted into digital form by its relevant p.c.m. terminal, allocated on a demand basis in a local exchange system. The digital signals are multiplexed into groups of 30 speech channels to form a single time division multiplex signal. Two additional channels (or time slots) provide control and supervisory information. This format uses a total of 32 time slots in each signal 'frame', with a frame repetition rate of 8 kHz . Each slot contains eight digital bits which define the polarity and amplitude of the speech sample being transmitted. Each incoming speech channel is thus sampled at a rate of eight thousand per second. These groups of channels enter the exchange switching system over digital transmission paths linked directly with the digital switch and its associated control system.
Concentrators will be employed which will enable large numbers of subscribers economically to be connected to a central digital exchange. The concentrators will replace small local exchanges, and will normally be connected to the main exchange through three digital links, providing for up to 90 subscribers to be connected simultaneously to the main exchange. With normal circuit usage, this would cater for 1,000 subscribers per
concentrator. Twenty five or more concentrators may be connected to the Martex switch to deal with up to 25,000 subscribers.

Each digital input circuit consists of thirty speech channels with an additional two supervisory and control signal circuits. This produces a stream of digital pulses in which every 32nd group of pulses, or time slots, relates to a particular speech channel.

Switching will require connection of input and output circuits in either the same time slot or in a different time slot. In the first case connection is by a relatively simple switching action, but in the second. time delays have to be introduced into both directions of transmission to match up the input and output circuits. This process is in addition to the normal switching process, and is also carried out under the direction of the central computer.

In both cases, any incoming signal, on a given digital input circuit, will need to be connected with another digital output. This part of the switching is carried out by providing physical connections between the appropriate wires on a matrix of crossed wires. The connection is made through solid-state digital switches which are pulsed at the 8 kHz repetition frequency of the appropriate pulse group in one of the 32 time slots in each multiplexed input.

However, in general, a second type of switching, incorporating a time delay, will have to be introduced to each switched circuit, in order that it will match up with the appropriate time slot in the output circuit.

If, for example, in order to establish a particular connection, it is necessary to connect the third time slot in one multiplexed input signal to the twelfth time slot of another multiplexed output channel (i.e. nine time slots later), it is necessary to delay the input signals by the equivalent of nine time slots in the forward direction of $35.2 \mu \mathrm{~s}$, and 23 time slots or $89.9 \mu \mathrm{~s}$ in the reverse direction. This is achieved by the use of 'junctor' units, which use shift registers, controlled by the central control computer, to provide the appropriate time delay.

The program control unit contains a number of processors in a fully triplicated system. Fixed program. read-only stores, provide the basic programming for the computer control system, while drum stores are used for channel routing instructions and other semi-permanent control data. Magnetic tape units are used to record call charge data and accounting information.
All critical parts of the system are fully triplicated, with a constant comparison of the data passing any point in the system. A majority voting technique is employed to ensure that a fault in one of the three systems will not introduce errors. In the event of two failures at parallel points in the system, the third channel can be switched to provide a continuous service. All three systems work in synchronism under the control of the exchange clock, to ensure that comparable data arrive at the voting point simultaneously.

# Elements of Linear Microcircuits 

## 10: Amplitude modulated radio receivers

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

Despite the increasing number of f.m. sets in use, most domestic and car radio receivers are still a.m. only, usually covering the m.w. band, 540 to 1640 kHz , and sometimes also the l.w. band, 155 to 280 kHz . In this article, we will take a look at the application of linear microcircuits in this field.

When off-the-shelf linears first began to come into the hands of set designers in the mid 1960 s , ther offered a possible alternative to the use of six to ten separate transistors in a conventional superhet circuit, which had by then become almost a way of living. This market presented a tempting large-scale outlet to semiconductor manufacturers, and as a result a lot of effort has gone into trying to develop microcircuits for a.m. receivers.

The ideal microcircuit design for this purpose would be a device with all active and passive circuit components incorporated with the exception of the aerial, tuning control anci indicator, volume control, loudspeaker and power supply. This may come some day, but for the present we must be satisfied with microcircuits which do not go as far as this.

Most approaches to the problem started from the corventional superhet circuit arrangement and were aimed at producing monolithic silicon chips containing as many of the transistors, resistors and capacitors of the discrete designs as possible. However, one school of design (using phase-locked-loop techniques to be described later) has abandoned the conventional superhet.

## Partitioning superhets

If you cannot reach the ideal solution of the single chip then you are faced with the problem of how to break the superhet down into sections. Receiver designs using i.cs have followed three main paths:

Discrete approach, in which only the active components are integrated. This fails to make use of the full potential of the monolithic circuit art because separate passive component counts are not reduced.
Functional approach, in which single functions of the receiver are fabricated in separate monolithic circuits and are

[^6]assembled with additional discrete components to form a complete radio.
System approach, in which multiple receiver functions (e.g. the mixer, oscillator and i.f. amplifier) are fabricated on the monolithic circuit chip.

The discrete approach soon proved to have no advantages over discrete assembly, and is of historical interest only. The functional approach, too, proved uncompetitive with discrete assembly but, although it has now been abandoned, we
will take a look at one example of it as a significant step towards current practice.

## Single i.f. stage

Fig. 1 (a) shows the internal circuitry of the Motorola MC1550G, a versatile common-emitter, common-base cascodecircuit high-frequency amplifier capable of 30 dB gain at 60 MHz but which can be used for a 470 kHz i.f. amplifier in the circuit of Fig. 1 (b).

It will be seen that all the resistors and


Fig. 1. Example of single-stage integration; (a) internal circuit of Motorola MC1550G r.f./i.f. amplifier; (b) MC1550G in single i.f. stage.


Fig. 2. Internal circuitry of Mullard TAD100 a.m. radio receiver microcircuit handling signal from local oscillator via mixer up to audio driver stage.
semiconductors for the single stage have been integrated, and apart from the $L C$ bandpass circuits, only three external capacitors are required.

## One chip, r.f. in to a.f. out

The Mullard TAD 100, whose circuit diagram is shown in Fig. 2, was one of the first i.cs designed specifically for a.m. radios. The design aim was a low-cost integrated circuit (not too expensive for economic
service replacement), with performance not worse than that of conventional discrete-component receivers, and in standard 14-lead dual-in-line package. It incorporates no fewer than 11 transistors and three diodes, together with many of the passive components from the mixer to the audio pre-amplifier.
$\operatorname{Tr}_{1}$ and $\operatorname{Tr}_{2}$ form a long-tail pair mixer stage, and $\operatorname{Tr}_{3}$ is the local oscillator. $\operatorname{Tr}_{4}, 5$ and $\operatorname{Tr}_{6}$ comprise a high-


Fig. 3. Broadcast-band a.m. receiver design (9V) utilizing TA D100 microcircuit.
gain wideband amplifier for i.f. amplification, and $\operatorname{Tr}_{7}$ is a transistor detector. $T r_{8}$ and, are a long-tail pair audio preamplifiers and $\operatorname{Tr}_{10} \quad \operatorname{Tr}_{11}$ a Darlington common collector audio driver stage. Diodes $D_{1}, D_{2}$ in parallel, back to back, across the oscillator transistor collector coil terminals, serve to stabilize the local oscillator. $D_{3}$ is a level shifting d.c. coupling diode to the input of the driver stage.

Typically the TAD 100 takes about 20 mA quiescent current in a 9 V circuit. Its sensitivity for a 26 dB signal-to-noise ratio (a standard index) is typically $25 \mu \mathrm{~V}$ at input terminal (1). Its a.g.c. range controlled by feedback from (8) to (1) is typically 62 dB change in r.f. input voltage for only 10 dB expansion in audio output. For 10 mV audio at the detector load, less than $6 \mu \mathrm{~V}$ r.f. input is required at the input.

You can see how the TAD100 is used in practice in the 9 V broadcast-band receiver arrangement of Fig. 3. A $180 / 280 \mathrm{pF}$ gang capacitor tunes the rod aerial coil $L_{1}$ and the oscillator coil $L_{3}$. The r.f. input is connected across (1) and (13), and the local oscillator drive feeds into (13); a.g.c. is fed back from (8) into (1) via a decoupling network and $L_{2}$. From (3) a $560 \Omega$ resistor to the negative supply (shunted by a series $56 \Omega$ resistor in series with $0.047 \mu \mathrm{~F}$ ) forms the tail of the input long-tail pair. The mixer output from (14) feeds into the input (a) of the $470 \mathrm{kHz} \mathrm{LP1175} \mathrm{block} \mathrm{filter}$, combination of two tuned $L C$ circuits with a ceramic resonator as shown separately inset in Fig. 3. The LP 1175 gives the typical normal 6 dB bandwidth of 7 to 8 kHz and a significant improvement in skirt selectivity over conventional fixedtuned i.f. transformers.

From the filter output (b), the i.f. signal passes into (10) and is amplified and detected to reappear from (8) to provide the audio drive to the top end of the volume control and the a.g.c. signal to be


Fig. 4. internal circuitry of S.G.S. TBA651 a.m. radio receiver microcircuit handling signal from r.f. amplifier up to i.f. output.


Fig. 5. 12v broadcast-band a.m. car radio receiver utilizing TBA651 microcircuit.
fed back to (1). From the volume control slider the audio is fed into (4) and reappears amplified at (6) to drive the output stage. In this design the output transistors are a discrete $n-p-n / p-n-p$ pair in singleended push-pull, capacitor-coupled to a $4 \Omega$ loudspeaker to give over 1W output.

At first sight there seems still to be a very large number of components outside the microcircuit, but it should be noted that most of them are passive and of wide tolerance, and unlikely to give trouble in assembly. Also the use of a block i.f. filter requiring no 'adjustment' simplifies set assembly.

## One chip, r.f. in to i.f. out

The TAD 100 was designed to integrate as much of the a.m. receiver as practicable. The a.f. output stage was left out because of dissipation limitations in the package used. A different partitioning was adopted by S. G. S. in their TBA651 linear integrated circuit that processes the whole high-frequency signal in a.m. receivers. It consists of five stages: r.f. amplifier, mixer, oscillator, i.f. amplifier, and a.g.c. control and voltage regulator and was designed primarily for high quality domestic and car radios. This explains the inclusion of a separate r.f. amplifier stage, and also the ability to work from voltage rails of 4.5 to 18 V . The circuit is packaged on a 'split' (staggered pins) 16-lead dual-in-line.

In Fig. 4 you will find details of the internal circuitry of the TBA651. $\operatorname{Tr}_{1}$ is an r.f. amplifier; $T r_{6}$ and $T r_{7}$ the mixer; $T r_{5}$ (with $T r_{4}$ ) the local oscillator; $T r_{2}$ and $T r_{3}$ the a.g.c. control on the r.f. amplifier; $\mathrm{Tr}_{8}$ and $T r_{9}$ (with $T r_{10}$ tail current source), $T r_{11}, T r_{12}, T r_{13}$ the i.f. amplifier; and $T r_{14}, T r_{19}, T r_{16}, T r_{17}$ a voltage regulator circuit providing three output voltages to set the d.c. bias conditions of the various transistors.

An a.m. car radio circuit using the TBA651 is given in Fig. 5. A three-ganged permeability unit tunes the aerial input, r.f. amplifier and local oscillator circuits. A double-tuned i.f. bandpass circuit $L_{4}$ and $L_{5}$ connected between (5) and (13) in series with the input to the i.f. amplifier section provides part of the required i.f. selectivity and the balance is provided by the single-tuned circuit $L_{6}$ at the i.f. output (10). The input $L C$ filter can be replaced by a ceramic-plus-LC filter similar to the LP1 175 for greater skirt selectivity.

In Fig. 5 it will be seen that a conventional a.m. diode detector is used externally to the TBA651; unlike the TAD 100 where a transistor detector is included in the microcircuit. After the volume control, a number of arrangements are possible. In Fig. 5 the monolithic TAA611/B is used to drive a pair of output transistors (medium power, with a current gain at 3A of greater than 20 ) to give 5 W output. A number of completely integrated $5 \mathrm{~W}, 12 \mathrm{~V}$ audio amplifiers are coming on the market with sufficient gain to be driven direct from the volume control in applications such as these, and ultimately we should see two-chip complete radio receivers.

## Phase-locked-loop alternative to the superhet

The difficulty of microminiaturizing frequency selective circuits has shown the lack of adaptability of the conventional superheterodyne system to an integrated radio receiver, particularly in the lower frequency bands. Because of this, designers are exploring systems.that do not call for such fixed-tuned frequency selective circuits.

One area where there is much activity is the p.l.1. (phase locked loop) receiver. This has been around as an idea since the early

1930s, when H. de Bellescize published an article on 'La Reception Synchrone' in e'Onde Electrique, Vol. 11, pp. 230-240. June, 1932. Nothing came of this, but in Electronic Engineering, pp. 75-76, March, 1947, D. G. Tucker raised the matter again in 'The Synchrodyne'. The p.l.l. receiver also goes variously under the names of 'Homodyne', 'Synchronous Detector', 'PL' (phase locked) and 'PC' (phase coherent).

Fig. 6 (a) shows the principle of the phase locked loop. A carrier of amplitude $A_{c}$ frequency $f_{c}$, and phase $\phi_{c}$, with modulation $S$ is applied to a phase detector which compares this input with the unmodulated output from a local oscillator of amplitude $A_{o}$, frequency $f_{o}$, phase $\phi_{o}$. If the local oscillator frequency is adjusted to equal the carrier frequency, the phase detector gives an output proportional to the phase difference $\theta=\phi_{c}-\phi_{o}$ between the input and oscillator phases. This output is then passed through a low-pass filter and an amplifier and fed back to vary the control voltage on the local oscillator in such a way as to reduce the phase difference between the two signals. The end result is that the local oscillator phase advances or retards until it is in phase with the carrier phase. The local oscillator need not be tuned exactly to the carrier frequency for the phase locked loop to operate. There is a capture effect, in that the local oscillator need be brought only roughly to the carrier frequency and the system then pulls into frequency and phase synchronism with the carrier.

The most elementary p.l.1. receiver can consist of a voltage-controlled local oscillator, a mixer (phase detector) and an audio amplifier with the audio signal fed back to control the local oscillator. In the mixer the signal carrier is converted to a

(a)


Fig. 6. The phase-locked-loop receiver alternative to the superhet; (a) basic phase-locked-loop; (b) system layout for phase-locked-loop a.m. receiver capable of implementation in microcircuit form.
zero-frequency intermediate frequency, the output from the mixer containing only demodulated information from the sidebands.

There are now indications from theoretical and experimental investigations that p.l.1. receivers are performance and costwise competitive with (even perhaps better than) conventional superhets. And the important thing is that the fixed-tuned $L C$ bandpass circuits of the superhet are avoided.

The p.l.1. receiver has some distinctive advantages over the superhet, apart from the lack of i.f. coils. Any interference will not be synchronous with the local oscillator, so that the mixer output resulting from an interference signal will be a beat note suppressed by the audio filtering. Also there is no image response in the system because the intermediate frequency is zero. These nearly ideal selectivity characteristics and the lower possible thresholds of reception have led to the wide use of p.1.1. receivers in difficult signal environments such as reception from artificial satellites. where low signal level, doppler shift and oscillator drift present problems. In the more mundane field of a.m. receivers, p.1.1. techniques have hitherto been prohibitively expensive, but now monolithics are appearing which would seem to make the p.l.I. domestic receiver a strong contender.

The National Semiconductor LM565
phase-locked-loop (although essentially a high quality professional microcircuit) is indicative of the sort of circuit that will soon become available to set designers. It contains a stable, highly linear voltage controlled oscillator and a double balanced phase detector. The v.c.o. (voltage controlled oscillator) frequency is set with an external resistor and capacitor, and a tuning range of $10: 1$ can be obtained with the one capacitor.

Fig. 6 (b) shows the outline of an a.m. p.l.I. receiver system that could be put together with currently available monolithic microcircuits. The r.f. input from the aerial is passed through a tunable r.f. amplifier. Unfortunately this still involves some form of inductance. The main purpose of the r.f. amplifier is to reject harmonics of the signal frequency to which the mixer might respond. The bulk of the receiver gain will still be at audio frequencies.

From the r.f. amplifier the input signal passes to the in-phase mixer (which can be a simple diode bridge) where it is mixed with the output from the v.c.o.-not directly but with a $+45^{\circ}$ phase-shift. The frequency of the v.c.o. will have been adjusted to approximately the right value from the tuning control. The in-phase mixer acts as a phase (and frequency) detector. The output then passes through the low pass amplifier and back via the
second phase detector, the a.p.c. (automatic phase control) filter to lock the v.c.o. to the frequency and phase of the r.f. input.

The output from the r.f. amplifier is also fed into the quadrature mixer where it is mixed with a $-45^{\circ}$ phase shifted output from the v.c.o. Through the second loop amplifier and the path phase detectora.p.c. filter it also helps to lock the v.c.o. on signal. The quadrature signal channel can be used to drive a visual tuning indicator.
A difficulty with p.l.I. receivers is that an annoying beat note 'heterodyne whistle' is heard as the receiver is tuned between stations. This can be eliminated by a threshold detector and a.f. squelch gate. When the receiver is off-tune, there is a significant output from the quadrature channel which activates the threshold detector and holds the squelch gate closed thus suppressing audio output. On tune. the quadrature channel output falls to virtually zero, the squelch gate is opened and audio output passes to the a.f. amp and the louspeaker.

Finally, an a.g.c. signal is taken from the in-phase channel via the a.g.c. amplifier to control the gain levels of both the r.f. amplifier and the local oscillator.

You can find a fuller discussion of the p.l.I. receiver described above in L.P. Chu 'A phase-locked a.m. radio receiver' in Trans. I.E.E.E. Vol. BTR 15, No. 3. pp 300-308, Oct, 1969. For the whole subject of phase-locked-loops an excellent standard reference is 'Phaselock Techniques' by F.M. Gardner. John Wiley and Sons. 1966.
(to be continued)

## Conferences and Exhibitions

Further details are obtainable from the addresses in parentheses
overseas
Aug. 11-13
St. Louis
Automatic
Automatic Control
(I.E.E.E., 345 E. 47 th St., New York, N.Y. 10017)

Aug. 16-20 Jerusalem
Impact of Computers on Developing Nations
(Jerusalem Conf. on Information Technology, 75 Grosvenor St., London WIX ODT)
Aug. 17-19
H.F. Generation Ithaca
(Prof. L. Eastman, Cornell School
Phillips Hall, Ithaca, N.Y. 14850) of E. Eng.,
Aug. 18-26
Budapest
Acoustics Congress
Aug. 23-28
Microwave Conference
(Dr. H. Steyskal, Fack 23, 104 50, Stockholm 80) Aug. 24-27

San Francisco
Western Electronic Show \& Convention
(WESCON, 3600 Wilshire Blvd, Los Angeles, Calif. 90005)
Aug. 25-27
Washington
Geoscience Electronics
(I.E.E.E., 345 E. 47th St., New York, N.Y. 10017) Aug. 27-Sept. 5

Berlin
International Radio \& TV Show
(A.M.K., Messedamm 22, 1 Berlin 19)

# Complementary Darlington Output Transistors in Audio Amplifiers 

## Product application note

Circuit shown right is designed around integrated Darlington power transistors, made by Motorola. With these, external bias componenss are not needed and their high gains limit the gain and power dissipation requirements of driver transistors, thus simplifying amplifier designs. Design is suitable for power outputs from 15 to 60 W working into a loudspeaker of 4 or $8 \Omega$-see table. This and a direct-coupled version are contained in Motorola application note AN-483A.



Circuit gives harmonic distortion of less than $0.2 \%$ at rated output from 50 Hz to 20 kHz and $0.1 \%$ at 100 mW output from 200 Hz to 20 kHz , rising to $0.25 \%$ at 20 Hz for both power levels. Intermodulation distortion is $0.2 \%$ at half power with 1 kHz and 10 kHz signals in 4:1 ratio. Resistor $R_{11}$ sets bias current-20mA-to minimize crossover distortion. As an alternative to bootstrapping, $\operatorname{Tr}_{5}$ base is connected to a constantcurrent source- $\operatorname{Tr}_{4}$ and diodes $D_{1}$. (Resistors marked with an asterisk should be $5 \%$ tolerance, others $10 \%$.)
Several short-circuit protection techniques can be used. The short-term one shown (left) allows a short-circuit to be driven for a few minutes-average power dissipation increasing by four times-using heat dissipators with thermal resistance specified in the table and at $25^{\circ} \mathrm{C}$ ambient temperature.

Components for 15 to 60 watt amplifier not specified in circuit

| Rated power W | $\begin{aligned} & \text { load } \\ & 2 \\ & S \end{aligned}$ | $\begin{aligned} & R_{12} \\ & \Omega \end{aligned}$ | $\begin{aligned} & V_{c c} \\ & \Omega \end{aligned}$ | $\begin{aligned} & \boldsymbol{R}_{5} \\ & \Omega \end{aligned}$ | $\begin{aligned} & R_{7} \\ & \mathrm{k} \Omega \end{aligned}$ | Tr ${ }_{1} .4$ | $\mathrm{Tr}_{2}$ | $\mathrm{Tr}_{3}$ | $T r_{5}$ | Tr 6 | $\begin{aligned} & C_{7} \\ & \text { rating } \\ & V \end{aligned}$ | $C_{2}{ }_{\text {rating }}$ V | $c_{4}{ }_{\text {rating }}$ v | heat <br> sink ${ }^{+}$ |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 15 | 4 | 330 | 32 | 620 | 33 | MPSA05 | MPSA55 | MPSU01 | MJE1100 | MJE1090 | 35 | 20 | 40 | 9.5 |
| 15 | 8 | 150 | 38 | 510 | 39 | MPSA05 | MPSA55 | MPSU01 | MJE 1100 | MJE1090 | 40 | 25 | 45 | 9.5 |
| 20 | 8 | 470 | 38 36 | 510 560 | 39 | MPSA05 | MPSA55 | MPSU01 | MJE1100 | MJE1090 | 40 | 25 | 45 | 7.0 |
| 20 | 8 | 180 | 46 | 470 | 47 | MPSAO5 | MPSA55 | MPSUO1 | MJE1 100 | MJE1090 | 50 | 30 | 55 | 7.0 5.0 |
| 25 | 4 | 510 | 38 | 560 | 39 | MPSA05 | MPSA55 | MPSUO1 | MJE 1102 | MJE1092 | 40 50 | 25 30 | 45 55 | 5.0 5.0 |
|  | 8 | 220 | 48 | 390 | 47 | MPSA05 | MPSA55 | MPSUO1 MJE520 | MJE 1100 MJ3000 | MJE1090 MJ2500 | 45 | 25 | 50 | 6.0 |
| 35 | 4 | 750 | 44 | 470 | 47 | MPSA05 | MPSA55 | MJE520 | MJ3000 | MJ901 | 60 | 35 | 65 | 5.5 |
|  | 8 | 390 | 56 | 330 | 56 | MPSA06 | MPSA56 | MPSU01 | MJ3000 | MJ2500 | 50 | 30 | 60 | 4.0 |
| 50 | 4 | 910 | 50 | 390 | 47 | MPSAO5 | MPSA55 | MJE520 | MJ3001 | MJ2501 | 65 | 35 | 75 | 4.0 |
|  | 8 | 560 | 65 | 270 | 68 | MPSAO6 | MPSA56 | MJE520 <br> MJE520 | MJ3001 | MJ2501 | 60 | 35 | 65 | 3.0 |
| 60 | 4 | lk 620 | 56 72 | 330 220 | 56 68 | MPSAO6 MPSAO6 | MPSA56 | MJE 520 | MJ3C01 | MJ2501 | 75 | 40 | 80 | 3.0 |

[^7] p. 22 of January 1 §.71 issue (instruction 5).

## Automatic Titration Potentiometer

by D. R. Bowman, m.I.E.R.E.

The instrument described employs dual-gate m.o.s.f.e.ts and was originally intended to monitor a chemical process known as titration. However the measuring circuit can be used for other applications in which an electrometer is required.

A measuring circuit was required that would link the output of a very high internal impedance probe with an indicating apparatus such as a char recorder. The probe in question had an internal impedance in the kilo-megohm region and an output of between 100 and 400 mV . One of the various thermionic electrometer valves available would have performed well but with the disadvantage of requiring h.t. and l.t. power supplies. Investigation of the various semiconductor amplifying devices available revealed that only the m.o.s.f.e.t. approached the input resistance requirement. Previous experience with these transistors has taught the author to be wary, for although the gate-to-source breakdown rating may be 20 or 30 V the high inherent resistance inevitably means that even the smallest charge cannot leak away and is liable to accumulate until the gate insulation is destroyed.

A number of transistor manufacturers
being alive to this problem have introduced devices with zener diodes internally connected across the gate electrode. The diodes exhibit a very high shunt resistance until the potential across them exceeds about $\pm 6 \mathrm{~V}$. At this potential their resistance drops to a low value and so protects the transistor's gate insulation.
The basic circuit, which is shown in Fig. 1, is a differential amplifier. The d.c. level drift with temperature, an always present problem in electrometer amplifiers, is not so serious here because the drifts in the two transistors are in opposition and therefore tend to cancel each other.
To maintain the maximum input resistance a gate leak resistor has not been included, however, the probe's series resistance provides an earth return for the gate electrode. The first device operates as a source follower, the inherent negative feedback tending to maintain the high input resistance. The second stage is connected as a common gate amplifier.


The overall power gain provided by the amplifier is of the order of 70 dB .

The second gate electrodes of the cascade devices are connected together and biased to about 0.6 of the drain potential. The two source electrodes are taken via a potentiometer to earth. This couple is adjusted for minimum thermal drift. The output potentiometer alters the gain slightly, but is primarily intended for setting the output to zero when there is no no input signal. The exact amplitude of the output signal is unimportant when the instrument is used in titration so this deficiency has not proved to be a great disadvantage. No attempt has been made to match the m.o.s.ts and yet the temperature stability has proved to be adequate.
Two transistor types are suitable, the 40673 and the 3N187. Of these the 40673 seems to be the best choice; it is identical in performance with the 3 N 187 , but is considerably cheaper.

## Power supply

The circuit shown in Fig. 2 exhibits a very low output ripple together with automatic overload protection. As the series regulating transistor is capable of supplying at least 200 mA other auxiliary equipment can be connected to the supply if required. In the diagram an unearthed 9 V unit is shown whose polarity can be altered by earthing either the positive lead to produce -9 V or the negative lead for +9 V . This instrument requires two such supplies, one of each polarity. The mains transformer used is a Radiospares miniature type with two 12 V secondaries, but any other transformer with two independent secondaries will do as the current requirement is only 10 mA . The two power supply circuits should be adjusted to provide about 9 V .
The setting up of the amplifier is extremely simple; the only point needing description being the minimum thermal drift adjustment. The dual gate m.o.s.ts are mounted in an electrically insulated dual heat sink. A hot soldering iron should be brought into thermal contact with this heat sink and the potentiometer adjusted


Fig. 2. Power supply circuit. Two of these are required.
for minimum drift as shown by the recorder. The gain potentiometer should be used to set the amplifier for zero signal out with the input short circuited.

## Titration

Many quantitative chemical analyses are made by adding measured amounts of acid to the unknown alkali solution until the two cancel one another out to leave a neutral solution. The stage at which balance occurs, the end-point, has to be determined very accurately and is normally done using one of the coloured chemical indicators of which Litmus is an example.

If two electrodes are dipped into the solution during the titration process the voltage across these electrodes will change as the solution goes through the neutral point. It was to detect this change that this instrument was designed.

The probe employed was a 'Silver billet combination electrode' (Cat. No. 39187). During the titration process the mixture was stirred using a magnetic stirrer and a piston burette was used to add one liquid to the other. The piston burette is driven by a motor and adds liquid at an accurately known rate. As this motor is synchronous the chart recorder and the piston burettes will automatically keep in step.

The titration probe and amplifier tend to be sensitive to noise generated by the


Fig. 3. Chart recorder A is set for 50 mV f.s.d. and is the potentiometric titration recorder. Chart recorder B is set for $5 m V$ f.s.d. and is the differential potentiometric titration recorder.
magnetic stirring system and for this reason a switched filter has been included in the input circuit of the amplifier. This filter should be used with care, only enough smoothing being used to reduce the noise or the response of the whole system may become excessively damped. Fig. 3 shows a basic differentiating circuit which if applied to the output of the amplifier and used in turn to drive the recorder makes the titration end point on the graph more easy to discern. With this simple circuit it will be necessary to increase the sensitivity of the chart recorder.

## Sixty Years Ago

August 1911. Two reports in this issue were concerned with mobile communications. An article 'Wireless Telegraphy and Aeroplanes' described an experimental installation as follows:
"In a lecture before the Royal Institution, Mr. T. Thorne Baker passed in review some of the work already accomplished in the application of wireless telegraphy to aerial navigation and referred to some satisfactory results obtained by Mr: Farman by using two trailing aerials, each consisting of rather thin wire about one hundred metres in length. Those experiments were carried out. some time after Mr. Baker had adapted a similar arrangement to a Bristol
biplane in this country. In the latter case no loose wires were used, and thus he had been limited to the amount of aerial that could be attached to the machine itself-about 50 ft . Instead, however, of using balanced aerials, he coupled them to each end of an inductance coil, and increased their effective length to the greatest extent possible without sacrificing efficiency. In the latest form of the apparatus he was using a 6 -in. induction coil with a $\frac{5}{8}$ in. spark gap, fixed at a considerable distance from the apparatus, so as to be away from the petrol tank. Two light brass rods extended from the coil well into the space between the two main planes of the machine, and to one side of the tank, and two $\frac{3}{8}-\mathrm{in}$. brass rods sliding over these and $\frac{5}{8}$-in. apart formed the spark gap terminals. Shunted across the spark gap was a condenser of the Leyden jar type, and an inductance coil consisting of seven turns of No. 14 copper wire wound on a light ebonite drum. This inductance had sliding contacts so that the number of turns used could be varied in the usual manner, in order to tune the two circuits. The two aerial wires were connected to the two ends of the inductance in use, and the aerial circuit was brought into tune with the shunt circuit. A secondary battery of eight or ten volts supplied the primary energy, about 50 or 60 watts being required.
"Two new arrangements have since been adopted, which should greatly enhance the efficiency of the plant. The chief of these is a long light brass tube attached to, but insulated from, one side of the tail of the aeroplane. This acts as a counter capacity, or 'earth', to a long aerial wire on the other side. This aerial starts from the nose of the machine, is carried thence to the extreme outer edge of the main plane, thence back to the tail, and thence to a loose extension, a length of 60 ft . of copper wire trailing behind."

Coming down to earth, another article, 'At the Royal Investiture', described how two Marconi portable wireless telegraph sets were used at the Investiture at Carnarvon of the Prince of Wales. These particular sets were normally employed by the Cumberland Yeomanry and as can be seen from the photograph, consisted of a motor generator and the wireless set itself. It is pictures like this which emphasize the tremendous advances that have taken place in just sixty years.


## World of Amateur Radio

## Morse outmoded?

Since the earliest days of amateur radio, the imminent demise of c.w. operation has been regularly forecast-yet dits and dahs still retain the interest of many amateur operators and account for a significant proportion of all activity. But c.w. has its critics. The notes in this column in May on the possible effects of the proposed F.C.C. changes to U.S. phone allocations brought a strongly contrary opinion from Dr John Irwin, (K6SE/5), of Louisiana State University. He feels that my notes showed a "negative attitude" towards "the switch from c.w. to s.s.b." This, he suggests, is happening all over the world and should be encouraged. "Phone is so much more efficient and interesting and satisfying than code that I have not used c.w. at all for the past two years", he writes. In that time he has worked over 900 different Japanese amateurs on s.s.b., many of them using less than 20 watts. "These Japanese are forced to use and speak English and I think this is a great thing for international fellowship and understanding, and they deserve to be commended for overcoming the severe language barrier. I only wish more Russians used s.s.b. . . . It is a complete misconception to believe that non-U.S.A. amateurs cannot work, do not want to work and do not work in the U.S. phone bands. . . Widening U.S. phone bands will thin out the interference, benefiting all amateurs, the world over.... Single-sideband equipment is now so satisfactory, so potent and so cheap that the present trend from code to voice cannot help but continue; and I'm all for it", he stresses.

Those of us who continue to believe there should be a future for c.w. will disagree with several of Dr Irwin's arguments, but must respect his right to express them-the more so since it now seems pretty certain that there will be an extension of the U.S. phone allocations. But two amateurs chatting on s.s.b. occupy as much frequency space as perhaps 30 or 40 would need for c.w. Where frequencies are under extreme pressure (e.g. 7 and 14 MHz ), surely narrow-band c.w. should be given reasonable priority? On other bands, the decision to opt for c.w. or phone is rightly one for individual amateurs to make.

It is worth noting that c.w. users retain an above average interest in the hobby. A breakdown of 100 British stations worked from G3VA on c.w. (3.5, 7 and 14 MHz ) in recent months showed that about $25 \%$ had been licensed during the past 5 years; about $13 \%$ from 5 to 10 years; $16 \%$ from 10 to 20 years; $18 \%$ from 20 to 25 years; and $28 \%$ over 30 years!

Beyond question s.s.b. is effective-but, because of the peaky nature and wide bandwidth of voice waveforms, c.w. of equivalent power is still a far more effective means of communication, provided that appropriate narrow-band filters are used in the receiver. Essential information can be passed as quickly, and more accurately. So most of us want to see both modes continuing in general use.

## Amateur finds radio "bug"

The recent disclosure, as the result of an Old Bailey trial, that W. H. Borland (G3EFS) of Bromley, Kent, had been responsible for first discovering and then tracking down illegal "bugging" equipment installed about half-a-mile from his home, highlights the continued interest in amateur direction-finding. For almost 20 years, each summer, a series of $D / F$ hunts is organized, culminating in the annual R.S.G.B. National Final. The contests usually take the form of hunting down, over distances up to ten miles, in the course of a single afternoon, two concealed 1.8 MHz transmitters.

## Space communications and amateurs

Amateurs who have been following the progress of the I.T.U. World Administrative Radio Conference on Space Matters in Geneva are concerned at the long-term implications of the extremely strong pressure for microwave frequencies for all forms of space communications. No longer are there any "unwanted" frequencies in this part of the radio spectrum. Amateurs have been disappointed at the apparent lack of liaison between the national amateur radio societies of a number of European countries and their official delegations, who often appear to be virtually unaware
of the amateur service. While it is still expected that some extensions will be granted to amateur space facilities (at present confined to 144 MHz ), a number of proposals, supported by the official U.K. delegation, are unlikely to be approved. The position taken up by the delegations from such countries as France, Norway. Sweden and the U.S.S.R. is contrasted with that of the U.K. where Minpostel invited the R.S.G.B. to nominate a member of its Council (Roy Stevens, G2BVN) to attend the meetings as an official adviser to the U.K. . delegation.

## V.H.F. activities

Several notable tropospheric and sporadic E "openings" were noted during June. TF3VHF, the 70 MHz beacon station in Iceland, was heard in the U.K. on several days. In just over two hours on June 13th, 9HIBL (Malta) worked 13 British stations cross-band $70 / 28 \mathrm{MHz}$ ( 70 MHz is not available in Malta). In a long series of observations on the London 70 cm beacon GB3GEC, two Dutch amateurs, PA0VZL and PA0GDV, have been hearing the station consistently, almost regardless of band conditions. A recent 144 MHz portable contest was won by G. W. Tibbetts, GW3NUE/P, who made 331 contacts. Peter Blair, G3LTF, has resumed 1296 MHz "moonbounce" contacts with W2NFA.

## In brief

The R.S.G.B. National Mobile Rally is at Woburn Abbey on Sunday, August 8th with talk-in stations GB2VHF, G3VHF and GB3RS on 14,70 and 144 MHz . Events will include a trade exhibition, demonstrations of amateur TV, bring-and-buy sale, etc. . . A special station, GB3ESP, will be operated by members of the International League of Esperantist Radio Amateurs during the 56th Universal Esperanto Congress in London from July 31st to August 7th . . . F.C.C. regional offices in America have been asking a number of "Technician" licensees to submit to re-examination; about half turn in their licences without trying. . . . F.C.C. have issued a Notice of Inquiry seeking to determine what improvement (including TV receiver design) could be made to achieve interference-free TV reception; the American Consumers Union intends to report more fully on the susceptibility of TV and hi-fi gear to interference from h.f. transmitters. . . An American amateur, WOWYX, has his home station located at a height of $11,500 \mathrm{ft}$ on Squaw Mountain, Colorado. . . . Increased subscriptions and the aftermath of the postal strike appear to have hit severely recruitment of new R.S.G.B. members; in the three months March to May only 165 new members were elected compared with 545 in the same period in 1970.

Pat Hawker, G3VA

# Personalities 

T. A. Duerden, B.Sc.. Ph.D.. who joined Plessey as manufacturing facilities planning executive just over a year ago. has been appointed general manager (Pentex). Dr. Duerden. who will be primarily responsible for the Pentex electronic telephone exchange business, will be based at the Group's Beeston. Nottingham. factory. A graduate of Manchester University, where he read physics and later received his doctorate. he was head of management services at the Preston Division of British Aircraft Corporation prior to joining Plessey.
G. C. F. Whitaker, F.I.E.E.. F.I.E.R.E. who was for two years on the staff of Yorkshire Television as senior planning engineer followed by a further two years as engineering consultant. has retired. Mr. Whitaker, who is 66. was educated at the Roval Naval Colleges Osborne and Dartmouth. He retired from the Navy in 1928. Re-joining the Navy at the outbreak of war he was, initially engaged on global. long-range h.f. direction finding. followed by a period in the Radio Physics Laboratory of the University of Sydney, where he studied radio location. At the close of hostilities he was re-instated on the Active List and after appointments in the Department of Naval Ordnance and. on two occasions as deputy superintendent of the Admiralty Signals and Radar Establishment, he was promoted to the rank of Captain. His final Naval appointment was on loan to the Australian Commonwealth Government as director of electrical engineering in the Department of The Navy. Melbourne. Victoria. Retiring in 1959. he was employed by Central Rediffusion Services Ltd. and from 1960 to 1967 was chief engineer of Rediffusion television operating the London weekday contract of the I.T.A.

Derek Stanners is appointed U.K. sales manager of Racal Instruments Ltd. of Windsor. Previously on the board of the B \& K Group. with overall marketing control of their instrumentation products
company. Mr. Stanners has also worked for the Plessey Group, at Northampton. He is an enthusiastic radio amateur. His call sign is G3HEJ.

John R. Brinkley, F.I.E.R.E.. A.M.I.E.E.. international manager of mobile radio for the I. T. \& T. Corporation since 1969. has joined Redifon Ltd as an executive director of the company. The Communications and Marine Division of Redifon is to be formed into a subsidiary company and it is intended that Mr. Brinkley should


John R. Brinkley
be its managing director. Mr . Brinkley received his early training with the Post Office. He transferred to the Home Office Communications Directorate in 1942 and six years later joined Pye. He was managing director of Pye Telecommunications Ltd from 1956 until 1966 when he joined Standard Telephones and Cables where he was executive director until his transfer to I.T.T.. the parent company.

Air Chief Marshal Sir Donald Evans, K.B.E.. C.B.. D.F.C.. R.A.F. (Ret'd). has joined Ferranti Ltd in Edinburgh. as a consultant on military aviation matters but will be based at Ferranti's London Office. Millbank Tower. S.W.1. Air Chief Marshal Evans. who is 59. commanded a night fighter trials unit during the war and later the Royal Radar Establishment's Flying Unit. His Air Force service included his appointment as Air Officer Commanding-in-Chief. Technical Training Command (1964-66): as Air Secretary
(1966-67): and as Commandant of the Imperial Defence College (1968-69).
C. J. Kent has joined A.P.T. Electronic Industries Ltd. of Byfleet. Surrey, as sales manager. Mr. Kent joins the company from Advance Electronics Ltd where he was employed for four years as senior sales engineer. He served his apprenticeship with A.E.I. at Trafford Park. Manchester.
J. E. Everitt, M.A.. M.I.E.E., joins the board of Rank Bush Murphy Ltd in the newly created post of director of overseas operations. Mr. Everitt. who is 35 and took his degree in mechanical sciences at Cambridge. joins Rank Bush Murphy from Ultra Electronic Holdings Ltd. of which he was marketing director.
G. Boris Townsend, B.Sc.. Ph.D.. F.I.E.E., F.Inst.P.. for the past six years head of engineering research at Thames Television, has joined the I.T.A. as deputy head of the Engineering Information Service. Dr. Townsend. a graduate of King's College. London. began his career at the General Electric Company where he worked on the development of colour television receivers. He is co-author with P. S. Carnt of the two volumes on colour television published by Butterworth and received his doctorate from London University for a thesis on colour television. In 1963 he joined Rank Cintel as technical manager of the Professional Television Equipment Division. Dr. Townsend was president of the British Amateur Television Club from 1960 to 1965.
A. R. Wilkinson, M.A.. M.I.E.E.. has been appointed technical director of Radiatron Ltd and Radiatron Components Ltd. of Twickenham. Middx. He will be chiefly engaged on development work and market research. Mr. Wilkinson - was formerly principal test equipment engineer with G.E.C. at Aycliffe. Co. Durham.

Ates Electronics Ltd have announced the appointment of Howard Prescott, who will have responsibilities for product marketing and technical liaison on the company's application circuits. Mr. Prescott. started his career with Ultra Electronics Ltd as a student apprentice, and moved to R \& D before joining Air-Tech Ltd as proiects engineer. Immediatelv prior to joining Ates, he was applications engineer with S.G.S. Ltd, where he specialized in linear i.cs.
C. Rhodes Oliver, B.Sc., M.I.E.R.E.. has joined Semiconductor Production Equipment Co. Lid. of West Byfleet. Surrey, as technical director. He will be responsible for all technical aspects
and development of the Centronic product range which includes diffusion furnaces. laminar flow cabinets. profilers. semiconductor ovens and lighting intensity controllers. After the Second World War. which was spent in the New Zealand Air Force working on radar and navigational aids. Mr. Oliver was with Pye Radio and Newmarket Transistors for several years before joining Standard Telephones \& Cables in 1958. This was followed by a period with A.E.I., Brimsdown. as development manager and with R.C.A. at Catania, Sicily.

## BIRTHDAY HONOURS

lew mell in the world of electronics were included in the Queen's Birthdav Honours List. Among those receiving honours are:

## Knights Bachelor

John Allen Clark, Companion I.E.E. chairman \& chief executive. Plessey.
John Henry Davis, chairman and chief executive. Rank Organisation.

## C.B.E.

H. Barker, director, net work planning. Post Office Telecommunications.
Rear Admiral B. J. Castles, F.I.E.R.E., R. Australian Navy. H. W. French, chief inspector. Dept. of Education \& Science.
L. S. Yoxall, chairman. FoxboroYoxall Co.

## O.B.E.

R. E. Burnett, M.A., F.I.E.E., managing director. Marconi Instruments.
R. W. P. Cockburn, controller (admin.) external broadcasting. B.B.C.
W. Nethercot, chairman. Min. Posts \& Telecoms advisory technical sub-committee on wireless interference from industrial apparatus.
L. A. Samson, sales \& service director. Guided Weapons Div.. Brit. Aircrafts Corp.
Wing Commander W. E. Satterthwaite. M.I.F.R.E.. R.A.F.

## M.B.E.

H. Ledger, senior engineer. Plessey Telecomm unications Ltd. Beeston. M. R. Neville for services to the Electrical \& Electronics Industries Benevolent Assoc.
H. J. Plater, asst. manager. studio operations. B.B.C. Television.

## OBITUARY

Lord Reith, under whose guidance broadcasting was started in this country in 1922 by the British Broadcasting Company, of which he was general manager, died in Edinburgh on June 16 th. He was 81. John Charles Walsham Reith. a mechanical engineer by profession. became the first director-general of the B.B.C. when in 1927 it became a public corporation. Although he resigned from the B.B.C. in 1938 he has left his mark indelibly on British broadcasting.

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# seconds make 10,000 hours. 



In nuclear physics you need absolute accuracy and long-term reliability from your electronic tubes. Especially thyratrons. EEV thyratrons can be fired with nano-second precision, with repetition rates of up to 50 kHz due to very rapid deionisation characteristics. Long life - 10,000 hours can be achieved - enables EEV ceramic thyratrons to be bolted into the circuit as with passive components.

EEV thyratrons meet the demands of major nuclear physics applications:
In linear accelerators they can withstand peak inverse voltages up to 20 kV following a pulse, and they give trouble-free operation in oil-filled equipment.

In particle accelerator work missed pulses are rare. Annular current-flow means rapid peak-current switching, too, without risk of arc extinction.

In spark chambers EEV thyratrons will eliminate spurious firing, and jitter can be kept as low as 1 ns . The CXI 154 for example operates over a wide range of H.T. voltages at currents up to 10 kA without significant change in characteristics, so drive units can be used with different chambers -and the low trigger voltage means that simple firing circuits are possible.

So, whether you're concerned about nano-seconds or thousands of hours, specify EEV thyratrons. And remember that EEV also make ignitrons, photo tubes, storage tubes, image intensifiers, vacuum capacitors, spark gaps, RF tubes (like tetrodes for driving RF separators) and magnetrons especially for linear accelerators. Send for details.

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# cut out the coupon and answer your soldering problems 



SK 1

## SOLDERING KIT

In rigid plastic "tool box" containing Model CN - 15 watts - 240 volts miniature iron fitted $\frac{3}{16}{ }^{\prime \prime}$ bit. Spare bits $\frac{5}{32}$ " and $\frac{3}{32}$ ". Reel of resin-cored solder, heat sink, cleaning pad, stand and booklet "How to Solder".


Model CN 240/2
15 watts -240 volts
£170

Fitted with nickel plated $\frac{3}{32}$ "bit and packed in handy transparent box.


## SK2

## Soldering Kit

In polystyrene pack, containing 15 watt miniature soldering iron, 240 volts fitted with $\frac{3}{16}$. bit, 2 spare bits $\frac{5}{32}{ }^{\prime \prime}$ and $\frac{3}{32}{ }^{\prime \prime}$. Coil of resin-cored solder, heat sink, 1A fuse and booklet "How to Solder".

£2.40

GSS Desoldering Tool Model GSS with $\frac{3}{32}{ }^{\prime \prime}$ tip diameter

ES240 D 25 watt soldering iron In transparent display pack, fitted with long life ironcoated bit $\frac{1}{8}$ " diam.

## £1-83

Interchangeable spare bits $\frac{3}{32}$ ', $\frac{3}{16}{ }^{\prime \prime}, \frac{1}{4}{ }^{\prime \prime}$ (extra) available. Improved design to ensure strong and reliable high speed iron. Heats up in 2 minutes.
group I of 40 kHz and an intermediate band of 20 kHz at 60 i.p.s. The record and reproduce heads, made of a wear-resistant material, have an edge voice channel for use with the optional voice logging accessory. An optional automatic tape-threading device is also available. Power requirements are $115 / 230 \mathrm{~V}$ a.c. ( $\pm 10 \%$ ), 48 to 420 Hz single phase. Maximum consumption is approximately 200VA. Bell \& Howell Ltd, Electronics \& Instruments Group, Lennox Road, Basingstoke, Hants.
WW322 for further details

## 50 MHz counter

Model FC50 from Wayne Kerr is a six-digit readout instrument with automatic location of the decimal point. The effective resolution can be increased, in some instances up to eleven digits, by under-ranging. The ranges are 0.1 Hz to 50 MHz and $1 \mu \mathrm{~s}$ to $10^{5}$ seconds, with a count facility to 999,999 . Start and stop can be manual or electrical (or a mixture of the two) and facilities are provided for inhibit, gating, storage and varying the up-dating rate. Clock signals are available for external use and there is an option of

b.c.d. outputs from the six number tubes. The display can be switched to show a 'non-blink' series of completed counts of the run as it proceeds. Acceptable input levels range from 20 mV (r.m.s.) to 100 V , and provision is made for correctly terminating $50 \Omega$ or 75 lines. The Wayne Kerr Co. Ltd, Roebuck Rd, Chessington, Surrey.
WW328 for further details

## Battery-operated soldering iron

The Antex MES 12 soldering iron operates from a 12 volt d.c. supply. Two large crocodile clips on 4.50 m of 2 -core cable provide connection to the battery terminals. The recommended U.K. price is $£ 1.95$. Anglo-Netherlands Technical Exchange Ltd, Mayflower House, Plymouth, Devon. WW 308 for further details

## Gunn oscillator

A Gunn o.scillator made by Mullard gives an output of 35 mW at 10.525 GHz $\pm 20 \mathrm{MHz}$. Type CL8631, it operates at a
fixed frequency over the temperature range -20 to $+50^{\circ} \mathrm{C}$ and can be used satisfactorily with any phase or load mismatch up to a v.s.w.r. of 1.3. It requires a power supply of 8 V , total consumption being less than 2 W . A square flange output mates directly with waveguide size RG-52 (WR90/WG16). The device can replace a klystron oscillator in many applications. Mullard Ltd, Mullard House, Torrington Place, London W.C.1.
WW323 for further details

## Shift registers

The MA86S/87S silicon gate 100/128-bit dual independent shift registers, from GEC Semiconductors, operate from a single t.t.l. level clock and the t.t.l. system noise immunity specification is preserved. The registers can be clocked from zero frequency to more than 3 MHz . All inputs and outputs (including the clock input) are t.t.l. compatible and the device operates from standard voltage levels. Since the registers are completely independent they may be clocked separately. The device is available in a TO-5 style package. GEC Semiconductors Ltd, Freebourne Rd, Witham, Essex.
WW 304 for further details

## Intensifier vidicon

A vidicon camera tube with more than 250 times the sensitivity of a conventional 26 mm vidicon is being produced by the Electron Tube Division of EMI-Electronics. The tube, which employs an intensifier and is designated the Ebitron type 9777 vidicon, is claimed to produce television pictures when illumination is equivalent to half moonlight. The vidicon employs electron-bombardment induced conductivity in the zinc sulphide target with a high sensitivity photocathode. The image section is all electrostatic and the scanning portion similar to a conventional 13 mm magnetic vidicon. The Ebitron can replace existing 26 mm vidicons in c.c.t.v. cameras, the 9777 tube and its coils being nobigger. The 18.2 mm photocathode makes it suitable for use with standard 26 mm vidicon lenses. The weight is 230 g potted,

100 g unpotted.
Typical operating conditions:

## image section

| overall e.h.t. <br> canning section | $14,000 \mathrm{~V}$ |
| :--- | :--- |
| cathode | 0 V |
| $\mathrm{~g}_{1}$ modulator | -30 V |
| $\mathrm{~g}_{2}$ limiter | 300 V |
| $\mathrm{~g}_{3}$ beam focus | 290 to 330 V |
| $\mathrm{~g}_{4}$ vidicon mesh | 500 V |
| axial magnetic focus field 550 V |  |
| output signal | $0.15 \mu \mathrm{~A}$ peak |
|  | white |
| overall sensitivity | $50 \mathrm{~mA} / 1 \mathrm{~m}$ |

The heater requires 90 mA at 6.3 V . EMI
Electronics Ltd, Hayes, Middlesex.
WW 310 for further details

## Thermally controlled soldering iron

A range of lightweight, thermally controlled, soldering instruments has been introduced by Adcola. Known as the Invader, the new models incorporate a proven element combined with a new 'pencil-slim' handle. The rectangular centre heat-shield allows the instruments to be placed on any surface without rolling, and the tool is balanced to keep the working bit clear of the surface. A hanging hook is moulded into the handle. Noryl plastic, used for the handle, does

not readily transmit heat-the company claim the 25 W and 27 W tools are the slimmest available in these powers. The plug-in element can be replaced in 90 seconds. The collet can also accommodate

the complete range of 70 standard and special-purpose bits. Standard Invader models are available for seven voltages$6,12,24,50 / 55,110,220$ and $230 / 250 \mathrm{~V}$. Three collet sizes $-\frac{1}{8} \mathrm{in}, \frac{3}{16}$ in and $\frac{1}{4} \mathrm{in}$-are available, and the recommended price for the largest tool is $£ 1.95$. Elements with bit temperatures between 250 and $410^{\circ} \mathrm{C}$ can be supplied at no extra charge. The temperature of the standard-bit face is $360^{\circ} \mathrm{C}$ controlled to $\pm 10^{\circ} \mathrm{C}$. Adcola Products Ltd, Adcola House, Gauden Rd, London S.W.4.
WW $\mathbf{3 0 7}$ for further details

## Rotary-action switches

A range of low-torque, rotary-action, miniature switches, with a mechanical life in excess of ten million operations, has been introduced by Honeywell. The 900 Series 'V4' switches can operate in clockwise or anti-clockwise direction with no change in operating characteristics, and alternative shaft positions are possible.


Both s.p.c.o. and s.p.d.t. versions are available with 0.187 in quick-connect or solder termination. They are rated at 5 A and 125 or 250 V a.c. Inrush current values should not exceed 10 A . Operating temperature range extends from -40 to $+100^{\circ} \mathrm{C}$. Honeywell Ltd, Charles Square, Bracknell, Berkshire.
WW327 for further details

## Inexpensive $X Y$ plotter

The $X Y$ plotter type PLI00 from J. J. Lloyd Instruments, is suitable for applications where extreme accuracy and high speed are not essential. It is sold as a basic potentiometric assembly with a sensitivity which may be adjusted from 150 to $300 \mathrm{~mm} / \mathrm{V}$. The response speed is approximately $200 \mathrm{~mm} / \mathrm{s}$ and adjustable damping is provided for the servos on both $X$ and $Y$ axes. The amplifiers for both axes are independent, with floating inputs, and a suppressed-zero facility is incorporated which enables the instrument to plot small changes in voltage or current about a given reference level. Calibrated plug-in amplifiers are available to extend the range and enable the instrument to
plot either voltage or current. Each amplifier has a calibrated reference, stepped attenuator and vernier sensitivity control, allowing the gain to be adjusted between $0.5 \mathrm{mV} / \mathrm{cm}$ and $40 \mathrm{~V} / \mathrm{cm}$ or $0.5 \mu \mathrm{~A} / \mathrm{cm}$ and $40 \mathrm{~mA} / \mathrm{cm}$. The accuracy

and repeatability is $\pm 1 \% \pm 1 \mathrm{~mm}$ and the maximum paper size is $254 \times$ 330 mm . Price of plotter only is $£ 124$. The plug in amplifier costs $£ 30$. J. J. Lloyd Instruments Ltd, Brook Avenue. Warsash, Southampton SO3 6HP.
WW314 for further details

## Soldering pencil

A soldering pencil, the MCP from Weller Electric, can be fitted with any of seven iron-plated tips ranging from 0.0 lin 'ṃicropoint' to 0.125 in double flat. Overall

reach is $2 \frac{1}{2} \mathrm{in}$. The element operates at 24 V supplied from its own power pack operating from the 240 V mains. The power unit carries a spring pencil holder, and a cleaning sponge. Price $£ 14.95$; tips 45p each. Weller Electric Ltd. Redkiln Way, Horsham, Sussex.
WW326 for further details

## Encapsulated regulators

The Roband Limpet range of encapsulated series regulators for stabilized power supply systems achieves high dissipation by providing an isolated metal heat transfer surface in one face of each module. The modules, which operate from a single unstabilized d.c. rail or from a battery, give well stabilized outputs up to 55 V or 20 A and have full over-current protection. The output voltage and protection levels
are each preset externally by a fixed resistor, or they can be remotely programmed. The modules fit a standard heat sink extrusion, but can be mounted on any conventional metal surface. A typical 2A unit which measures $47 \times 30 \times 22 \mathrm{~mm}$ can give a stabilized rail set anywhere between 6 V and 24 V with a maximum internal dissipation of 25 W . The cost of

such a unit is $£ 15.50$. Roband Electronics Ltd, Charlwood Works, Charlwood, Horley, Surrey.
WW 306 for further details

## Coaxial connectors

Sealectro have introduced a new range of r.f. coaxial front panel connectors. The 'Kwick Konnect' range provides locking and exhibits a v.s.w.r. of better than 1.30:1 at frequencies up to 18 GHz . Assembly to cables is by crimp or clamp of the outer conductor, and by crimp or solder to the inner conductor. Once mated, it is virtually impossible to break the connection by pulling on the connecting cable. To disconnect a knurled ring is pulled back and the connectors disengage. Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hants.
WW $\mathbf{3 0 9}$ for further details

## T line connectors

Pressac have developed a new system of T line connectors. They are designed to allow electrical accessories to be connected into main wiring harnesses without cutting

the conductors. They can be applied directly to insulated wire without stripping. Each connector has an insulating sleeve which is threaded over the accessory lead
and a brass. contact is crimped to the conductor. The brass connector cuts through the insulation to make an electrical connection. The insulation sleeve is then wrapped around the contact and fixed by an integral latch. The connectors can be supplied either on reels, for machine assembly or loose. Pressac Ltd, Leopold Street, Long Eaton, Nottingham.
WW 311 for further details

## Power transistors

A range of hometaxial silicon power transistors from Ates, suitable for highpower amplifier circuits, employs a structure in which the base region exhibits homogeneous resistivity in the axial direction-i.e. emitter-to-collector-eliminating secondary voltage breakdown within the maximum ratings of the device. The 2N3771 of this TO-3 range provides 150 W output, with $30 \mathrm{~A}\left(I_{C}\right)$ at $50 \mathrm{~V}\left(V_{C B O}\right)$. For 100 V operation, the 2 N 3772 gives 20 A , and the 2 N 377316 A at 160 V . Ates Electronics Ltd, Mercury House, Park Royal, London W.5.
WW $\mathbf{3 0 3}$ for further details

## Reed switch

Reed switch type DRA-291 from F.R. Electronics is capable of switching up to 5 A at 50 VA and up to 1 A at 100 VA . It is

standard size, and has rhodium contacts with low contact resistance. F. R. Electronics Ltd, Wimborne, Dorset, BH21 2BJ.
WW 312 for further details

## Logic level pulse generator

From Grange Electronics (Production) we have received details of a wide-range pulse generator which covers repetition frequencies from 1 Hz to 5 MHz in seven overlapping ranges. Delay and output pulse widths are variable between 100 ns and 100 ms in six overlapping ranges. Additional features include manual and external triggering, a pre-pulse output and simultaneous complementary outputs at t.t.l. levels. The price is $£ 66$. Grange Electronics (Production) Ltd, Stone Lane, Wimborne, Dorset, BH21 1HD.
WW3 18 for further details

## Oven for TO-5 devices

Jermyn's 4ST2 self-regulating oven is designed for devices in TO-5 size packages when lead lengths are restricted to 12.5 mm . Devices having up to eight leads may be accommodated and can be installed without the use of special tools. Ovens having control temperatures of 65 , 80 and $115^{\circ} \mathrm{C}$ are available and will operate in ambient temperatures from -50 up to 50,60 and $100^{\circ} \mathrm{C}$ respectively.


The ovens have no moving parts or electronic circuitry but incorporate a semiconductor heater to provide a self-regulating proportional temperature control. Power requirements are 24 V $( \pm 4 \mathrm{~V})$ a.c. $/$ d.c. 0.6 W (at $25^{\circ} \mathrm{C}$ ambient). Maximum warm-up time from $-55^{\circ} \mathrm{C}$ is 3 minutes. Jermyn Industries, Manufacturing Division, Vestry Estate. Sevenoaks, Kent.
WW319 for further details

## Transient voltmeter

Model 3206 voltmeter from Sintrom Electronics will measure and hold the peak value of a single pulse which has a 10 ms duration or longer. The instrument has a

four-figure digital readout and an accuracy of $1 \%$ of full scale. There are four switched ranges with full-scale values ranging from 10 mV to 19 V . The input impedance is 1 $\mathrm{M} \Omega$. The peak value is held in store until reset. Automatic reset for driving a printer or recorder is provided. The input is floating and is double screened to reject radiated transients. Other models in this range include instruments capable of measuring
pulses up to 30 kV and as short as 50 nanoseconds. Analogue and digital readouts are available. Prices range from $£ 580$. Sintrom Electronics Ltd, 2 Castle Hill Terrace, Maidenhead, Berks.
WW 302 for further details

## Sub-miniature chokes

Cambion's 550-339 sub-miniature radio frequency choke is available in a wide range of inductance values- 0.1 through to $1,000 \mu \mathrm{H}$ in 49 steps. Each choke has a small moulded body 6 mm long and 24 mm in diameter. Cambion Electronic Products Ltd, Castleton, near Sheffield, S30 3WR.
WW313 for further details

## Axial-lead electrolytic capacitors

A series of axial-lead miniature aluminium electrolytic capacitors, type EN 12.12, has been added to the range of ITT single ended miniature capacitors type EN12.35.


The axial-lead versions are available from $1 \mu \mathrm{~F}$ to $4,700 \mu \mathrm{~F}$ rated up to 500 V (dependent on capacitance value). These capacitors have an operating temperature range of -25 to $+70^{\circ} \mathrm{C}$. Plastic sleeves are employed for case insulation. ITT Components Group Europe, Standard Telephones and Cables Ltd, Edinburgh Way, Harlow, Essex. WW316 for further details

## High-frequency counters

Series 7900 counters from Dana Electronics are seven-digit units with an optional eighth digit, and all have optional systems interface units. Sensitivity is 1 mV up to 500 MHz . Three counters typical of the range are the 7910 (to 150 MHz ) at
$£ 595$ (illustrated), the 7920 (to 550 MHz ) at $£ 750$, and the 7960 (to 3 GHz ) at £1395. Dana Electronics Ltd, Bilton Way, Dallow Road, Luton, Beds.
WW324 for further details

## Panel drilling bit

A Bradrad (Type A). from West Hyde Developments. provides panel holes of different sizes. drilling and deburring in a single operation. Two versions are available providing holes of $1 \frac{1}{2}$ to $2 \frac{1}{2}$ inches in $\frac{1}{8}$ in steps, or 36 mm to 60 mm in 3 mm steps.


The bit is made of cobalt 'high speed' steel and has a 12.5 mm diameter shank. Price $£ 23$ plus 35 p postage and packing. West Hyde Developments Ltd, Ryefield Crescent, Northwood Hills, Northwood, Mddx.
WW321 for further details

## Voltage-dependent resistors

A new range of silicon carbide and diffused junction silicon voltage-dependent resistors (varistors) is available from ITT. Silicon carbide voltage-dependent resistors are available in rod or disc form, and can be supplied with leads for direct wiring into position or without leads for direct mounting. These devices have a wide range of applications for voltage control and component protection. Silicon diffused junction varistors are particularly suitable for a very wide range of currents at low

voltage levels. A particular application of this type is for temperature compensation

in semiconductor circuits. ITT Components Group Europe, Resistor Products Sales, Edinburgh Way, Harlow, Essex.
WW 301 for further details

## Variable transformers

Variable transformers from the Zenith Electric Company, in the Variac-Setavolt range, are fuily encapsulated for 200-250 volt operation, covering the current ranges 0.75 to 4 A in 5 sizes. The frequency range is $50-400 \mathrm{~Hz}$ Motorized two-gang and three-gang units are also available. The Zenith Electric Co. Ltd, Wavendon, Bletchley, Bucks.
WW317 for further details

## Tantalum capacitors

The MT series of moulded tantalum capacitors, available from General Instrument (UK) Ltd, are dry sintered anode units, moulded in epoxy resin and not subject to gassing or electrolyte

leakage. Capasitance range is from 0.068 to $47 \mu \mathrm{~F}$ rated up to 50 V . The working temperature can be as high as $85^{\circ} \mathrm{C}$. General Instrument Ltd, Stonefield Way, Ruislip, Middx
WW325 for fuether details

## Improved recording tape

A new family of recording tapes which exhibit increased output and a 4 dB improvement in signal-to-noise ratio with no modification to existing equipment has been developed by the 3M Company. Known as High Energy tapes, they are based on a cobalt-modified ferric oxide formulation. Unlike chromium-dioxide tape, which requires separate circuitry to be switched in, High Energy tape can be used on existing cassette machines without any modification to the standard low-noise bias and equalization levels to give greater undistorted output and an increase in dynamic range from 2 dB at low frequencies to 6 dB at the upper end of the scale. Circuitry designed around the potential performance characteristics of the new tapes could improve reproduction still further. It is expected that the new tapes will be marketed in the U.K. later this year in helical-scan video form, and that broadcast video and audio cassettes will follow. 3M Company, 3M House, Wigmore Street, London W1A IET.

## Miniature trimmer pot

The $\mathrm{T}-200-\mathrm{K}$ single-turn wirewound potentiometer in the Contelec range of trimmers has a knurled plastic-moulded knob, with bifurcated

shaft that pushes into the pot and interlocks with the keyway. It can equally easily be turned by a screwdriver. Power rating is 2 W at $40^{\circ} \mathrm{C}$. Resistance range is $10 \Omega$ to $50 \mathrm{k} \Omega$. Size is $20 \times$ 10 mm . Operating temperature is -25 to $125^{\circ} \mathrm{C}$. The T-200 series is available in eight standard versions, in either bush mounting or printed circuit types. Kynmore Engineering Co. Ltd, 19 Buckingham Street, London WC2. WW 331 for further details

## Zero-voltage switch for thyristors

A low-cost version of the RCA zero-voltage switch for thyristor gate triggering is the CA 3079. It has the same temperature range as the earlier CA3059 ( -40 to $85^{\circ} \mathrm{C}$ ) but the fail-safe, inhibit and over-ride functions are not included. The economy type includes a power supply, allowing operation from an a.c. line of 24 to 277 V at .50 to 400 Hz , a differential sensing amplifier; a zero-crossing detector and a triac gating circuit. The zero-crossing detector, of course, allows thyristor switching at the voltage zeros of the a.c. line, eliminating r.f. interference when used with resistive loads. The circuit is packaged in a 14-lead dual in-line
plastic case. Price is 79 p for $1-24$ and 59 p for 100 up. RCA Ltd, Solid State Division, Sunbury-on-Thames, Middx.
WW 320 for further details

## Stylus balance

The BIB stylus balance model 32 is produced specifically for determining the 'pressure' of modern cartridges and is calibrated in 0.25 g

divisions. It has a non-magnetic base mounted on foam plastic. The cross-bar of the beam has recesses which are mounted on a pair of lowfriction pivot points. Price £1.80. BIB Division, Multicore Solders Ltd, Hemel Hempstead, Herts.
WW $\mathbf{3 3 0}$ for further details

## Printed circuit elements

Conducting elements for wiring semiconductor devices to printed circuits are made by Circuit-Stik Inc. of California. With an adhesive backing, the 1000 and 2000 series of

elements are designed to suit most types of TO-5 and TO-18 packages. The former are drilled to match a 0.1 in grid and the latter undrilled to save space. Available in the U.K. from Bourns (Trimpot) Ltd, 17 High Street, Hounslow, Middx.
WW 329 for further details

## Press-button switches

The Arrow Adapt-a-Switch range of illuminated and non-illuminated press-button switches is based on a small number of components that fit together simply. The actuator can be chosen for momentary or alternate action. Press-in lenses give a range of three shapes-round, square and oblong-in six colours. The standard duty ratings are 5 A at 125 V a.c., 2 A at 250 V a.c., and 5 A at 28 V d.c. Electrical and mechanical life is 100,000 cycles minimum at full rating. Arrow Electric Switches Ltd, Brent Road, Southall, Middlesex. WW 332 for further details

# Real \& Imaginary 

by "Vector"

## On Stopping the Home Fires Burning

I wonder whether you've ever thought of the domestic 'telly' as a lethal instrument? I must confess I hadn't until I read a study of statistics relating to fires in television sets. This paper was written by a member of the Joint Fire Research Organization and the figures quoted give food for thought.

For in 1968 (the last year for which figures were presumably available) 1244 fires occurred in Britain which were directly attributable to the magic box. In 1960, the figure stood at 528 and rose significantly in every subsequent year.

You may well say 'Ah yes, but the number of sets in use increases every year'. True. But other figures given show that the number of fires increased at a considerably higher rate than licences did. In 1965 the incidence of fires to licences was 61.8 per million; three years later it was up to 82.4 per million and after another three years I wouldn't be surprised to find that it had taken another comparable jump. The older the set, the greater the risk, is a logical conclusion and possibly, with the cost of living steadily rising, people are hanging on to their sets longer.

One rather less sombre side is that (taking the 1968 statistics) about $83 \%$ of these fires occurred between 3 p.m. and midnight when someone is likely to be able to initiate prompt action. Only $5 \%$ of the tota-roughly 4.5 per million licencesoccurred between 1 and 2 a.m. Compared with the annual incidence of fires from all causes between these hours, which amounted to some 500 per million dwellings, the number of television fires are chickfeed; but they are nonetheless dangerous, since at that time most people are asleep in bed and totally unprepared for disaster. You may remember that recent fire in a hotel, in which eleven people died. That was attributed to a television receiver
What effect has the advent of colour, with its higher operating voltages, had on the figures? No significant alteration up to 1968, but that doesn't mean much because colour hadn't got going, and even today colour sets are not in wide enough use to make much difference.

How were these fires caused? It was no part of the report's aim to specify and so we don't know. Component breakdown must have played a palt but, to be fair on the manufacturers, by no means all TV receiver fires are started in this way. All
dealers know the old lady who drapes a blanket over the top of the set to let her cat sleep on, and how, by drooping over the back, this (the blanket I mean, not the cat!) can block all ventilation. Tatty do-ityourself flex wiring (often using bare staples) with the lead to the set permanently 'live' is another well-known phenomenon. And again, smoke pouring from the cabinet may panic the householder into calling the fire brigade when in fact no fire, as such, exists. The statistics given seem to indicate the presence of this last factor, for of the 1244 fires quoted for 1968,612 were 'confined to the set' and might therefore have been smoke only-or does a fire brigade have to see flames to record the incident as a fire? As to the remainder, a further 560 were 'confined to the room', while 72 spread to other areas. These 632 were, without doubt, genuine no-nonsense fires, but it would be instructive to know whether outbreaks originating in the mains lead to the set or in its feed wire along the skirting ( where this exists) are classified as television fires, or whether the outbreak must originate in the set in order to qualify.

In the U.S.A. the incidence of fires in TV sets is causing considerable concern. In August 1969 the Federal Government's National Commission began to put the whole question under the microscope and in due course came up with the pronouncement that more than 10,000 such fires occurred annually. Predictably, this was hotly denied (no pun intended!) by the Electronic Industries' Association, which put up a rival figure of 2600 over a fiveyear period. Subsequently, other reports were produced from various sources with figures that fell somewhere between the two extremes.

One such (the 'Jitco') was especially enlightening. This was in essence a tabulation of data supplied by set manufacturers concerning fires reported for each of their models. It did far more than tabulate, however; it also pinpointed the components that were responsible. One startling fact that emerged was that colour sets were forty times more likely to cause fires than black-and-white models. Forty times. That's a fantastic jump.

In the list of delinquent components the line output transformer emerges as the worst offender by a considerable margin ( $29.26 \%$ of the total fire/smoke cases, rising
to $40 \%$ in colour sets). Then come highvoltage components ( $18.1 \%$ ), the receiver on-off switch ( $12.7 \%$ ), the yoke ( $7.4 \%$ ), controls ( $5.9 \%$ ) and so on through seven more groupings, ending with fuses at $0.4 \%$. Nothing much to surprise the British service engineer here, I fancy. Understandably, fires occurred in chassis runs-that is, if a given component was fitted which subsequently proved to be unequal to its job, an epidemic of fires would be experienced with the particular model that embodied the component.

Now, to judge from a comprehensive report on the subject in Electronics, the United States can scarcely be set up as a pattern upon which to model future British procedure. The bible in the matter of standards seems to be the Underwriters' Laboratories UL492, which runs to 402 paragraphs and which is continuously being updated. But apparently there is no legal obligation to conform to it and it is left to individual cities to decide whether sets used within their boundaries should carry the U.L. stamp. Only three cities insisted on this at the time of the survey (August 1970) and so many manufacturers just don't bother with it. The U.L. standards, it is stated, are not so much those which ensure public safety as ones which the manufacturers can conveniently work to. One example cited is the permissible leakage between case and earth which is 5 mA sufficient to pack a nasty wallop; efforts are now being made to reduce this to 0.5 mA . Again, the permitted level of X-ray radiation from TV sets has been set at 0.5 milliröntgen/hour at 5 cm , but not because this gives a good margin of safety; it is merely a level that manufacturers can conveniently meet. Recently, however, some improvement has been effected; from January 1st, 1971, all sets have had to conform to this level even if all controls are maladjusted to 'worst case' and component failure increases emission.

Signs are not wanting, in fact, to indicate that American television manufacturers are at last treating the fire hazard much more seriously than formerly. This may reflect an improved sense of social conscience. On the other hand the more cynical might think that it stems from a test case in the U.S. Courts concerning a man who died in a TV-originating fire. The receiver manufacturers were ordered to pay $\$ 212,000$ dollars compensation to the man's family. If this establishes a precedent as to where responsibility lies, it could make for an expensive future for television manufacturers.

Returning now to the British scene, one benefit from our delayed entry into colour is that we have a breathing space before colour receivers become the rule rather than the exception. This gives us opportunity to benefit from American mistakes.

For the information contained in the above I am indebted to:- 'Fires in television sets', S. E. Chandier, Fire, Sept. 1970. 'Customer hazards: why they happen', and 'Customer hazards: how they can be fixed', Electronics, Aug. 3, 1970.


# Super IC-12 



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[^8]
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SPECIFICATIONS-Number of transistors: 16 plus 20 in I.C. Tuning range: 87.5 to 108 MHz Capture ratio: 1.5 dB . Sensitivity: $2 \mu \vee$ for 30 dB quieting: $7 \mu \mathrm{~V}$ for full limiting. Squelch level $20 \mu \mathrm{~V}$. A.F.C. range: $\pm 200 \mathrm{KHz}$. Signal to noise ratio: $>65 \mathrm{~dB}$. Audio frequency response : $10 \mathrm{~Hz}-15 \mathrm{KHz}( \pm 1 \mathrm{~dB})$. Total harmonic distortion : $0.15 \%$ for $30 \%$ modulation. Stereo decoder operating level: $2 \mu \mathrm{~V}$. Cross talk: 40 dB . Output voltage: $2 \times 150 \mathrm{mV}$ R.M.S.
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| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | Amps | Weight | Size cm. Se | econdery Windings | $1-24$ | 25-99 | No |
| Na. | 12 V 24 V | 16 az |  |  |  |  | Np |
| 111 | $\begin{array}{ll}0.5 & 0.25 \\ 1.0 & 0.5\end{array}$ |  | $7.6 \times 5.7 \times 4.4$ $8.3 \times 5.1 \times 5.1$ | 0.12 at 0.12 V at $0.5 A \times 2$ | ${ }_{0} 0.88$ | 0.81 | 22 |
| 213 | 1.005 | 1 +1 1 | $8.3 \times 5.1 \times 5.1$ $7.0 \times 6.4 \times 5.7$ | - $0.12 V$ at $0.5 A \times 2$ | 1.88 1.16 | 1.87 1 | 22 |
| 18 | 2 | 24 | $8.3 \times 7.0 \times 7.0$ | $0-12 V$ at $2 A \times 2$ | 1.62 | 1.50 | 36 |
| 70 | 6 | 312 | $10.2 \times 7.6 \times 8.6$ | $0-12 \mathrm{~V}$ at $3 A \times 2$ | 1.95 | 1.81 | 42 |
| 72 | 105 | 63 | $7.9 \times 10.8 \times 10.2$ | $0-12 \mathrm{~V}$ at 5A $\times 2$ | 2.56 | 2.37 | 52 |
| 17 | 168 |  | $12.1 \times 9.5 \times 10.2$ | $0-12 \mathrm{~V}$ at $8 \mathrm{~A} \times 2$ | 3.95 | 3.16 | 52 |
| 15 | $20 \quad 10$ | 116 | $12.1 \times 11.4 \times 10.2$ | $0-12 \mathrm{~V}$ at $10 \mathrm{~A} \times 2$ | 5.03 | 4.70 | 67 |
| 187 | $30 \quad 15$ | 1612 | $13.3 \times 12.1 \times 12.1$ | $0.12 \mathrm{Vat} 15 \mathrm{~A} \times 2$ | 9.28 | 8.58 | 82 |
| 30 VOLT RANGE |  |  |  |  |  |  |  |
| Re | Amps. | Weightis az. | Size cm. | Secondory Tops | Qty | $\begin{aligned} & \text { Qty. } \\ & 25-99 \end{aligned}$ | $\begin{aligned} & \text { P.P. } \\ & \text { each } \end{aligned}$ |
| No. |  |  |  |  | ${ }_{0}^{6}$ | ¢ 6 | each |
| 112 | 0.5 |  | $8.3 \times 3.7 \times 4.9$ | 0-12-15-24-30V | 1.18 | 1.10 | 36 |
| 79 | 1.0 2.0 |  | $7.0 \times 6.4 \times 6.0$ $8.9 \times 7.0 \times 7.6$ | ". ${ }^{\text {.", }}$ | 1.75 | 1.63 | 36 |
| 20 | 2.0 3.0 |  | $8.9 \times 7.0 \times 8.6$ $10.2 \times 8.9$ | ". ", | 2.16 | 1.95 | 42 |
| 21 | 4.0 | 60 | $10.2 \times 9.5 \times 8.6$ |  | 2.56 | 2.37 | 52 |
| 117 | 6.0 |  | $12.1 \times 9.5 \times 10.2$ | ., " | 3.79 | 3.51 | 52 |
| 89 | 10.0 | 12 | $14.0 \times 10.2 \times 11.4$ |  | 6.21 | 5.74 | 67 |
| Ref. | Amps. | Weight | m. | 50 VOLT RANGE | $\begin{gathered} \text { Qty } \\ 1-24 \end{gathered}$ | $\begin{aligned} & \text { Qty } \\ & 25-99 \end{aligned}$ | $\underset{\substack{\text { P.P. } \\ \text { eoch } \\ \text { Nat }}}{ }$ |
|  |  |  |  |  |  |  | Np |
| 102 | 0.5 | 111 | $7.0 \times 7.0 \times 5.7$ | 0-19-25-33-40-50V | 1.16 | 1.07 | 30 |
| 103 | 1.0 | 210 | $8.3 \times 7.3 \times 7.0$ | , " | 1.69 | 1.57 | 36 |
| 104 | 2.0 |  | $10.2 \times 8.9 \times 8.6$ | " ., | 2.34 | 2.16 | 42 |
| 105 | 3.0 |  | $10.2 \times 10.2 \times 8.3$ | " " | 3.18 4.20 | 3.89 | 52 |
| 106 | 4.0 |  | $12.1 \times 17.4 \times 10.2$ | " ${ }^{\text {, }}$ | 4.20 | 3.89 5.74 | 52 67 |
| 107 | 6.0 |  | $12.1 \times 11.1 \times 13.3$ | " ${ }^{\prime}$ | 6.21 8.10 | 7.49 | 97 |
| 118 119 | 8.0 10.0 |  | $13.3 \times 13.3 \times 12.1$ $16.5 \times 11.4 \times 15.9$ |  | 8.10 10.15 | 7.49 | 97 |
| 119 | 10. |  | Size cm. | 60 VOLT RANGE | $\begin{aligned} & \text { Qty. } \\ & 1.24 \end{aligned}$ | $\frac{\text { Qty }}{25-99}$ | $\underset{\substack{\text { each } \\ \text { each }}}{\text { Nop }}$ |
|  | Amps. |  |  |  |  |  |  |
| No. 124 |  | ${ }^{16} \mathrm{OL}$ |  |  |  | ${ }_{109}$ | ND 36 |
| 124 126 | 0.5 1.0 | 2 3 | $8.3 \times 9.5 \times 6.7$ $8.9 \times 7.6 \times 7.6$ | 0-24-30-40-48-60V | 1.64 1 | 1.52 | 36 |
| 127 | 2.0 | 56 | $10.2 \times 8.9 \times 8.6$ | " | 2.56 | 2.37 | 42 |
| 123 | 4.0 | 10.6 | $11.4 \times 9.5 \times 11.4$ |  | 5.03 7.28 | 4.65 | 87 |
| 120 | 6.0 | 1612 | $13.3 \times 12.1 \times 12.1$ $16.5 \times 12.7 \times 16.5$ | " ${ }^{\text {", }}$ |  |  | 82 |
| 122 | 10.0 | 232 | $16.5 \times 12.7 \times 16.5$ |  | 12.05 | 11.15 |  |
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|  |  |  |  |  | $1-24$ | 25-99 |  |
| ${ }^{\mathrm{No}}$ | 1.5 | [bol ${ }^{\text {az }}$ | $7.0 \times 6.0 \times 6.07$ |  | 1.17 | 1.08 | N 30 |
| 5 | 4.0 | 311 | $10.2 \times 7.0 \times 8.3$ | Please note, these | 1.77 | 1.64 | 42 |
| 86 | 6.0 | 512 | $10.2 \times 8.9 \times 8.3$ | units do not in- | 2.67 3.04 | 2.47 2.82 | 52 |
| 146 | 8.0 2.5 | 64 114 | $8.9 \times 10.2 \times 10.2$ $13.3 \times 10.8 \times 12.1$ | J clude rectifiers | 3.04 4.52 | 4.18 | 67 |

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| CA3047 | 21.37 | MC724P | ${ }^{68 p}$ | SN74160 | 81.80 |
| CA3048 | 22.04 | MC780P | 22.47 | SN74161 | 81.80 |
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| FCL101 | 22-80 | SN7411 | 25p | UA741C | 87p |
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| 600 PLV | 1A | 50p | 200 P | PIV 4A | 75p |
| 50 PrV | 2A | ${ }^{5.50}$ | 400 P | PIV 4A | 80 p |
| 100 PIV | 2A | ${ }^{60} \mathrm{p}$ | 30 P | PIV 6A | ${ }^{68 p}$ |
| 200 PIV | 2 A | 878 | 100 P | PIV 6A | 75 p |
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| 50 PIV | 4A | ${ }^{60 p}$ | 400 | PIV 6A | \&1.10 |
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| 4003200 P | Piv |  | 10p | 12p | 229 |
| 4004400 P | PIV |  | 10p | 12p | 25p |
| 4005600 P | PIV |  | 12. | 15 p | 26p |
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| 1N34A | 10. |  | 12p | GJ7M | 37D |
| 1 N 914 | 7 p | BAX13 | 12p | OA5 | 17p |
| IN916 | 70 | BAX16 | 12p | OA6 | 12p |
| AAI19 | 7 D | BAY31 | 7 p | OA10 | 82p |
| AA129 | 10 p | BAY38 | 25p | OA9 | 10 p |
| AAZ13 | 10 p | BY100 | 15p | OA47 | 7 D |
| AA715 | 120 | BY103 | 22p | OA70 | 7p |
| AAZ17 | 12p | BY122 | 37p | OA73 | 109 |
| BA100 | 15p | BY124 | 15p | OA79 | ${ }^{8 p}$ |
| BA102 | 22 p | BY126 | 15D | OA81 | 7 D |
| BA110 | 32 p | BY127 | 17 p | OA85 | 7 p |
| Balli | 27 p | BY164 | 67p | OA90 | 7 p |
| BA112 | 70p | BY210 | ${ }^{35}$ p | OA91 | 7 p |
| BA115 | 7 D | BYZ11 | 32 D | 0 A95 | 7 D |
| BA141 | 32p | BYZ12 | ${ }^{30} \mathrm{p}$ | OA200 | 100 |
| BAl42 | 32 p | BYZ13 | 25D | OA202 | 10 D |
| BA144 | 12p | BYZ16 | 40p | OA210 | 179 |
| BA145 | 20 p | FST3/4 | 22p |  |  |


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| PIV | 50 | 100 | 200 | 300 | 400 |
| 1 A | 25p | 27 p | 37p | 40p | 47p |
| 4 A |  | 470 | ${ }^{55}$ | 57p | 77p |
| 5 A | - | 65p | ${ }^{65 p}$ |  | 75 p |
| 7 A | - | 55D | 65 p | - | 97p |

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2526
3015
30017
3018
30 FF
$30 \mathrm{FL1}$
30 FL 12
30 FL 14



| 25 L 6 | 45 D EL91 |
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|  | Add 12p in 2 for postage |
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| 400 MW | NER DI | 10 watt |
| :---: | :---: | :---: |
| 3.3-33 Volt | 2-4-100 Volt | 3.9-100 Volt |
| 15p each | 20p each | 25p each |


| 2N3055 | 75p | AF239 | 87p |
| :---: | :---: | :---: | :---: |
| $\begin{gathered} 25+ \\ 100+ \end{gathered}$ | $\begin{aligned} & 82 \mathrm{p} \\ & 50 \mathrm{p} \end{aligned}$ | $25+$ | 32p |
|  |  | $100+$ | 28p |
|  |  |  | 20 p |
|  |  | $1000+$ | 200 |
| BCl13 | 15D | BCl 48 | 11p |
| $25+$ | 18p | $25+$ | 9 p |
| $100+$ | 12 D | 100+ | 8 D |
| $500+$ | 10p | $500+$ | 7 p |
| $1000+$ | 8D | 1000+ | ${ }^{80}$ |
| BYZ13 | 25p | BC168C | 15p |
| $25+$ | 20 p | $25+$ | 12p |
| $100+$ | 17p | $100+$ | 10 p |
| $500+$ | 15p | $500+$ | 8p |
| $1000+$ | 13p | $1000+$ | 6p |
| BC107/8/9 | 10p | Be169C | 15p |
| $25+$ | 9p | $25+$ | 129 |
| $100+$ | 87 | 100+ | 10D |
| $500+$ | 7 p | $500+$ | 8 p |
| $1000+$ | 60 | $1000+$ | 69 |
| 0071 | 12D | AD161/AD162 | 35 p |
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| 9N74107 | 0.45 | 0.40 | 0.85 | SL403A | 2.12] |  | Sheets 5 p per | type |
| 8N74121 | 0.90 | 0.85 | 0.80 | 3 watt amp. |  |  |  |  |
| SN74151 | 1.10 | 1.00 | 0.00 | sL701C | 1.45 |  |  |  |
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|  |  |  |  | P.C. | 2 | 100 | 0.60 Therm | apath |
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|  |  |  |  | For PA246 | 4 | 200 | $0.75 \quad 2$ |  |
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131 g-inte nand
131 -input NAND gate
151 Expandable dual
2-wide
AND-OR-INV
2-int
$161 \begin{gathered}\text { Duate } \\ \text { Dual } \\ \text { AND-OR-INe } \\ \text { 2-ingut }\end{gathered}$
$171 \begin{gathered}\text { Expandable } \\ \text { E-input AND-OR } \\ \text { 2ate }\end{gathered}$
2-inPut AND-OR-
INVERT gate
1814 wide $\begin{aligned} & \text { 2-input } \\ & \text { AND-OR-INVERT }\end{aligned}$ gate
191 Quadruple 2-input
201 Quadruple 2-input with open
collecter output
221 Hex inverter
${ }_{241}^{231}$ 2.tite binary fult-add

7400 20p 16p 14p 7410 20p 16p 14p $\begin{array}{llll}7420 & 20 p & 16 p & 14 p \\ 7430 & 20 p & 16 p & 14 p\end{array}$ 7440 24p 20p 17p 7450 20p 16p 14p 7451 20p 16p 14p 7453 20p 16p 14p 7454 20p 16p 14p 7402 20p 16p 14p
$7401 \quad 20 \mathrm{p} \quad 16 \mathrm{p} \quad 14 \mathrm{p}$ $\begin{array}{llll}7481 & 23 \mathrm{p} & 21 \mathrm{p} & 18 \mathrm{p} \\ 7480 & 67 \mathrm{p} & 56 \mathrm{p} & 48 \mathrm{p} \\ 7482 & 87 \mathrm{p} & 73 \mathrm{p} & 62 \mathrm{p}\end{array}$ $7483 \quad 11.32 \quad 11.16 \quad 11.00$

## 271 Hex inverter with

$281 \begin{gathered}\text { Output } \\ \text { Beco decimal } \\ \text { decoder TTL }\end{gathered}$
$291 \begin{aligned} & \text { Quadruple } \\ & \text { NAND } \\ & \text { gate with }\end{aligned}$
7442 61.16 94p 81p
NAND gate with 341 Quadruple e -input exement
351 Schement
361 Sxess 3 trigger.
Execimal 371 Excess 3 gray to
38। Quad 2-input positiv
ANO gate
pole output
391 $\begin{gathered}\text { pole output } \\ \text { Quad } \\ \text { AND -input positive }\end{gathered}$
7408 25p 21p 18p
AND gate Op
collector
FLY 101 Dual $\begin{gathered}\text { colinput } \\ \text { coll }\end{gathered}$
FL $101 \mathrm{~J}-\mathrm{K}^{\mathrm{x} p \text { flip }}$ flop
121 Dual Jl-K master
131 Duave flip-flop
131 Dual J-K master.

Part No.
141 Dual D
151 Ouad bistable latch
161 Decade counter
171 Divide-by- 12 counter
191
201 A-bit shift register
Synchronous do down
$201 \begin{gathered}\text { Synchronous up } \\ \text { 4-bir decade }\end{gathered}$
4-bir decade
$211 \begin{gathered}\text { Syne mode control } \\ \text { Synchronous sup down } \\ \text { abit bin }\end{gathered}$
4-bit binary
with one line

 241 Synchronous up down 4-bit dec
counter
As above)
$251 \begin{gathered}\text { (As above) } \\ \text { counter }\end{gathered}$$74192 \quad € 1.74<1.45<1.25$
2615 -bit shife register

27 Dual J-K master-slav flip-flop with preser
and clear
301 Dual quadruple
FLK101 Monostabie multi-
74107 52p 43p 36p 101 Bybator decoder and nixio
driver $\quad 74141 \quad$ \& $1.22 \quad$ \&1.02 87p

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|  |  | Collector) | 0.250 | 0.200 | 167 |
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|  |  | Collector) | 0.250 | 0.20 | 0.167 |
| DM7404N | (SN7404N) | Hex Inverter | 0.2 | 0.225 | .188 |
| DM7405N | (SN7405N) | Hex Inverter (Open Collector) | 0.27 | 0.225 | 0.188 |
| DM74ION | (SN7410N) | Triple Three-Input Gate | 0.250 | 0.200 | 0.167 |
| DM7420N | (SN7420N) | Dual Four-Input Gate | 0.250 | 0.200 | 0.167 |
| DM7430N | (SN7430N) | Eight-Input Gate | 0.250 | 0.200 | 0.167 |
| DM7440N | (SN7440N) | Dual Four-Input Buffer | 0.250 | 0.200 | 0.167 |
| DM7450N | (SNT7450N) | Expandable Dual AND-ORINVERT Gate | 0.250 | 0.200 | 0.167 |
| DM7451N | (SN745IN) | Dual AND-OR-INVERT Gate | 0.250 | 0.200 | 0.167 |
| DM7453N | (SN7453N) | Expandable AND-OR-INVERT | 0.250 | 0.200 | 0.167 |
| DM7454N | (SN7454N) | AND-OR-INVERT Gate | 0.250 | 0.200 | 0.167 |
| DM7460N | (SN7460N) | Dual Four-Input Expander | 0.250 | 0.200 | 0.167 |
| DM7472N | (SN7472N) | J-K Master Slave Flip Flop | 0.325 | 0.263 | 0.221 |
| DM7473N | (SN7473N) | Dual J-K Flip Flop | 0.525 | 0.417 | 0.350 |
| DM7474N | (SN7474N) | Dual D Flip Flop | 0.450 | 0.363 | 0.300 |
| DM7476N | (SN7476N) | Dual J-K Flip Flop with Preset and Clear Inputs .. | 0.563 | 0.450 | 0.375 |
| DM7486N | (SN7486N) | Quad Exclusive-OR Gate | 0.575 | 0.488 | 0.425 |
| DM74107N | (SN74I07N) | Dual J-K Flip Flop with Vce and |  |  |  |
|  |  | GND on Corners | 0.525 | 0.417 | 0.350 |

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|  |  |  | 1-24 | $25+$ |  |  |  | $1-24$ |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | $\because$ | .. | $\because 10 \mathrm{l}$ | ${ }_{8 \rho}^{8 p}$ | ${ }_{2}^{2 N} 1132$ |  | $\because$ | - ${ }^{25 p}$ | ${ }_{13}^{21 p}$ |
| BC BC 109 109 | $\cdots$ | . | $\because 10 \mathrm{p}$ | ${ }^{8 p}$ | ${ }^{2} \mathbf{2 N 1 3 0 4}$ |  |  | 22, | 18, |
| BC 114 | $\because$ | $\because$ | $\because 14 \mathrm{p}$ | ${ }_{12}{ }^{2}$ | ${ }_{2}{ }^{2} \mathrm{~N} 1653$ |  |  | ${ }_{\text {20, }}^{22}$ | 188p |
| BC 115 | $\because$ | $\because$ | $\therefore 165$ | 14 Pp | 2 N 2193 |  |  | ${ }^{27 p}$ | ${ }^{20}$ |
| ${ }_{\text {BC }}{ }^{\text {BC }} 116$ 16 | $\because$ | $\because$ | $\because 19$ | ${ }^{163}$ | 2N2218 |  |  | ${ }_{23+\mathrm{p}}^{23 \mathrm{p}}$ | 17tp |
| BC 118 | $\because$ | $\because$ | $\because 10 \mathrm{p}$ | cisp | 2 N 2221 |  |  | ${ }_{23}{ }^{23}$ | ${ }_{178}$ |
| ${ }^{\text {BC }}$ BC 125 | . | . | $\cdots{ }^{20} 5$ | 17 P | ${ }^{2} \mathbf{N} 2222$ | $\cdots$ |  | ${ }^{23,5}$ | ${ }^{12}$ |
| BC 147 | $\cdots$ | $\because$ | - ${ }^{\text {a }} 10$ | , | ${ }_{2} \mathrm{~N} 2369 \mathrm{~A}$ |  |  | ${ }_{18 p}$ | ${ }_{13+}$ |
| ( BC ${ }_{\text {BC }} 1488$ | $\because$ | $\because$ | ${ }_{13}{ }^{\text {p }}$ | ${ }_{18}^{8 p}$ | ${ }_{\text {2 }}^{2}$ |  |  | ${ }^{27 p}$ | ${ }^{209}$ |
| ${ }^{\text {BC }} 153$ | $\because$ | $\because$ | $\because 18$ 18p | ${ }^{16+p}$ | 2 N 2907 |  |  |  | 22, ${ }^{\text {ap }}$ |
| ¢CC ${ }_{\text {BC }} 178$ | $\because$ | $\because$ |  | 1719 | - ${ }_{\text {2N2924 }}$ |  |  |  | 9 |
| BC 182 | $\because$ | $\because$ | $\cdots{ }^{\text {a }}$ | 2p | ${ }_{2} \mathrm{~N} 2926$ |  |  | ${ }^{\text {c/ }}$ | ${ }_{7}{ }^{\text {P1P }}$ |
| ${ }_{\text {BC }} 183$ | $\cdots$ |  | 918 | ${ }^{8,1}$ | ${ }^{2} \mathbf{2 N 3 0 1 1}$ |  |  | ${ }^{15}+\mathrm{p}$ | ${ }_{12}^{12}$ |
|  | $\because$ | $\cdots$ | $\because{ }^{\circ} 119$ | 910 | ${ }^{2} \mathbf{2 N 3 0 5 3}$ |  |  | ${ }^{18 p}$ | ${ }^{12+8}$ |
| BCY 59 | $\because$ | . | $\because 27$ | ${ }^{22}$ | 2 N 3133 |  |  |  | ${ }_{178}^{127 p}$ |
| ${ }_{\text {BCY }} \mathrm{BCY} 70$ | $\because$ | . |  | ${ }_{\substack{121 \\ 15 \\ 15}}^{18}$ | 2N3134 |  |  |  | ${ }^{1818} 18$ |
| BCY 72 | . | $\cdots$ | 12\% | 10 P | ${ }^{2} \mathrm{~N} 3136$ |  |  | $\because 278$ | 22 P |
| ${ }^{\text {BF }} 167$ | $\because$ | $\because$ | - $\because 8$ | 17p | 2N3390 |  |  |  | ${ }_{17}^{2519}$ |
| BF ${ }_{\text {BF }} 173$ | $\cdots$ | $\cdots$ |  | ${ }^{20}$ | 2 N 3391 A |  |  | -. $22+$ P | 19 p |
| BF 180 2N697 | $\because$ |  | - | ${ }_{128}^{28 p}$ | ${ }_{\text {2N3392 }}^{2 N}$ |  |  |  | ${ }_{12 \mathrm{p}}^{11}$ |
| ${ }^{2} \mathbf{2 N 6 9 9}$ | $\cdots$ |  |  | ${ }^{22}$ | 2 N 3414 |  |  | $\cdots 14 p$ | 12+p |
| ${ }^{2} \mathbf{2 N 7 0 6}$ | $\because$ |  |  | ${ }_{14 \mathrm{p}}$ | 2 N 3415 |  |  |  | 15p |
| ${ }^{2 N} 122$ | .. |  | 79 | ${ }_{\text {ctip }}$ | ${ }_{2}^{2 N 3643}$ |  |  | $\cdots 27$ pp | ${ }^{221} \mathrm{p}$ |
| 2 N 988 |  |  | 42 ${ }^{\text {P }}$ | ${ }^{36}$ | 2N3646 |  |  | ${ }^{26,5}$ |  |
| 2N299 | $\cdots$ | . | 178 | $12 \pm$ P | 2 N 4392 |  |  | 11.42 | 6120 |
| 2N930 | . |  |  | 12tp | 2N4393 |  |  | 61.42 |  |

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| :---: | :---: | :---: | :---: | :---: |
|  |  |  | $\underline{E}$ | $f$ |
| SSC41B | 6 | 200 | 0.865 | 0.693 |
| *SSC40B | 6 | 200 | 1.016 | 0.814 |
| SSC4ID | 6 | 400 | 1.146 | 0.915 |
| *SSC40D | 6 | 400 | 1.302 | 1.050 |
| SSC46B | 10 | 200 | 1.167 | 0.932 |
| *SSC45B | 10 | 200 | 1.318 | 1.050 |
| SSC46D | 10 | 400 | 1.520 | 1.218 |
| *SSC45D | 10 | 400 | 1.675 | 1.398 |
| SSC5IB | 15 | 200 | 1.201 | 0.966 |
| *SSC50B | 15 | 200 | 1.352 | 1.075 |
| SSC5ID5 | 15 | 400 | 1.806 | 0.882 |
| *SSC50D | 15 | 400 | 1.953 | 1.562 |
| *SSC61B | 25 | 200 | 2-108 | 1.701 |
| *SSC60B | 25 | 200 | 2.297 | 1.822 |
| *SSC6ID | 25 | 400 | 3.008 | 2.402 |
| *SSC60D | 25 | 400 | 3-192 | 2.541 |
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| IN4001 |  | . | . | . | . | . | 50 | 7p | 6 p | 4p |
| IN4002 |  | . | . | . | . | . | 100 | 8p | 7p | 41p |
| IN4003 |  | . | . | . | $\cdots$ | . | 200 | 10p | 9 p | 5p |
| IN4004 | . | . | . | . |  |  | 400 | 10 p | 9p | 5p |
| 1N4005 | $\cdots$ | . | $\cdots$ | . | $\cdots$ | $\cdots$ | 600 | 12p | 10p | 7p |
| 1N4006 | $\cdots$ | . | . | . |  | . | 800 | 14p | 12p | 9p |
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VRC.19X Trans-ceiver, $150-170 \mathrm{Mc} / \mathrm{s}, 2$ Channel, 20 Watts, Output $12 / 24 \mathrm{~V}$ d.c. operation. General Electric Transmitter, $410-419 \mathrm{Mc} / \mathrm{s}$, thin line tropo scatter
system, with antennae. W.S. Type 88, Crystal controlled, $40-48 \mathrm{Mc} / \mathrm{s}$. W.S. Type system, with antennae. W.S. Type 88, Crystal controlled, $40-48 \mathrm{Mc} / \mathrm{s}$. W.S. Type
$\mathrm{HF}=156, \mathrm{Mk}$. II, Crystal controlled, $2.5-7.5 \mathrm{Mc} / \mathrm{s}$. W.S. Type 62 , tunable, $1.5-12$ $\mathrm{HF}-156$, Mk. II, Crystal controlled, $2.5-7.5 \mathrm{Mc} / \mathrm{s}$. W.S. Type 62 , tunable, $1.5-12$
$\mathrm{Mc} / \mathrm{s}$. C.44, Mk. II, Radio Telephone, Single Channel, $70-85 \mathrm{Mc} / \mathrm{s}$, 50 watts,

 watt output, 110 V or 230 V input. STC Tx/Rx Type 9X, TR1985; RT1986;
TR1987 and TR1998, $100-156 \mathrm{Mc} / \mathrm{s}$. TRC-1 Tx/Rx, Types T.14 and R.19, TR1987 and TR1998, $100-156 \mathrm{Mc} / \mathrm{s}$. TRC-1 Tx/Re, Types T.14 and R. 19 ,
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455 Tx/Rx. Directional Finding Equipment CRD 6 and FRD. 2 complete 455 Tx/Rx. Directional Finding Equipment CRD. 6 and FRD. 2 co
Sets available and spares. Complete system with full set of Manuals.

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each
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TEST SET TS-147C: Combined signal generator, frequency meter and power meter for $8500-9600 \mathrm{Mc} / \mathrm{s}$. CW or FM signals of known freg, and power or measurement of same. Signal Generator: O/put - to - 85 dbm . Trans-mission-FM, PM, CW. Sweep Rate-0-6 Mc/s per microsec. Deviation-0$40 \mathrm{Mc} / \mathrm{s}$ per sec. Phase Range- $3-50 \mathrm{microsec}$. Pulse Repetition Rate-to 4000 pulses per sec. RF Trigger for Sawtooth Sweep-5-500 watts peak. Sawtooth Sweep-Positive polarity, $10-50 \mathrm{~V}$ peak. $0.5-20$ microsec duration at $10 \%$ max. amplitude, less than 0.5 microsec rise time between $90 \%$ and $10 \%$ $10 \%$ max. amplitude, less than 0.5 microsec rise time between $90 \%$ and $10 \%$ max.
$+2.5 \mathrm{Mc} / \mathrm{s}$ per sec. absolute, $+1.0 \mathrm{Mc} / \mathrm{s}$ per sec. for freq. increments of less than $60 \mathrm{Mc} / \mathrm{s}$ relative, $\pm 1.0 \mathrm{Mc} / \mathrm{s}$ per sec. at $9310 \mathrm{Mc} / \mathrm{s}$ per sec. calibration point. Accuracy measured at $25^{\circ} \mathrm{C}$ and 60 hurnidity. Power Meter: Input: +7 to +30 dbm . Output -7 to -85 dbm . Price: $£ 75$ each $+£ 1$ carr.
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MARCONI DEVIATION TEST SET TF-934: $2.5-100 \mathrm{Mc} / \mathrm{s}$ (can be extended up to $500 \mathrm{Mc} / \mathrm{s}$ on Harmonics). Dev. Range $0-75 \mathrm{Kc} / \mathrm{s}$ in modulation range $50 \mathrm{c} / \mathrm{s}-$ $15 \mathrm{Kc} / \mathrm{s} .100 / 250 \mathrm{~V}$. a.c. $£ 45$ each, $£ 1.50$ carr.
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"OMRON" OCTAL BASE. A.C. mains. $2 \times 5 \mathrm{amp}$. A.E. Perspex enclosed, plug in $50 \Omega 6 \mathrm{y} .2 \mathrm{c} / \mathrm{o}$. 63 p ea. $470 \Omega 12 \mathrm{v} .4 \mathrm{c} / \mathrm{o} .73 \mathrm{p}$ ea. 2,780 $\Omega 48 \mathrm{v} .4 \mathrm{c} / \mathrm{o} .73 \mathrm{p}$ e2. $1,260 \Omega 48 \mathrm{v} .6 \mathrm{c} / 0.83 \mathrm{p}$ ea.
MAGNET DEVICES. $12 \mathrm{v} .3 \times \mathrm{H} . \mathrm{D} . \mathrm{c} / \mathrm{o}$ Contacts




NEW "F.I.R.E."' PLUG-IN
RELAY.-li5v. Coil $50 / 60$ c.p.s. 3 heavy duty silver change-over contacts. Very robust. 63 p .

NEW "'ISKRA' 240 Y. A.C RELAY:- ${ }^{3}$ contacts. 63 p .

SIEMENS HIGH SPEED RELAY, Type $89 \mathrm{~L} .1,700 \Omega+1,700 \Omega$ MINIATURE "HATCH MASTER" RELAY 6, 12 , or 24 v . D.C. operation. One make one break, contacts rated 5 amps.
at 30 v . Once current is applied, at $\begin{aligned} & \text { revay remains latched until input }\end{aligned}$ polarity is reversed. Manufacured for high acceleration re
 quirements by Sperry Gyroscope Co. Size: Length is dia. F" (including mount). Please state vertical or horizontal mount and voltage. $£ 1.63$ each.
ELECTROLYTIC CAPACITORS MULLARD.
 70p eac., 66.00 per doz. $1,600 \mu \mathrm{~F}$ 64v. $11^{\prime \prime}$ dia. $\times 3^{3}$
38 p -ea., E 3.50 per doz. $10,000 \mu \mathrm{~F}$ 10v. $1^{\prime \prime}$ dia. $\times 3^{\prime \prime}$
 50 p ea., $£ 4.50$ per doz.
HUNTS $1,000 \mu \mathrm{~F} 50 \mathrm{v}$. $1 \mathrm{z}^{\prime \prime}$ dia. $\times 2^{\prime \prime}, 25 \mathrm{p}$ ea., $10,000 \mu \mathrm{~F}$ 6 V . $18^{\prime \prime}$ dia. $\times 2^{\prime \prime}, 30 \mathrm{p}$ ea. 43.00 per dox. $16 \mu \mathrm{~F} 350 \mathrm{v}$
 dia. $\times 3,30 \mathrm{p}$ ea., 300 per doz. $12232 \mu \mathrm{~F} 275 \mathrm{v}$.
$\times 2^{\prime \prime}, 38 \mathrm{p}$ ea. $100 \mu \mathrm{~F} 100 \mathrm{v}$. $1^{\prime \prime}$ dia. $\times 2^{\prime \prime}, 25 \mathrm{p}$ ea.
ERIE. Ceramicon capacitor. Type CHV4IIP. 500 P.F. 30 KV Size $1.5^{n}$ dia. $\times 1.44^{\text {² }}$ long. 50 p ea. Carriage paid. HIGH CAPACITY ELECTROLYTICS. Cylinder. type with screw terminals on top. Average size 3 dia. $X$
$4 t^{"}$ high. "Mallory" $20,000 \mu \mathrm{~F} 30 \mathrm{v}$. D.C. $45 v$. D.C. surge. "Mallory $25,000 \mu \mathrm{~F} 25 \mathrm{v}$. D.C., 40v. D.C. surge. Mallory $35,000 \mu \mathrm{~F}$ 15v. D.C., 20 v . D.C. surge. "Mallary" $40,000 \mu \mathrm{~F} 10 \mathrm{v}$. D.C., 12 v . D.C. surge. "Spraguve" 40,000 \%F 10 v . D.C., 12v. D.C. surge. "General Electric" 46,500 $\mu$ F 25v. D.C., 30v. D.C. surge. "General Electric" $55,000 \mu$ F 15 v . D.C., 20 v . D.C. surge. 50 p each. Minimum
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MOTORS
AMPEX $7.5 v$. D.C. MOTOR. This is an ultra-precision tape motor designed for use in the AMPEX model AG20 portable recorder. Torque $450 \mathrm{GM} / \mathrm{CM}$. Stall load at 500 ma . Draws 60 ma on run. $600 \mathrm{rpm}+5 \%$ speed adjustment, incernal $\bar{A} F / R F$ suppression.
motor $3^{\prime \prime}$ dia." dia. $\times 1^{\prime \prime}$ spindle,,$~$ E16.50. Our price $\mathrm{E}^{4.25}$. P. \& P. 25 p Large quantity available (special quorations). Mu-metal enclosure available 75 p each.


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63 p ea. P. \& P. 12 p . Stock now 63p ea. P. \& P. 12p. Stock now
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THERMOSTAT. Adjustable THERMO STAT. Adjustable internal adjuster takes the maximum up to $120^{\circ} \mathrm{C}$. Circuit cuts in again at $3^{\circ}$ below cut-out setting. $42^{\prime \prime}$ capillary and sensor probe. The thermostat actuates a 15 amp . 250v. c/o switch. A second single pole on/off switch is incorporated in the adjustment mechanism. 88p.


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HEAVY DUTY PORTABLE BATTERIES. New ex WD. 12v. 75 AH. Built in stout metal cases with carrying handles and nifam socker outlet. Size $15 \frac{1}{2}^{\prime \prime} \times$
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ELECTRO CONTROL (CHICAGO). Shaded pole $240 \mathrm{v} .50 \mathrm{~Hz} .200 \mathrm{rpm} 10 \mathrm{lb} . / \mathrm{in}$. £2-50. P. \& P. 25p. MYCALEX. Open frame, shaded pole motors. 240 v . $50 \mathrm{~Hz}, 7 \mathrm{rpm} .28 \mathrm{lb} . / \mathrm{in} .80 \mathrm{rpm} .12 \mathrm{lb} . / \mathrm{in} .62 \cdot 25 \mathrm{each}$.
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SYLVANIA MAGNETIC SWITCH-a magnetically activated switch operating in a vacuum. Switch speed-4ms. temperature $-5\{$
to $+200^{\circ} \mathrm{C}$. Silver contacts normally closed to $+200^{\circ} \mathrm{C}$. Silver contacts normally closed
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Rotary Switches, 4 pole 3 way or 2 pole 6 way
Switch clearer, aerosol cans
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BM I Phonc pre-amp
BM 2 Tape preamp BM 3 Mike ore-amp BM21 F.M. Transmitter BM22 F.M. Wireless Guitar BM31 Electric Organ BM4I Code Oscillator BM42 Wireless Oscillator

Electrolytic Capacitors
$\mathbf{f 1} \mathbf{2 5} \quad 2,000 \mu \mathrm{f} \mathbf{2 5}$ volt Rev.
£I-25 1,000 $\mu \mathrm{f} 70$ volt
\&1. $\mathbf{2 5} \quad 10,000 \mu f 35$ volt
€1. $25 \quad 10,000 \mu \mathrm{f} 25$ volt
f1. $25 \quad 2,000 \mu \mathrm{f} 18$ volt
$2,000 \mu \mathrm{f} 18$ volt
$60 \mu \mathrm{f}+200 \mu \mathrm{f} 300$ volt
$60 \mu \mathrm{f}+200 \mu \mathrm{f}$
$400 \mu \mathrm{f} 275$ vole
$400 \mu f 275$ vol
$10 \mu f 6$ volt
$10 \mu f 6$ volt
$10 \mu f 25$ volt
$10 \mu f 25$ volt
$16 \mu f 250$ volt $32 \mu f 275$ volt

40p $\pm 1.00$ 50p 20p
$f 1.25$ 1.25
$98 p$ $98 p$
50 p 50p 5p $15 p$ 50p 25p
35p
50p
35p
20p
$30 p$
$25 p$
$2 p$
$4 p$
$8 p$
$8 p$ $8 p$
$8 p$

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| ASY22 | 10p | OC45 | 10p | 2N709 | 50p | 2N3703 | 13p |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| ASY29 | 25p | OC46 | 15p | 2 N 1302 | 15p | 2N3704 | 18p |
| ASZI7 |  | OCl41 | 22p | 2N1309 | 23p | 2N3707 | 15p |
| (OC35) | 25p | OCI39 | 22p | 2 N 1613 | 25p | 2N3877A | 40p |
| BC167 | 15p | OC74 | 20p | 2 N 1711 | 25p | 7401 | 40p |
| BCY70 | 18p | OC204 | 25p | 2N2646 | 58p | 7410 | 40p |
| BFX12 | 20p | 2G345 | 10p | 2N2926 | 15p | 7430 | 40p |
| OC41 | 20p | 2G371 | 10p | 2N3053 | 25p | 7472 | 55p |
| OC42 | 23p | 2G378 | 10p | 2N3055 | 75p | 7473 | ${ }^{90}$ |
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| OC44 | 15p |  |  |  |  |  |  |

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$2 \frac{1}{2}$ in $\times \operatorname{lin} \times 0.15$ in $6 \mathrm{p} \quad$ in $\times 33 \mathrm{in} \times 0.15$ in 28p $\quad 3 \frac{3}{3}$ in $\times 3 \frac{3}{2}$ in $\times 0.1$ in 24p
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18p. Special Offer Pazk consisting of $5 \frac{1}{2} \times$ lin boards and a Spot Face 18p. Special
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G90 Magnetic Stereo Cartridges, Diamond Needle, 6 mV output, $\mathrm{E4}$. ACOS GP 67/2 (Mono, Crystal) 75p. ACOS GP 91/3 (Compatible, Crystal) El. ACOS GP 93/I (Stereo, Crystal, Sapphire) \&1.25. ACOS GP 93iro (1-50. ACOS GP $94 / 1 \mathrm{D}$ (Stereo, Ceramic, Diamond) $£ 1-88$. ACOS GP 95/I (Stereo, Crystal with two L.P./Stereo needles) $\mathbf{f 1 - 2 5}$.

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$1,000 \mathrm{pf}, 1,200 \mathrm{pf}_{\text {, }} 1,500 \mathrm{pf}, 1,00 \mathrm{pf}, 2,20 \mathrm{pl}, 15 \mathrm{p}$ per dozen (all 400 V working). $0.15 \mu f, 0-22 \mu f, 0.27 \mu \mathrm{f}, 30$
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These small portable instruments measure
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| 95 |  |  |  |

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 DC volts 10 MV to 1,000 volts. Resistance 0.02
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 XL
K-11
6-P-Pole Plug

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Offered Brand New. Sealed Packs at a
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RECORDER \& Sensitive Controller Series 60 . Offered in good used con-
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200-yard reels equipment wire, size $1 / 024$, reels only 75 p. P. \& P. $12 \frac{1}{2}$ p.

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Cossor Electronic Invertors type CRA 200. A high quality device for producing a hisv 400 HZ single phase utput. Incorporating the following eatures: Input 23-28V D.C
Sull overioad prot
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* Buile to Aireraft specifications
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May be run in series operation for 3 phase requirements. Offered brand new boxed units. Price $\mathbf{£ 1 7 . 5 0}$
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Constant Voltage DC Power Supplies A stabilised unit supplying 48
A stabilised unit supplying 48 vdc at 4 amps
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 amp type fully shrouded with scale plate \& control knob. Good used condition. Price E 10 Carriage 75p. Aly
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No waiting, straight off the shelf and into your equipment, the Catalogue Nos 250 2202A, 4/33A63/1; coll resistance is ohms. Complete with base, and the price is $\mathbf{6 5}$. Limited quantity only Also: 2203A, 2200A, 2202A SEARCH RECEIVERS AN/APR/4
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 ally for mobile equipments, coil voltage 2 v . frequency up to 250 MHz at 50 watts. Small size only, 2 in. $X H$ in. Offeredbrand new, boxed. Price $\mathbb{f} \cdot 50$, inc. P.\&P.

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$\left.\begin{array}{l}\text { SK ohms } \\ \text { OK ohms }\end{array}\right\}$ ALL TEN TURN
20 K ohms
PRICE TEN TURN
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NR/R, 24 p: chirome bezel, round NR/R, ${ }^{24} \mathrm{p}$ : chrome bezel, round
amber NR/A, 24 p ; chrome bexel, round clear NR/3, 24 P . Neon,
square red square red
amber type
type $L S 5 C / C$, S $18 \mathrm{p} /$ Ail 18 p : clear
above are for 240. mains operation.
Filament typent: 6 V . 0.04 A
square red square red type LS5C/R-6v. 30p;
$6 y .0 .04 \mathrm{~A}$ amber type LS5C A-6v.,
 $\begin{array}{ll}\text { LS5C/R-12v., } & 3 \mathrm{P}^{2} \mathrm{p} \text { : } 28 \mathrm{~V} \text {. } 0.04 \mathrm{~A} \\ \text { LS5C/R-28v., } & \text { 45p. Other colours }\end{array}$ DIN CONNECTORS

| DIN CONNECTORS |  |  |  | WPO2 | 140 | 2A | 60.95 |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| Pole |  | Plug | Socket | BY164 | 42 | 1.4A | 80.49 |
| 2 ( | (Spkr) | 12p | 10p | 81912 | 80 | -1.5A | ¢0.66 |
| 3 |  | 13 p | 10p | C1412 | 80 | $\cdot 3 \cdot 2$ A | 11.02 |
| 4 |  | 14 p | 12p | E2512 | 80 | -15A | ¢1. 64 |
| 5 | $180^{\circ}$ | 15p | 12p |  |  |  |  |
| 5 | $240^{\circ}$ | 15p | 12p | *Reduc |  | by $30 \%$ | f nos |
| 6 |  | 15p | 13p | contac | led. |  |  |

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| 40361 | 55p | 2N2905 | 44p | 2N4291 | 15p | BCl 48 | Pp | BFX87 | 29p |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 40362 | 68p | 2N2905A | 47p | 2N4292 | 15p | BC149 | 10p | BFX88 | 26p |
| 2N696 | 17p | 2N2924 | 20p | ACl07 | 46p | BCl53 | 19p | BFY50 | 23p |
| 2N697 | 18 p | 2N2925 | 22p | AC126 | 20p | BC154 | 20p | BFY5I | 20p |
| 2N706 | 12p | 2N2926 | 11p | ACI27 | 20p | BC157 | 12p | BFY52 | 23p |
| 2N930 | 29p | 2N3053 | 27p | ACI28 | 20p | $8 \mathrm{BC158}$ | $11 p$ | BS $\times 20$ | 16p |
| 2NII31 | 29p | 2N3055 | $60 p$ | ACl53K | 22p | BC159 | $12 p$ | C407 | 17p |
| 2 N 1132 | 29p | 2N3702 | 13p | ACI76 | 16p | 8 BC 167 | $11 p$ | MCl 40 | 25p |
| 2N1302 | 19p | 2N3703 | 13p | ACY20 | 20p | BC168 | 10p | MPS6531 | 35p |
| 2 N 1303 | $19 p$ | 2N3704 | 13p | ACY22 | 16p | BC169 | $11 p$ | MPS6534 | $30 p$ |
| 2NI304 | 26p | 2N3705 | 13p | ADI40 | 63 p | $\mathrm{BCI}^{\text {Cl7 }}$ | $14 p$ | NKT211 | 25p |
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| 2 N 1308 | $36 p$ | 2N3709 | $11 p$ | ADI62 | 36p | BC183L | 10p | NKT403 | 65p |
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| 2N1613 | 23p | 2N3711 | 13p | AFI15 | 24p | $\mathrm{BC}^{\mathrm{BC} 212 \mathrm{~L}}$ | 16p |  |  |
| 2N1711 | 26p | 2N3819 | 23p | AFII7 | 22p | $\mathrm{BC2}^{\text {B }} 13 \mathrm{~L}$ | 16p | OCBI | 35p |
| 2N1893 | $54 p$ | 2N3904 | 35 p | AFl24 | 33p | $\mathrm{BC2}^{\text {BCY }}$ - | 16 p | OC83 | 20p |
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| 2N2218 | 34 p | 2N4058 | 13p | AF139 | 33p | BCY71 | 33p | ZTX300 | 14p |
| 2N2218A | 44p | 2N4059 | 10p | AF239 | 36 p | ${ }^{\text {BCY72 }}$ | 15p | ZTX301 | 16p |
| 2N2219 | 38p | 2N4060 | $11 p$ | ASY26 | 27p | BF115 | $23 p$ | 2TX302 | $22 \%$ |
| 2N2219A | 53p | 2N4061 | $11 p$ | ASY28 | 27p | BF167 | 18 p | ZTX303 | 22 p |
| 2N2270 | $62 p$ | 2N4062 | 12p | ${ }^{\text {BCl }} 107$ | $12 p$ | BFI73 | $19 p$ | 2TX304 | 27\% |
| 2N2369A | 19p | 2 N 4124 | 18 p | BC108 | $11 p$ | BF194 | 140 | ZTX ${ }^{\text {2 }}$ | 12p |
| 2N2483 | 35p | 2N4126 | $27 p$ | 8 BC 109 | 12 p | BF195 | $15 p$ |  | 25 |
| 2N2484 | 42p | 2N4284 | 15p | BCl 25 BCl 26 | 15\% | BF BF $\times 84$ | 31p | ZTX503 | 22p |
| 2N2646 | 47p | $2 N 4286$ $2 N 4289$ | 15p | BC1 BC | 220 | BFX85 | $34 p$ | ZTX504 | 52p |

## RESISTORS-10\%, 5\%, 2\%

| Code | Power | Tolerance | Range | Volues available | $\begin{aligned} & \text { to } 9 \\ & \text { (see } \end{aligned}$ | to 99 low). | 100 up |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| C | 1/20W | $5 \%$ | 82 $\Omega-220 \mathrm{~K} \Omega$ | E12 | 9 | 8 | 7 |
| C | 1/BW | 5\% | $4.7 \Omega-470 \mathrm{~K} \Omega$ | E24 | , | 0.8 | 0.7 |
| c | 1/4W | 10\% | $4 \cdot 7 \Omega-10 \mathrm{M} \Omega$ | E12 | , | 0.8 | 0.7 |
| c | 1/2W | $5 \%$ | $4.7 \Omega-10 \mathrm{M} \Omega$ | E24 | 1.2 | I | 0.9 |
| C | IW | 10\% | 4.78-10M | E12 | 2.5 | 2 | 1.8 |
| MO | 1/2W | 2\% | 100-1M | E24 | 4 | 3.5 | 3 |
| WW | iw | $10 \% \pm 1 / 20 \Omega$ | 0.22n-3.9 | E12 | 7 | 7 | 6 |
| WW | 3W | 5\% | $12 \Omega-10 \mathrm{~K} \Omega$ | E12 | 7 | 7 | 8 |
| WW | 7W | 5\% | $12 \Omega-10 \mathrm{~K} \Omega$ | El2 | 9 | 9 | 8 |

Codes: $C=$ carbon film, high stability, low noise.
MO = metal oxide, Electrosil TR5, ultra low noise.
WW= wire wound, Plessey.

## Values

E12 denotes series: $10,12,15,18,22,27,33,39$ 47, $56,68,82$ and their decades.
E24 denotes series: as E12 plus 11, 13, 16, 20, 24, CARBON TRACK POTENTIOMETERS, long spindles. Double wiper ensures minimum ong spind
Single gang linear 100 n to 2.2 Ma , 12p; Single gang log. 4.7 KO to 2.2 MO , 12 p ; Dual gang linear 4.7 kQ to $2.2 \mathrm{Ma}, 42 \mathrm{p}$; Dual gang $\log , 4.7 \mathrm{~K} \Omega$ to $2.2 \mathrm{MB}, 42 \mathrm{p}$; Log/antilog, $10 \mathrm{~K}, 47 \mathrm{~K}$, 1 MQ oniy 42 p ; Dual antilog, IOK only, 42p. Any type with 1 A D.P. mains switch, $12 p$ extra.

Only decades of 10,22 \& 47 available in ranges quoted.

CARBON SKELETON PRE-SETS
Small hish quality, type $P R$, linear only: $100 \Omega$, Small high quality $220 \Omega, 470 \Omega$, IK $2 \mathrm{~K} 2,4 \mathrm{~K} 7,10 \mathrm{~K}, 22 \mathrm{~K}, 47 \mathrm{~K}, 100 \mathrm{~K}$ $220 \mathrm{~K}, 470 \mathrm{~K}, \mathrm{IM}, 2 \mathrm{M} 2,5 \mathrm{M}, 10 \mathrm{M} \mathrm{\Omega}$. Vertical or horizontal mounting, 5 p each.

COLVERN 3 watt Wire-wound Potentiometers. $10 \Omega, 15 \Omega, 25 \Omega, 50 \Omega, 100 \Omega, 150 \Omega, 250 \Omega, 500 \Omega$, IK, SK, 2SK. SK, 10K, 15K, 2SK, SOK, 32p eacherem ZENER DIODES, $5 \%$ full range 6.4 V , yalues:
$400 \mathrm{~mW}: 2.7 \mathrm{~V}$ to 30 V , 15 p , ach; $1 \mathrm{~W}: 6.8 \mathrm{~V}$. 82 V . 27p each; $1.5 \mathrm{~W}: 4.7 \mathrm{~V}$ to 75 V , 60 p each. Clip to increase 1.5 W rating to 3 watts (type 266F), 4p.

[^10]Prices are in pence each for quantities of the same ohmic value and power rating. NOT mixed values. (lignore order.)

## CAPACITORS

MULLARD polyester C280 series $250 \mathrm{~V} 20 \%=0.01,0.022,0.033,0.0473 p$ each: $0.068,0.1,4$ p ach; $0.15,4 p ; 0.22,5 p .10 \%$. $0.33,7 p ; 0.47,8 p ; 0.68,11 p ; 1 \mu F, 14 p ; 1.5 \mu F$, 21 p ; 2.2 $\mu \mathrm{F}, \mathbf{2 4 p}$.
MULLARD SUB-MIN ELECTROLYTICS C426 range, axial lead $30 \cdot 1.2 \cdot \cdot 2.116 \mathrm{p}$ each 4/10; 4/40; 5/64:6.4/6.4;6.4/25; 8/4:8/40; $10 / 2 \cdot 5$ 10/16; 10/64; 12.5/25; 16/10; 16/40; 20/16; 20/64; $25 / 6$. $; 25 / 25 ; 32 / 4 ; 32 / 10 ; 32 / 40 ; 32 / 64 ; 40 / 16$ 40/2.5; 50/6.4; $50 / 25 ; 50 / 40 ; 64 / 4 ; 64 / 10 ; 80 / 2.5 ;$ $\begin{array}{llllll}80 / 16 ; & 80 / 25 ; & 100 / 6 \cdot 4 ; & 125 / 4 ; & 125 / 10 ; & 125 / 16 ;\end{array}$ 160/2.5; 200/6.4; 200/10; 250/4; 320/2.5; 320/6.4i

LARGE CAPACITORS
High ripple current types: 1000/25, 28p; 1000/50, $41 p ; 1000 / 100,82 p ; 2000 / 25,37 p ; 2000 / 50,57 p ;$ 2000/100, \&1-44; $2500 / 64$; 77p; 2500/70, 98p; 5000/25, $62 p ; 5000 / 50$, \&i-10; $5000 / 100$, L2.91: $10000 / 50,22 \cdot 40$.

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\& 15.00 . Carr. 90 p.

GARDNERS HEAVY $V$ DUTY HT TRANSFORMERS GARDNERS HEAV
pri. $110-220-204$ Sec. $255-0-6 a$. Conservatively rated. ' C' core
Table top connections. Size $10 \times 8 \times 7$ in. E12.75. Carr. $£ 150$.

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 Neptune 5 eries. Pri. $110-200-220-240 \mathrm{v}$. Sec. $250-0-250 \mathrm{v} .70 \mathrm{~m} / \mathrm{a} .$,

 Pri. T. $200-240 \mathrm{v}$. Sec. $500-0-500 \mathrm{v} .120 \mathrm{~m} / \mathrm{a} ., 6$. 3 v . $3.5 \mathrm{a} ., 6.3 \mathrm{v}$. 3 a ., 5 v . 3 a . $\mathbf{6 2 - 5 0 . ~ P . P . ~} 45 \mathrm{p}$.

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$240 \mathrm{v} .-110 \mathrm{v}$. or 100 v . Completely Shrouded fitted with
Two-pin American Sockets or terminal blocks. Please state which type required.

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| :---: | :---: | :---: | :---: | :---: | :---: |
| 1 | 80 | 2i | lb. | 62.00 | $3{ }^{30}$ |
| 2 | 150 |  | 16. | 6.2.75 | 35 p |
| 3 | 300 | 61 | lb . | 63.75 | 35 p |
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$220-240 \mathrm{v}$. 50 cycles, 3 watts 8 r.p.m
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\begin{abstract}
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| B12H | 1.75 | RCH84 | 0.45 | PC900 |
| :---: | :---: | :---: | :---: | :---: |
| CY31 | 0.35 | ECH200 | $0 \cdot 62$ | PCC84 |
| DAF96 | 0.88 | ECL80 | 0.45 | PCC89 |
| DF96 | 0.37 | BCL82 | $0 \cdot 32$ | PCC189 |
| DK96 | 0.41 | ECL83 | 0.65 | PCE800 |
| DL92 | 0.82 | ECL86 | 0.42 | PCF80 |
| DL94 | $0 \cdot 40$ | EF36 | 0.45 | PCFR2 |
| DL96 | 0.41 | EF37A | 0.45 | PCF84 |
| DM70 | 0.30 | EF39 | 0.40 | PCF86 |
| DY86 | 0.30 | EF40 | 0.50 | PCF200 |
| DY97 | 0.32 | EF41 | 0.62 | PCF201 |
| DY802 | 0.48 | EF80 | 0.25 | PCF801 |
| E88CC/01 | 1.80 | EF83 | 0.55 | PCF802 |
| E180CC | 0.42 | EF8S | $0 \cdot 32$ | PGF805 |
| E181CC | 0.90 | EF86 | 0.31 | PCF806 |
| E1s2CC | 1.05 | EF89 | 0.26 | PCF808 |
| Eabc80 | 0.82 | EF91 | 0.15 | PCH200 |
| EAF42 | 0.50 | EF92 | 0.37 | PCLI 81 |
| Eb91 | 0.15 | EF95 | 0.30 | PCL82 |
| EbC33 | 0.50 | EFF183 | 0.32 | PCL83 |
| EBC41 | 0.52 | EF184 | $0 \cdot 35$ | PCL84 |
| ECC81 | $0 \cdot 30$ | E: ${ }^{\text {L }}$ L200 | 0.75 | PCL85 |
| EBF80 | 0.42 | ELL34 | 0.52 | PCL8 ${ }^{\text {b }}$ |
| EbF83 | 0.42 | EL41 | 0.57 | PFL200 |
| EbF89 | 0.30 | EL42 | 0.53 | L36 |
| ECC81 | 0.30 | EL84 | 0.23 | L8 |
| ECC82 | 0.28 | ELL85 | 0.40 | PL82 |
| ECC83 | 0.30 | E.L86 | 0.40 | PL83 |
| ECC84 | 0.30 | ${ }^{\text {E.LP90 }}$ | 0.35 | PL84 |
| ECC85 | 0.40 0.60 | ELL95 | 0.35 0.85 | PL500 |
| ${ }_{\text {ECC88 }}$ | 0.50 0.37 | ${ }_{\text {k. }}^{\text {k. } 1.500}$ | 0.85 0.25 | PL504 |
| ECC189 | 0.52 | E M80 | 0.40 | PY33 |
| ECF80 | $0 \cdot 35$ | EM84 | 0.35 0.55 | PY80 |
| ECF82 | 0.35 | EY51 | 0.40 | PY81 |
| ECF83 | 0.75 | E Y86 | 0.35 | PY82 |
| ECF801 | $0 \cdot 62$ | EY81 | 0.35 | PY83 |
| ECF802 | 0.62 | E Y88 | 0.40 | PY88 |
| ECH35 | 0.60 | F240 | -0.25 | PY800 |
| ECH42 | 0.65 | ER81 | 0.27 | PY801 |
| ECH81 | 0.28 | GZ34 | 0.52 | QQvo |
| ECH83 | 0.42 | 上 T66 | $1 \cdot 60$ | 3-10 |

VALVE VOLTMETER TYPE TF 958 Measures AC 109 mV ; $20 \mathrm{c} / \mathrm{s}$ to $100 \mathrm{mc} / \mathrm{s}$,
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MARCONI YHF ALIGNMENT OSCILLOSCOPE TF 1104 . Combined sweep gmerator and CRO for $41-216 \mathrm{kHz}$. If range $10-40 \mathrm{MHz}$. VF range 5 kHz to 1 JMHz . Output 10 uV to 250 MV continucus at 50 ohms. Sweep
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 Bigex $\begin{array}{ll}\text { Q3 } & 0.35 \\ \text { Q5W } & 0.50 \\ \text { S6 } & 0.37 \\ \text { T6 } & 0.80 \\ \text { U6 } & 0.30 \\ \text { X4GT } & 0.40 \\ \text { X5GT } & 0.85 \\ 7 & 0.40 \\ \text { K7 } & 0.40 \\ \text { A6 } & 0.25 \\ \text { E6 } & 0.30 \\ \text { G6G } & 0.55 \\ \text { J6 } & 0.45 \\ \text { Q7A } & 0.35 \\ \text { R7 } & 0.80 \\ \text { W6 } & 0.80 \\ \text { W7 } & 0.70\end{array}$





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klystrons，etc．

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SPECIAL PURPOSE VALVES．PLEASE OFFER US YOUR SURPLUS STOCK．MUST BE UNUSED．

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Imperial College of Science \& Technology DEPARTMENT OF AERONAUTICS

There is a vacancy in this Department for an

## ELECTRONICS TECHNICIAN

 orSENIOR TECHNICIAN
to work primarily on a general purpose instrumentation project. Salary ranges are £1,136-£1,535 p.a. and $£ 1,493-£ 1,832$ p.a. el,136-Eively. Superannuation scheme, four
respectivel weeks holida, and canteen facilities.
Apply in writing giving details of experience and qualifications to the Assistant Director, Deportment 備Aeronautics, Imperial College, $^{\text {A }}$ DepartmentisAeronautics, Imperial College,
Prince Consort Road, London, S.W.7.

1264

## UNIVERSITY OF SHEFFIELD

## SENIDR TECHNICIAN AND TECHNICIAN

required for University Television Service itrom September 1971
One post requires appropriate qualifications and
experience in field of electronics, particularly T.V.: for experience in field of electronics, particularly T.V.: for the other, trairing in electrical and/or laboratory
techniques. Farmiliarity with woad and metal work an advantage.
Training given ir T.V. operations to enable appointees to become members of a team working on educational T.V. productions-

Applicants for Senior Technician should be at least
25 years old and have C. $\&$ G. Final Certificate 25 years old and have C. \& G. Final Cerrificate or
equivalent: Tech ician, minimum age 20 , with C. G . equivalent: Techmician, minimum abe subject.
Salary: Senior Technician $\mathbf{E 1 , 3 9 8 - 6 1 , 7 0 7}$ p.a.
Technician $\mathbf{E 1 , 0 4 1 - \mathrm { fl } , 4 1 0}$ p.a. each with basic qualification. Supplement for approved higher qualification. Superannuation Scheme.
Write immediataly to the Bursar (Ref. B.854), The University, Sheffeld $510^{\text {the }} 2 \mathrm{TN}$.

## EXPANDING COMPANY IN SAUDI ARABIA REQUIRES

 EXPERIENCED CERTIFICATED ENGINEERSFOR THE FOLLOWING POSTS CHIEF ENGINEER
B.Sc. or equivalent with 10 or more years experience in Operation and Maintenance of Transmission and Broadcasting Equipment.

## ENGINEERS TECHNICIANS

Experience in Operation and Maintenance of Broadcasting Equipment, Studio Equipment and Teleprinters.

Please submit a complete resume and state availability and salary required. Box WW 1270

## Service Technicians

Move to Harlow and enjoy the benefits of a good job with a successful company in the pleasant surroundings of our New Town.

Your job will be to service and repair products from our wide range of Airborne Instruments, Scopes and Test Gear. You will be working in our Harlow base workshop with the opportunity for occasional field trips. We will give you product training but we'll expect a good basic knowledge of Electronics preferably backed up with fault finding experience on transistorised and solid state devices.

Your starting salary would be from $£ 1,310$ per annum with excellent opportunities of promotion to Section Leader grades. In many cases we can assist with local New Town housing and help with your removal costs.

If you want to find out how to secure your position then 'phone or write now to:

R. T. Reid,

COSSOB

## LONDON BOROUGH OF HILLINGDON

## EDUCATION DEPARTMENT

Two suitably qualified and experienced technicians are required to undertake the maintenance and repair of visual and aural aids equipment in all parts of the Borough schools and other educational establishments. These are new posts.

## POST ONE

$£ 1,605-£ 1,866$ incl. LW. Additional responsibility for planning and operating the maintenance and repair programme.

## POST TWO

£1,179-£1,362 incl. LW.
8 cwt. van and tools provided for each post. Work base in Uxbridge. Current clean driving licence essential.

Application form and further particulars from the Establishment Officer, Ref. E/186/30, Manor House, Church Road, Hayes, Mx. Closing date August 2.

## Opportunities with Redifon in Radio Communications

Experienced Test Engineers are invited to write to Redifon with regard to vacancies in our Test Department at Wandsworth.
The salary range for these positions.is $£ 1,248-$ $£ 1.749$ plus. The Company is engaged in the design and manufacture of a wide range of radio communications and allied equipment from military pack-set to broadcast transmitter, including communications receivers, M.F. beacons, teleprinter terminals, complete radio office installations for the Merchant Marine and mobile H.F. S.S.B. stations. Our Test Engineers have sound technical knowledge coupled with good practical experience in the alignment and test of H.F. and V.H.F.

Communications equipment.
The work is varied and interesting and offers excellent opportunity to broaden experience in semiconductors S.S.B. and Frequency Synthesis.
Please write in the first instance to Norman Manion.
The Recruitment Officer, Redifon Limited Broomhill Road, Wandsworth, S.W. 18


## T

## Telecommunications Engineers

required for the installation, maintenance and supervision of modern electronic systems used in our offshore oilfield complex at Das Island in the Arabian Gulf. These are bachelor postings but carry generous home leave and allowances.
Candidates, aged 23 to 40 , should possess a minimum of HNC or equivalent, and have several years' practical experience with radio systems ranging from MF to Microwave multi-channel, with a good working knowledge of digital telemetry and automatic telephone systems.

- Please write, quoting reference R.943/ZH and giving relevant information about yourself to: G. I. Andrews, External Recruitment, The British Petroleum Company Limited, Britannic House, Moor Lane, London, EC2Y 9BU, or ring 01-920 6522 for an application form.

POOLE GENERAL HOSPITAL, POOLE, DORSET

Applications are invited from qualified candidates for the following post in the Electranics Department at Poole General Hospital:

## ELECTRONICS <br> TECHNICIAN III

Qualifications: ONC. HNC, City \& Guilds or equivalent.<br>Salary: $\quad £ 1,356 \times 8$ increments to £1.764 p.a.

The Department will be primarily concerned with the installation, testing and maintenance of an extensive range of diagnostic/therapeutic and allied electronic equipment, and ultimately with research and development of bio-medical equipment in consultation with medical staff.
The position offers adequate scope for initiative and career progression, including the possibility of assistance with further training.
Applications, giving full details, including qualifications, experience and the names and addresses of two referees, to the Hospital Secretary, Poole General Hospital, Poole, Dorset.

## EDINBURGH CITY POLICE REQUJRE A <br> WIRELESS TECHNICIAN

for Servicing and Maintenance of fixed and mobile broadcasting receiving system.
Salary scale $\{1,413$ rising by annual increments to \{1,611.
Applicants will be required to have a knowledge of UHF and VHF apparatus used on fixed and mobile stations, and be able to diagnose and repair faults.
They would be expected to have atrained the Ciry and Guilds Telecommunications Technicians A Current or an equivalent qualification.
Applications to
EDINBURGH CITY POLICE
7 CHAMBERS ST., EDINBURGH, EHI IHR [I269

## THE UNIVERSITY OF SUSSEX ELECTRONICS TECHNICIAN

An interesting post is available in a small growing department for a technician with experience of transistor applicants should be capoble of designing and constructing simple apparatus for a variety of experlments.

Salary scale: (a) $£ 1,011-£ 1,380$ or (b) $£ 1,041-£ 1,410$ Salary scale (a) is applicable to those not holding an approved basic qualification
Further particulars and forms of application can be obtained from the Secretary (Establishment). Office of Arts and Social Studies. Arts Building. University of Sussex, Falmer, Brighton, BN1 9 QN to whom applicatelephoning Mr. Crook, Brighton, 66755, ext. 339 1276

## ST. BARTHOLOMEW'S HOSPITAL LONDON, E.C. 1

Applications are invited for two TECH NICIAN posts in the DEPARTMENT OF MEDICAL ELECTRONICS. The work involves routine servicing of electronic apparatus and the construction of new equipment for special purposes.
Applicants must have an O.N.C. or the final City and Guilds certificate, or two 'A' level passes in science subjects and at least four years' relevant technical experience. Experience of hospital work is not essential. Salary will be on the Technician III and IV scales. $£ 1,446$ rising to $£ 1,854$ and $£ 1,296$ rising to $£ 1,590$ respectively. Applications in writing with the names of two referees should be sent to the Clerk to the Governors.

# SECTIONAL ENGINEER GRADE II 

## EAST AFRICAN COMMUNITY

$\star$ Up to $£ 2,718$

- 25\% gratuity
$\star$ Low taxation
$\star$ Contract 21-27 months
$\star$ Subsidised accommodation
$\star$ Education allowances
$\star$ Appointments Grant payable in certain circumstances

Required by the Meteorological Department for the installation, operation and maintenance of their radio telecommunications, radio sounding and radar equipment.
Candidates, up to age 45 , must possess O.N.C. or the City and Guilds Final Certificate (Telecommunications) plus 7 years relevant experience or have equivalent experience in one of the armed services. They should have a good theoretical and practical knowledge of FSK, ISB and SSB receivers and transmitters, Mufax and facsimile transmitters and recorders. A good working knowledge of radar systems is essential.

Apply to CROWN AGENTS, 'M' Division, 4 Millbank, London, S.W.I, for application form and further particulars, stating name, age, brief details of qualifications and experience and quoting reference number M2K/6904 I3/WF.

## up to $£ 1741$ p.a. and all the variety you want as a RadioTechnician

Variety is the keyword. As a Radio Technician with the National Air Traffic Services, you would be installing and maintaining a wide range of sophisticated electronic systems and highly specialised equipment. You would be involved with RT, radar, data transmission links, navigation aids, landing systems, closed circuit T.V. and computer installations. All custom-built to meet the stringent operational requirements of air traffic control throughout the U.K.

If you're aged 19 or over and have at
least one year's electronics experience, preferably with O.N.C. or C. \& G. (Telecoms.), you could qualify for entry to our training course. Your starting salary would be $£ 1,143$ (at 19) to $£ 1,503$ (at 25 and over), scale max. $£ 1,741$ - shift duty allowances. Good career prospects.

Write NOW for full details to:
A. J. Edwards, C.Eng., MIEE,

Room 705, The Adelphi, John Adam
Street, London WC2N 6BQ,
marking your envelope
'Recruitment-в/ww/27'.
Not applicable to residents outside the United Kingdom.


## ELECTRONIC ENGINEERS required

for new Technical Service Centre to be established at Hemel Hempstead by British Manufacturers and Servicing Group of a wide range of Business Equipment Products. Ideally suited for engineers experienced in Radio/T.V. H.M. Forces, Industrial electronics.
Please write to: Mr. D. D. Davies, Technical Services Manager, Control Systems Ltd.,
Technical Services Centre,
1 Frogmore Road, Apsley, Hemel Hempstead, Herts.

## Closed Circuit Television Engineer

This interesting and responsible position involves all aspects of the installation and service of a wide range of monochrome C.C.T.V. for use with medical $X$-ray apparatus. The equipment would include vidocon, orthicon, plumbicon and isacon tubes, light intensifying systems and 35 mm . video tape recording apparatus.
The position would ideally suit an engineer experienced in C.C.T.V. systems preferably with ONC/HNC, looking for a responsible position and a secure future in a progressive firm.
A good salary and several fringe benefits including a Company car will be offered to the successful applicant.

Please apply for an application form to:

> The Personnel Officer, G.E.C. Medical Equipment Ltd., East Lane, Wembley, Middx. Tel. 9041288

WESSEX REGIONAL HOSPITAL BOARD and WESSEX HOSPITAL MANAGEMENT COMMITTEES REGIONAL ELECTRONICS SERVICE
Suitably qualified Engineers and Technicians are required for the Board's new Regional Department of Electronics and Bio Medical Engineering and in similar departments in Hospitals located in Hampshire and Dorset.

## 1. ELECTRONICS ENGINEER

Qualifications: Chartered Member I.E.E., I.E.R.E.

## 2. ELECTRONICS TECHNICIAN I

Qualifications: H.N.C.-H.N.D. Full Technologica Certificate C. \& G.

## 3. ELECTRONICS TECHNICIAN III

Qualifications: O.N.C.-H.N.C.-C. \& G.

## 4. ELECTRONICS TECHNICIANS V

Qualifications: O.N.C. or A.2.
Salary Scales:

1. $£ 2,088$, rising by nine annual increments to £2,868 per annum.
2. $£ 1,877$ rising by five annual increments to £2,346 per annum.
3. $£ 1,800$ rising by eight annual increments to £2,500 per annum.
4. $£ 900$, rising by seven annual increments to £1,160 per annum.
Point of entry to the scale dependent on qualifications and/or experience.
Posts (1) and (2) will be based at the Board's Héadquarters in Winchester; Posts (3) and (4)
in various centres in the Region.
Departments will be concerned with all aspects of design-installation-testing and commissioning of a wide range of diagnostic/therapeutic and allied electronic equipment and data transmission systems

Research and Development in conjunction with Medical Staff will be undertaken in the short term future.
Application forms available from the Personnel Department, Highcroft, Romsey Road, Winchester, to which they should be returned by 2nd August, 1971

## UNIQUE OPPORTUNITY

Electronic engineer to join the management team of a small but fast expanding company supplying a wide range of advanced projection, sound and lighting control systems.
We want an experienced inventive engineer fully capable of designing and developing, relay and solid state sequence control equipment sound amplifiers, lighting control equipment, etc. Salary by negotiation.

Appiy: Technical Director, Audio Visual Equlpment Ltd. 73 Surblton Road. Kingston, Surrey 01-546-4565 1285

## TECHNICIAN REQUIRED

September for electronics workshop. Salary according to qualifications,

Senior technician H.N.C. £1,305-£1,712
Technician O.N.C. $£ 902-£ 1,415$
Junior technician 'O' level maths \& science
£525-£803

Day release possible for technicians and juniors. Written applications stating age, qualifications and experience, and names of two referees to Administrator, University Laboratory of Physiology Parks Road, Oxford.

# Sea-going Radio Officers can now make sure of a shore job and good pay. 



# SENIOR TELECOMMUNICATIONS TECHNICIAN 

$\star$ Up to $£ 2942$

* $\mathbf{2 5} \%$ gratuity
$\star$ Low taxation
$\star$ Appointments Grant payable in certain circumstances.

> Required by the Posts and Telecommunications Department to be responsible for the implementation of the planning, the installation and maintenance of all telecommunications facilities, the control of stores and the technical training of local staff.
> Candidates should possess the City and Guilds Full Technological Certificate (Telecomms.) or H.N.C. They should have at least IO years relevant experience in the provisioning, installation and maintenance of HF, MF, and VHF communications installations in the AM, CW and SSB modes; both valve type and transistorised solid state radio beacons; radio teleprinter using both tone on/off and two tone keying; multi channel VHF equipment and manual CB telephone exchanges.

[^11]
## RADIO OPERATORS

DO YOU HOLD

PMG II OR PMG I OR NEW GENERAL CERTIFICATE
OR HAD TWO YEARS' RADIO OPERATING EXPERIENCE? LOOKING FOR A SECURE JOB WITH GOOD PAY AND CONDITIONS?

Then apply for a post with the Composite Signals Organisation-these are Civil Service posts, with opportunities for service abroad, and of becoming established, i.e. non-contributory pension scheme.

Specialist training courses (free accommodation) starting January. April and September, 1972.
If you are British born and resident in the United Kingdom write NOW for full details and application form from

Recruitment Officer, Government Communications Headquarters, Oakley, Priors Road, CHELTENHAM, GIos. GL52 5AJ.
(Telephone: Cheltenham 21491, Ext. 2270)

## LABORATORY TECHNICIANS ELECTRONICS

(£1,056-£1,881 p.a. inc.)
The Central Electricity Research Laboratories, Kelvin Avenue, Leatherhead, Surrey, wish to recruit Laboratory Technicians for the construction and testing of a varied range of electronic and electro-mechanical apparatus and equipment, mostly prototypes, including chassis construction and layout, working from circuit diagrams, sketches and verbal instructions.

Applicants must be at least 25 , have served a craft apprenticeship or recognised period of training with several years' practical experience and possess ONC or equivalent. A radio and television engineer with suitable practical experience in this field would also be considered.

Write or phone for application form to the Personnel Officer at above address (L'head 4488, ext. 363) as soon as possible. Full details of the work and conditions of employment will be discussed with short-listed applicants during interview. Ref. WW/193.

1308

## UNIVERSITY OF ESSEX <br> DEPARTMENT OF ELECTRICAL ENGINEERING

TECHNICIAN
A Technician vacancy exists in the VISUAL SYSTEMS RESEARCH LABORATORY
Applicants should have an interest and preferably some experience in tele vision. The position offers interesting work on cameras and CRT displays, both colour and monochrome, for use in video telephone experiments being carrid out
research contact with the British Post Office.

Salary scale (with approved basic qualifications) Salary scale (with approved basic qualifications)
61,041-61,410 plus E 51 higher qualification allowance where appropriate.

Applications, giving age, technical qualifications and details of experience to the Registrar, University of Essex, Wivenhoe Park, Colchester, Essex.

## BRUNEL TECHNICAL COLLEGE, BRISTOL Department of

 MARINE AND AERO-ELECTRONICSApplications invited for following post. Duties to commence 1st September, or as soon as possible thereafter.

## LECTURER GRADE II in AERO-ELECTRONICS

Applicants must hold current Aircraft Radio Maintenance Engineers Licences, with Radar Ratings. Additional qualifications such as ' X ' Electrics, ' $X$ ' Instruments, etc., an advantage.
Further particulars and application form from: Registrar(S) Brunel Technical College, Ashley Down, BRISTOL BS7 9BU. Please quole reference number $71 / 33$. Closing date 30 th July. 1290

## EXPERIMENTAL OFFICER in mechanical engineering

Required to assist in development and research activities and provide technical support for maintaining laboratory equipment. Experience in designing experimental engineering equipment and in using electronic instrumentation are considered essential and some practical knowledge of pneumatic and/or hydraulic control systems would be desirable.

Candidates should hold a B.Sc. degree, H.N.D. or an H.N.C. with considerable industrial experience, and would be expected to organise the perience, and would be expected to organise the
work of a small technical force as necessary. It is work of a small technical force as necessary. It is
unlikely that candidates under 25 years of age would be considered.

Salary Scale $£ 1,902$ to $£ 2,592$ per annum.
Applications should be sent to the Staff Officer, University of Surrey, Guildford, Surrey.

## SITUATIONS VACANT

## O.E.M. require <br> ELECTRONIC ENGINEERS

to service a range of desk calculators and/or visible record computers. If you have experience in this field or in servicing digital equipment employing bipolar or M.O.S. semiconductors and are looking for a change, why not ring $01-4073191$ or write to:
E. J. LANDON, OFFICE AND ELECTRONIC

140/148 Borough High Street, London S.E.I, for an interview.

## BUSINESS OPPORTUNITY

Earn a substantial extra income through a fascinating part-time business of your own that you could share
with your wife and operate from your own home. with your wife and operate from your own home. exceeding 55000 per annum at the higher levels. We are looking for arganisational and managerial ability. VISTA MARKETING for an MPAintment MEAD 28754
A FULL-TIME technical experienced salesman re-
previous experience. salary, required to-The Manager,
Henry's Radio, Ltd., 303 Edgware Rd., London, W.2.
DRAUGHTSMEN. Mechanical and Electrical required
Dy expanding electronics company specialising in
$\begin{aligned} & \text { lighting control and audio visual products. This posi- } \\ & \text { tion is salaried and gives ample opportunity for advance- }\end{aligned}$
$\begin{aligned} & \text { tion is salaried and gives ample opportunity for advance- } \\ & \text { ment. Please apply Electrosonics Ltd., } 47 \text { Old Woolwich }\end{aligned}$
$\begin{aligned} & \text { moad. Greenwich. London, S.E.10. Tel. } 8584784 \text {. [22 }\end{aligned}$
INSTALLATION ENGINEER required for the servicing.
testing and installation of audio projection and light-
ing control equipment. An excellent opportunity for
electronics. Starting salary $£ 1,250$. The post offers
opportunities for travel in England and overseas. Apply
to The Personnel Director, Electrosonic Ltd., 47 Old
Woolwich Road. Greenwich, S.E. 10
TRANSMITTER Technician (34) seeks new position
I giving test and maintenance experience on (a) VHF
$\begin{aligned} & \text { Transmitters, or (b) professional quality Audio equip- } \\ & \text { men for Broadcasting, Sound Recording, or Public }\end{aligned}$
$\begin{aligned} & \text { men for Broadcasting, Sound Recording, or Public } \\ & \text { Addresses, etc.), anywhere in U.K. Box w.W. } 1311\end{aligned}$
Wireless World.
Wanted: Ambitious young man with good elec-
ham, Surrey. Ring Caterham 42515.


## 17" BBC/ITV <br> TELEVISIONS $£ 5$

plus p. \& D. \&1.00 C.W.O.
SUITABLE FOR ANY AREA
3 Channel $19^{\prime \prime} \mathrm{D} / \mathrm{S}$ TVs. ITV, BBC 1, BBC 2 , $£ 25$ inc. carriage. 17" 13 Channel, complete but untested, $£ 1.50$ each, plus $£ 1$ P. \& P., C.w.o.

SPEAKERS
$6^{\prime \prime} \times 4^{\prime \prime}, 7^{\prime \prime} \times 4^{\prime \prime} 30 \mathrm{HM}$
20p plus \&p P. \& P. each, C.W.O. regular deliveries throughout england

TRADE TV's
407 Thornton Road, Girlington, Bradiford 8, Yorks.

## TV's TV's TV's

SPECIAL OFFER-LIMITED PERIOD ONLY Thorn 800 ('hassis 13 (hannel stim TVs. (tood
working order. Polished cabinets. Only $\operatorname{E9.50}$ PLTS working or
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Complete with 13 ehammel tuners. ( (ood cabinets,

 $19^{\circ}$ BBC2 Sets C 14.50 .

PERFECT SPEAKERS EX TV Pm 3 ohm (minimum order two 5 in. round, 8 int. by $2 i^{2}$, r.
p. and $p$.

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| EB91 | 5p | 3014 | 12] D \| | PL,36 | 22 ${ }^{\text {d }}$ D |
| EBF89 | $12 \pm$ p | $1{ }^{2} \mathrm{C} 97$ | 1710 | PL 81 | $17 \frac{1}{}{ }^{\text {d }}$ |
| ECV82 | 12id | PCF86 | 1712 | 1'181 | 15p |
| EC180 | 710 | PC84 | $7{ }^{1} \mathrm{p}$ | PY800 | 15p |
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| EF85 | 12 LD | P(r88) | 1210 | 1 Y 33 | 2218 |
| EF183 | 12.0 | PCL85 | 2212 | [T191 | 17ำ P |
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| 630 LZ | $12 \frac{12 p}{20}$ | PCL86 | $17 \frac{1}{2} \mathrm{D}$ | 301P12 | 20p |
| 30 L 15 | 12ib ${ }^{\text {p }}$ | $1^{\prime}{ }^{1} 1883$ | 12tp | 30F5 | 10p |
| $2 \frac{2}{2} \mathrm{pp}$ | valv | and p . free | lers |  |  |

To suit lerguson 850, 900 chassis $\mathbf{E 2} \mathbf{5 0}$. p. \& 1). 50p.
SLOT METERS-SPECIAL OFFER Smiths Mk. 11 6d. Convertible to 5 y . (Smiths hit $\pm 5$ incl. post and vacking.
Please write with SAE for quotations on any spares. RADE DISPOSALS (Dept. T.S.), Thornbury

## now under new management

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Anhydrous, Technical Quality
Packed in steel drums containing $100 \times 1 \mathrm{lb}$. bags
Guaranteed High Quality-Low Price
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8 M.F.D., 2,500 volt capacitors $\mathbf{£ 3}$ 50 yard Interpolation Oscillator $\mathbf{£ 3}$
GREENWELD ELECTRONICS
24 Goodhart Way, West Wickham, Kent Prone: 01-777 2001

## COMPUTER PERIPHERALS

Closure of computer site makes available the following tape editing equipment:

CREED Model 75 Teleprinter with soundproof
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Model 25 punch at $£ 150$.
CREED Teleprinter, type 54 also keyboard per
forators (in Ferranti pegasus Code) at $\not 225$ Teletype punches BRPE
proof box at $K 155$ also BRPE 8 hole with sound-
5 hole converted
to phonic wheel sync, choice of two at $£ 40$ each.
ELLIOTT T.2/94. 250 c.p.s. 8 hole Optical Tape readers, choice of four at $£ 150$ each.
For further details contact
Business Engineering Services,
Gt. Bentley, Colchester, Essex. Telephone: 0473 (lpswich)/77197 or 0206 (Gr.
Bentley)/25/550.

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TELEPHONE: ARBORFIELD CROSS 610
Oscilloscape H,P. 175A 0-50MHz c/w 1781B $£ 350$
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Oscilloscope Textronix Plug-in Type CA..
Frea/Timer Counter HP $5233 \mathrm{~L} 0-2 \mathrm{MHz}$
Freq/Timer $\stackrel{6}{\text { Cougit }}$ Cor HP 3 3734 A © $0-5 \mathrm{MHz}$
Freq/Timer Counter Venner TSA $53 / \dot{H} \dot{P}$
6160 $£ 210$

PSU Solartron AS 1410 0-30V IA
650
PSU Advance PP6 Twin 0-30V 3A
$\begin{array}{r}\text { £ } 75 \\ \hline\end{array}$

Recorder HP 7035B X-Y Plotter. Almost $£ 235$
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$0-100 \mathrm{mV}$ and 0.200 mV FSD
Chart width $8^{\prime \prime}$. Chart speed
$1^{\prime \prime}, 2^{\prime \prime}$ and $4^{\prime \prime} / \mathrm{hr}$. Chart speed
Wave Analyser Airmec 853 HF Analyser $£ 40$
Signal Generator Airmec 210 HF
Generator $30 \mathrm{KHz}-30 \mathrm{MHz}$
Stability $005 \%$
Stability 005\%
Signal Generator Marconi TFI446 Millivoltmeter Advance VM 78
Digital Voltmeter Dynamco DM 2006 .. $£ 325$ looolb Thrust Vibration Systems Consisting of Shaker and Amplifier. Also available 2501b Thrust. Please ring for fuller details.
Weare always inneed of Good Test Equipment.
Contace us you are considering disposal. 1306

COLOUR, UHF and TV SERVICE SPARES. SPECIAL OFFER. leading Brit. maker's Colour Monito Panels designed to BBC standards. Pal filter and
 35p). Also quantity Colour TV Camera Panels.
Plessey colour scan coils $£ 5.75 \mathrm{P} / \mathrm{P}$
35 p , convergence coils $£ 3.80 \mathrm{P} / \mathrm{P} 25 \mathrm{p}$. Blue lateral $£ 1.25 \mathrm{P} / \mathrm{P} \quad 10 \mathrm{p}$ (or complete set $\begin{gathered} \\ 10 \\ 10 \\ P / P \\ 50 p \text { ). Latest type colour scan and }\end{gathered}$ convergence coils, with electrical control of static convergence $£ 6.25$ P/P 35p. Colour LOPT assembly incl. EHT output and focus control $£ 4.50 \mathrm{P} / \mathrm{P} 35 \mathrm{p}$, luminance/ chrominance panel $£ 1 \mathrm{P} / \mathrm{P}$ 25p. Integrated transistd.
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## INDEX TO ADVERTISERS

## Appointments Vacant Advertisements appear on pages 85-92

|  | Page |  | Page |  | Page |
| :---: | :---: | :---: | :---: | :---: | :---: |
| A1 Factors. |  | Hall Electric Ltd.......... | 8 | Quality Electronics Ltd. | 16 |
| Acoustical Mfg. Co., L | 29 | Harris Electronics (London) Lt | 24 | Quartz Crystal Co., L | 95 |
| Adcola Products Ltd. | Cover iii | Harriss P..... | 93 |  |  |
| Advance Electronics Ltd |  | Hart Electronics. | 64 |  |  |
| A.E.G. (G.B.) Letd. |  | Hatfield Instruments Ltd | 30 |  |  |
| Anders Electronics Ltd | 28, 34 | Heath, Gloucester) Ltd. | - 6,57 | Radio \& TV Components Ltd. . | 54 |
| Andor Electronics Ltd |  | Henry's Radio Ltd |  | Radio Components Specialists Lid.. | 97 |
| A.N.T.E.X. Ltd. | ${ }^{43}$ | Henson, R., Ltd. | 94 | Ralfe, P. F................ | 78 |
| Arrow Electric Switches Lid | ${ }^{23}$ |  |  | Rank Wharfedale Ltd. | 32 |
| Audio Supplies.......... | 36 |  |  | R.C.S. Electronics. | 20 |
| Audix, B. B. Ltd. | 28 | I.C.S. Ltd. | 60 | Reslosound Lid.. | 16, 25 |
|  |  | I.M.O. Precision Controls Ltd. |  | R.S.C. Hi-Fi Centres Ltd. | . 71 |
|  |  | Instructional Handbook Supplies | 96 | R.S.T. Valves Ltd.. . | . 54 |
| Barrie Electronics. | 53, 93 | Integrex Ltd. ${ }^{\text {a }}$ | 60 |  |  |
| Batey, W., \& Co. |  | Intercontinental Components... |  |  |  |
| Bentley, K. J., \& Partners. | 96 | I.T.T. Mobile Communications. | Cover ii |  |  |
| Bentley Acoustical Corporation Ltd. | - ${ }_{13}$ | Ivoryet Ltd................ |  | Samson (Electronics) Ltd. | 81 |
| B.I.E.T....... | - ${ }_{80}^{13}$ |  |  | Scientific \& Technical Services. | 70 |
| Bi-Pak Semiconductors |  |  |  | Service Trading Co....... | 75 |
| Bi-Pre-Pak Ltd. | - ${ }^{52}$ | J.E.F. Electronics. |  | Servo \& Electronic Sales L | 51 |
| Black, J.... ${ }^{\text {Bowtilerman }}$ L |  | Jackson Bros. (London) Litd | 32 | Shure Electronics Ltd. . | 40 |
| Bull, J. (Electrical) Ltd.... | . 72 | Jackson, J. D., Electronics. | 12 | Sinclair Radionics Ltd. | 47, 48, 49 |
|  |  | Jermyn Industries. |  | S.M.E. Ltd. ${ }^{\text {a }}$ (Radio |  |
|  |  |  |  | S.N.S. Communications Ltd. |  |
|  |  |  |  | Special Product Distributors Lit |  |
| Cambridge Audio Laboratories | . ${ }^{37}$ | Keytronics. | 64 | Starman Tapes. | 95 |
| Cesar Products Ltd. (Yukan). | . 94 |  |  | Steed, John, Research Ltd |  |
| Chiltmead Ltd......... | 73, 96 |  |  | Stephens Electronics | 74 |
| Colomor (Electronics Ltd.)..... |  | Labgear Ltd. |  | Sugden, J. E., Ltd. | 14 |
| Combined Precision Components |  | Labhire Ltd.. | . 98 | Surplus Electronic Trading. | 94 |
| Computer Sales and Service Ltd. | - ${ }^{57}$ | Lasky's Radio Lid. | . 53 | Sution Electronics Ltd..... | 94 |
| Concorde Instrument Co.. | . 96 | Lawson Tubes. | 93 |  |  |
| Croydon Precision Inst. Co. |  | Leda Tapes.. |  |  |  |
|  |  | Ledon Instruments Ltd | 30 |  |  |
|  |  | Levell Electronics Ltd.. |  |  |  |
| Dewtron |  | Light Soldering Developments | 26 | T.B. Technical Ltd. | 26 |
| Dexter \& Co. | .. 22 | Lowe Electronics....... | 95 | Telequipment Ltd. |  |
| Diathane Ltd. | . 94 | L.S.T. Components Ltd. |  | Teleradio, The (Edmo |  |
|  | - ${ }_{20}$ |  |  | Tinsley, H., \& Co., | 15 |
| Dixons Technical (CCTV) Ltd. . | - 20 |  |  | Trio Corporation Ltd.. |  |
|  | 96 | Marconi Instruments.............. | 21 | Tro Corporaion Ld... |  |
| D.T.V. Group Lid. |  | Marshall, A., \& Sons (London), Ltd. |  |  |  |
| Dumet Products. |  | McKnight Crystal Co. |  |  |  |
|  |  | Mills, W.......... | 68, 69 | Valradio Ltd. . |  |
|  |  | Milward, G. F... | . 77 | Vitavox Ltd. . | 70 |
| E.B. Instruments . |  | Modern Book ${ }^{\text {Multicore Solders }}$ | Cover ${ }^{22}$ | Vortexion Ltd. | 2 |
| Edwards. Scientific Int. Ltd.. | $\square^{-24}$ |  |  |  |  |
| Electronic Brokers. . . . . . . | 8, 59, 97 |  |  |  |  |
| Electronics Design Associates | . 70 |  |  |  |  |
| Electro-Tech Sales. | . 76 | Newmarket Transistors Ltd. | 11 | Watts, Cecil E., Ltd.. |  |
| Electrovalue. . |  | Nettlefold \& Moser Ltd. (GKN) | 24 | Wayne Kerr, The, Co., |  |
|  |  | Nicholls, E. R. | 93 |  | 10 |
| English Electric Valve C | $\begin{array}{r} 42, \\ .43 \\ \hline \end{array}$ | Nombrex Ltd | 28 | West Hyde Developments Ltd | 36 50 |
| Enthoven Solders Ltd. . |  |  |  | West London Direct Supplies. | 50 70 |
|  |  |  |  | Weyrad (Electronics) Ltd. | 54 |
|  |  | Osmabet Ltd. | 95 | Wilkinson, L. (Croydon), Lid.. | 70 |
| Farnell Instruments Ltd. . |  | Oxley Developments Co., Ltd.. | 20 |  |  |
| Ferrograph, The, Co., Ltd.. | .... 41 |  |  |  |  |
|  |  |  |  | Young Electronics. | 96 |
|  |  | Parker, A. B........ |  |  |  |
| Goldring Mfg. Co., Ltd...... | .. 33,35 | P.C. Radio Ltd. | 83 |  |  |
| Grampian Reproducers Ltd. | .. 36 | Plessey Electronics Ltd. |  |  |  |
| Greenwood, W., Electronic Ltd | 15 | Powertran Electronics. | $60$ | \& I. Aero Services Ltd. | 84 |

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