Digital wristwatch
Active notch filters
Cover Story

Some Marconi Instruments are designed to be mobile. Others are not – but do a lot of travelling all the same. In fact, nearly three-quarters of mi’s total sales stem from export orders.

So there are plenty of people in Milwaukee or Mannheim or Melbourne or Montevideo who are just as discerning about Marconi Instruments as you are. And they’re equally enthusiastic about mi service, too. We’ve service organisations in New Jersey, Munich, Paris and a whole lot of other places to see to that.

There are mi distributors and representatives in more than 60 countries throughout the world and we have 14 associated companies in Africa, the Middle, Near and Far East, North and South America and Europe.

mi, then, doesn’t only cover all the intricacies of planning and producing some of the world’s finest electronic testing and measuring instruments . . . .

It covers the world, as well.

mi: THE INTERNATIONALISTS

MARCONI INSTRUMENTS LIMITED

Longacres • St. Albans • Hertfordshire AL4 0JN • England • Telephone: St. Albans 59292 • Telex: 23350

A GEC-Marconi Electronics company
LOW COST TESTERS

LEVELL PORTABLE INSTRUMENTS

INSULATION TESTER

A logarithmic scale covering 6 decades is used to display either insulation resistance or leakage current at a fixed stabilised test voltage. The current available is limited to a maximum value of 3mA for safety and capacitors are automatically discharged when the instrument is switched off or to the CAL condition. The instrument operates from a 9V internal battery.

**RESISTANCE RANGES**
- 10M Ω to 10T Ω (10^11 Ω) at 250V, 500V, 750V and 1kV.
- 1M Ω to 1T Ω at 25V, 50V and 100V.
- 100k Ω to 10G Ω at 2.5V, 5V and 10V.

**CURRENT RANGE**
- 100pA to 100mA on a 6 decade logarithmic scale.

**MEASUREMENT TIME**
- < 3s for resistance on all ranges relative to CAL position.
- < 10s for resistance of 10G Ω across 1µF on 50V to 500V.

**RECORDER OUTPUT**
- 1V per decade ±2% with zero output at scale centre.
- Maximum output ±3V. Output resistance 1k Ω.

TRANSMITTER TESTER

Tests bipolar transistors, diodes and zener diodes. Measures leakage down to 0.5 nA at 2V to 150V. Current gains are checked from 1µA to 100mA. Breakdown voltages up to 100V are measured at 10µA, 100µA and 1mA. Collector to emitter saturation voltage is measured at 1mA, 10mA, 30mA and 100mA for I_C/I_B ratios of 10, 20, 30. The instrument is powered by a 9V battery.

**TRANSISTOR RANGES (PNP OR NPN)**

- I_CBO: 10nA, 100nA, 1µA, 10µA and 100µA f.s.d. acc. ±2% f.s.d. ±1% at voltages of 2V, 5V, 10V, 20V, 30V, 40V, 50V, 60V, 80V, 100V.
- BV_CBO: 10V or 100V f.s.d. acc. ±2% f.s.d. ±1% at currents of 10µA, 100µA and 1mA ±20%.
- I_B: 10nA, 100nA, 1µA, 10mA f.s.d. acc. ±2% f.s.d. ±1% at fixed I_C of 1µA, 10µA, 100µA, 1mA, 10mA, 30mA, 100mA acc. ±1%.
- h_FE: 3 inverse scales of 2000 to 100, 400 to 30 and 100 to 10 convert I_B into h_FE readings.
- V_BE: 1V f.s.d. acc. ±20mV measured at conditions on h_FE test.
- V_CE(sat): 1V f.s.d. acc. ±20mV at collector currents of 1mA, 10mA, 30mA and 100mA with I_C/I_B selected at 10, 20 or 30 acc. ±20%.

**DIODE & ZENER DIODE RANGES**

- I_D: As I_CBO for transistors.
- V_Z: Breakdown ranges as BV_CBO for transistors.
- V_DF: 1V f.s.d. acc. ±20mV at I_D of 1µA, 10µA, 100µA, 1mA, 10mA, 30mA and 100mA.

LEVELL ELECTRONICS LTD.
Moxon Street, High Barnet, Herts. EN5 5SD
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**WW—015 FOR FURTHER DETAILS**

Prices include batteries and U.K. delivery. V.A.T. extra.
Optional extras are leather cases and mains power units. Send for data covering our range of portable instruments.
A new range of sound equipment from Vortexion, System 2000 has been designed by our engineers to combine the aesthetics of design in the domestic equipment field with the near flexibility of a modular system. Like all our equipment Vortexion System 2000 is built to last.

No matter what your sound problem, whether hotel or local pop group, ask our Design Consultants how it can be solved with System 2000.
The Dymar 1785 portable AM-FM modulation meter.

No need to ask who's in control. It's you!

The Dymar Type 1785 is quickly and easily tuneable anywhere across the entire VHF band and into UHF to encompass the mobile 470MHz band.

Designed to measure the depth of modulation or frequency deviation of today's demanding mobile and portable transmitters, the 1785 offers four ranges of both peak or trough percentage modulation (3% fsd to 100%) and both positive and negative deviation (3kHz to 100kHz).

The sensitivity over the entire frequency range is better than 2.5mV into 50 ohms (-40dbm), which permits loose coupling to the transmitter under test. And internal noise is typically 44db below 3kHz.

Then, like most Dymar instruments, the 1785 is equally at home working from mains supply or in action in the field operating on its own rechargeable NiCd batteries.

With such value-for-money performance, you'll want to drive the 1785 to the limit - and that's why we emphasise that the 1785 is fully tuneable.

Want to know more? Use the Reader Reply Service or contact Dymar direct.

The Dymar range of instruments - designed for the mobile land, marine and air communications industry.

DYMAR ELECTRONICS LIMITED,
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Again and again and again

Given the time, the patience, and the money, one can connect* fifty 303 amplifiers nose to tail so that the programme goes through one after the other gradually deteriorating along the way.
Deteriorating? The fact is, that apart from a very slight background hiss – akin to a good tape recording – the programme will sound exactly the same at the end as when it started.

*Of course one must fit an attenuator to reduce the signal back to its original level between each amplifier.

Send postcard for illustrated leaflet to Dept.WW Acoustical Manufacturing Co. Ltd., Huntingdon PE18 7DB. Telephone (0480) 52561.

QUAD
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W/W—049 FOR FURTHER DETAILS

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The Miniaturised Bi-stable polarised relay type VPR and the P.O. approved relay type 23 are but two from a range used and approved throughout the electronics world. Each is built to uncompromising quality standards... with ultrasonic cleaning throughout coupled with exacting performance and timing checks.

The same goes for our new AC range.

These miniature plug-in relays have the same physical dimensions as the DC range. Shown: 2 and 4-contact versions.

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Oliver Pell Control Ltd.,
Tel: 01-854 1422  Telex: 897071

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Company__________________________
Address____________________________
Stable companions

Wide-range universal bridge **B602**
0.1-100MHz source/detector **SR268**
from Wayne Kerr

The B602 transformer ratio arm bridge measures impedance in all four quadrants of the complex plane over the frequency range 100kHz to 10MHz. Because of novel features incorporated in the design, values from virtually a short circuit to an open circuit can be measured. This bridge has established a standard of performance and flexibility which is unobtainable from any other radio frequency bridge.

A standard inductor is included in the bridge network in addition to standards of capacitance and resistance enabling periodic calibration of the scales which are correct at any frequency between 100kHz and 10MHz.

There are only two balance controls. One is direct reading in resistance and conductance, the other in capacitance and inductance and there is no interaction between them.

The stability realised allows a discrimination of 0.1% to be obtained for all types of measurement with a general accuracy of 1% over most of the impedance and frequency range.

The bridge is shown together with the SR268 Source and Detector which can also be used with other bridges in the Wayne Kerr range over the frequency band 100kHz to 100MHz. Nine frequency ranges are provided by this instrument and a single tuning control adjusts both source and detector to the exact frequency required.

Meticulous screening between the two sections provides freedom from bridge measurement errors due to leakage of the source signal into the detector. Common mode rejection transformers are incorporated in the input and output networks to reduce interference from unwanted signals, and push button attenuators are included to assist the logarithmic detector circuit to indicate approach of the bridge balance point.

For more information, either phone Bognor Regis (02433) 25811 or write to the address below:

**WAYNE KERR**
Durban Road, Bognor Regis, Sussex
Telex: 86120. Cables: Waynkerr, Bognor
A member of the Wilmot Breeden group

**SPECIFICATION**

**B602**
Frequency range: 100kHz to 10MHz
Accuracy: 1% up to 3MHz, 1pF to 10nF,
10Ω to 100kΩ, 1μH to 10mH
Overall impedance range: 1F to 1mF,
100μΩ to 100MΩ, (10nF to 10kΩ)
10pH to 10H
**SR268**
Frequency Range: 100kHz to 100MHz in 9 bands
(SR268L 46.5kHz to 46.5MHz)
Frequency accuracy: 2 - 3% according to band used.
Short Term Frequency Stability: 0.01%
Output level: 0 - 2.0V according to band used
Output attenuator: 3 - 6 - 10 - 20 dB additive steps, 75Ω
Input sensitivity for 10% meter deflection: 1 to 30μV according to frequency setting
Input attenuator: 4 steps of 20 dB, 75Ω
Detector bandwidth: 100kHz to 10MHz in 9 bands

WW—056 FOR FURTHER DETAILS
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Cray Avenue, St. Mary Cray,
Orpington, Kent BR5 3QJ.
Telephone: Orpington 27099
Telex: 896141

WW—005 FOR FURTHER DETAILS
The HY5 is a complete mono hybrid preamplifier, ideally suited for both mono and stereo applications. Internally the device consists of two high quality amplifiers: the first contains frequency equalisation and gain correction, while the second caters for tone and balance.

**Technical Specification**
- Inputs: Magnetic Pick-up 3mV RIAA, Ceramic Pick-up 30mV, Microphone 10mV, Tuner 100mV, Auxiliary 3.100mV, Input impedance 47kΩ. 1kHz.
- Outputs: Tape 100mV, Main output 0db (0.775 volts RMS).
- Active Tone Control: Treble -12db at 10kHz, Bass -12db at 100Hz.
- Distortion: 0.05% at 1kHz.
- Signal/Noise Ratio: 68db.
- Overload Capability: 40db on most sensitive input.
- Supply Voltage: 10-25 volts.
- Price: £4.75 + £1.19 V.A.T. P&P free.

The HY50 is a complete solid state hybrid Hi-Fi amplifier incorporating its own high conductivity heatsink hermetically sealed in black epoxy resin. Only five connections are provided: input, output, power lines and earth.

**Technical Specification**
- Output Power: 75 watts RMS into 8Ω.
- Lead Impedance: 4-16Ω.
- Input Sensitivity: 0.775 volts RMS.
- Input Impedance: 4Ω at 1kHz.
- Distortion: Less than 0.1% at 25 watts typically 0.05%.
- Signal/Noise Ratio: Better than 75db.
- Frequency Response: 10Hz - 50kHz / 3db.
- Supply Voltage: 25 volts.
- Size: 105 x 50 x 25 mm.
- Price: £6.20 + £1.56 V.A.T. P&P free.

The PSU50 incorporated a specially designed transformer and can be used for either mono or stereo systems.

**Technical Specifications**
- Output Voltage: 50 volts (25-0-25).
- Input Voltage: 210-240 volts.
- Size: L70, D90, H60 mm.

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r.f. test equipment

TM6 R.F. MILLIVOLT METER
1 mV to 300V f.s.d.
50kHz to 1.5GHz
Useful to 3GHz
Near true r.m.s. readings on low ranges

A range of general purpose r.f. signal generators, synthesizers and sweepers is offered covering 100kHz to 512MHz. Specialist instruments include a Marine Test Set covering 0.1 to 12MHz for use with receivers with narrow i.f. filters for S.S.B. reception and u.h.f./v.h.f. Test Sets suitable for work on alerters, pocket pagers and two-way personal radios. All the equipment is programmable and may therefore be used in A.T.E. systems or operated manually.

Complementary r.f. test equipment includes an r.f. millivoltmeter, a programmable attenuator and an x-y recorder.
A clear view of the band

Eddystone EP961 MkII Panoramic Display Units provide visual monitoring of all signals in a selected band. MkII-A is tunable from 50kHz to 800kHz, matching the IFs used in MF and HF communication receivers.

MkII-B covers 500kHz to 36.5MHz. It is ideal for use with VHF and UHF receivers for monitoring FM broadcasts and communication transmissions, and its usefulness extends into the laboratory field. Both versions can be used with direct aerial input in many applications.

Eddystone Radio Limited
Member of Marconi Communication Systems Limited
Alvechurch Road, Birmingham B31 3PP, England
Telephone: 021-475 2231 Telex: 337081

For full information on the new WO.33B, contact RCA Electronic Components
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BLACK ARROW ELECTRONICS Ltd Millbrook Road, Yate, Bristol BS17 5NX Tel: Chipping Sodbury 315824

RCA’s new 3-inch scope... an entire servicing system

1. It’s an 8MHz general purpose scope. Typical composite TV video signal.

2. It’s a “Quicktracer” Transistor/Diode and Component Tester. Typical junction wave form.

3. It’s a Vectorscope for colour TV AFP C alignment. Colour bar generator used for test signal.

4. It’s a “ringing” tester for coils, yokes, transformers. Typical ringing test pattern.

5. It includes the WG-400A Direct Law-Capacitance Switch Probe and Cable with BNC type connector, and a special “Quicktracer” probe.

WW—035 FOR FURTHER DETAILS

WWW.americanradiohistory.com
Minimod, the versatile British range of encapsulated power supplies first introduced in 1973, has now been extended to cover European and North American mains voltages (and is interchangeable with most American types). Normally available ex-stock, all units are fully stabilised with fold back current limiting – the 5V models have over voltage crowbar too!

### STANDARD MODELS

<table>
<thead>
<tr>
<th>Type</th>
<th>Output Voltage</th>
<th>Output Amps</th>
<th>Short Circuit Current mA (Typical)</th>
<th>% Regulation Line and Load</th>
</tr>
</thead>
<tbody>
<tr>
<td>PU01</td>
<td>5 ± 0.1</td>
<td>0.5</td>
<td>370</td>
<td>0.3</td>
</tr>
<tr>
<td>PU02</td>
<td>5 ± 0.1</td>
<td>1.0</td>
<td>770</td>
<td>0.5</td>
</tr>
<tr>
<td>PU03</td>
<td>5 ± 0.2</td>
<td>0.10</td>
<td>37</td>
<td>0.1</td>
</tr>
<tr>
<td>PU04</td>
<td>5 ± 0.2</td>
<td>0.20</td>
<td>84</td>
<td>0.1</td>
</tr>
<tr>
<td>PU09</td>
<td>5 ± 0.2</td>
<td>0.12</td>
<td>45</td>
<td>0.1</td>
</tr>
<tr>
<td>PU06</td>
<td>12 ± 0.2</td>
<td>0.24</td>
<td>120</td>
<td>0.2</td>
</tr>
<tr>
<td>PU11</td>
<td>18 ± 0.2</td>
<td>0.15</td>
<td>50</td>
<td>0.1</td>
</tr>
<tr>
<td>PU10</td>
<td>15 ± 0.2</td>
<td>0.10</td>
<td>37</td>
<td>0.1</td>
</tr>
<tr>
<td>PU12</td>
<td>12 ± 0.2</td>
<td>0.10</td>
<td>45</td>
<td>0.1</td>
</tr>
<tr>
<td>PU13</td>
<td>18 ± 0.2</td>
<td>0.05</td>
<td>23</td>
<td>0.1</td>
</tr>
</tbody>
</table>

Input voltage ranges 103 - 126V, 200 - 240V. 210 - 250V. Frequency 50 - 400 Hz all types.

Comprehensive specification given in brochure GT 29b which is available on request.

**SPECIAL DESIGN SERVICE**

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"He's asking for a reed relay assembly with a 30kV isolated coil"

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138 GRAYS INN ROAD, W.C.1 Phone: 01/837/7937

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Valradio sinewave and square wave transvertors now incorporate SILICON transistors resulting in greater reliability and more stable performance at high ambient temperatures, including tropical climates.

TYPE D12/400S
A wide selection of types are available to drive practically any equipment within the power rating.
A random selection of types:

<table>
<thead>
<tr>
<th>Input</th>
<th>Output</th>
<th>Price</th>
</tr>
</thead>
<tbody>
<tr>
<td>C12/30S</td>
<td>12v DC 115/230v</td>
<td>£40.00</td>
</tr>
<tr>
<td>C24/60S</td>
<td>115/230v 60W</td>
<td>£58.20</td>
</tr>
<tr>
<td>D12/400S</td>
<td>12v DC 115/230v</td>
<td>£206.85</td>
</tr>
<tr>
<td>D12/500T</td>
<td>12v DC 115/230v</td>
<td>£110.55</td>
</tr>
<tr>
<td>D24/150T</td>
<td>24v DC 115/230v</td>
<td>£39.60</td>
</tr>
<tr>
<td>D12/250/24</td>
<td>12v DC 24v DC 8A</td>
<td>£83.10</td>
</tr>
</tbody>
</table>

Please send for literature WW675

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K12 Meter with drive components ............................. £11.00
K1-12 package price ........................................... £85.00
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SL3018 dual trans. ........................................... £1.40
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Postage per item ................................................. £0.10

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We thought it was about time somebody supplied direct reading meters for high voltage, so we've produced three—one for up to 5kV, one for 15kV and one for 30kV to complement our range of HV power supplies. The meters are operated by two 9V internal batteries (800 hours life) linked with a built-in checking facility. Positive or negative ground is available, selected by a front panel switch. And, as with all Brandenburg products, there is a 12 months unconditional guarantee.

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<th>Height</th>
<th>Depth</th>
<th>Case no vents</th>
<th>Case with vents</th>
<th>Chrome leg</th>
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Types 21, 22, 23 and 24 are finished in olive green hammertone with front panels in light straw gloss enamel. Fitted with ventilated rear panels only. No louvres in the base.
Telequipment's new dual trace 10 MHz battery operated oscilloscope

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*Exclusive of VAT
This month's front cover is a composite view of David Clegg's digital watch, a full description of which starts this month.

IN OUR NEXT ISSUE

Electronics and the railways, a survey of past, present and future developments in communications from signalling to fail-safe systems in high-speed travel.

Audio level meter. Constructional design, using columns of I.e.ds for logarithmic display, has characteristics similar to BBC peak programme meter.

Solid-state wristwatch. The second part of this unusual project gives the printed circuit board design and other constructional details.
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The analogue to digital conversion

Until about twenty years ago, the majority of measuring instruments displayed their results in an analogue manner, usually by means of a pointer meter. The accuracy of measuring electronics was such that the precision of this kind of indication was adequate. At this point, frequency measurement using flip-flops and logical gates came into large-scale use, and a numerical indication was the obvious way to display the reading. The numbers did not imply a spurious, mythical accuracy of measurement, because these instruments were capable of a high degree of accuracy.

Since then, measurement capabilities in many fields have improved and digital indication has been adopted to take advantage of this improvement. We have now reached the stage where an instrument is sometimes regarded as not quite "with it" unless a row of numbers appears on the front panel. Digital readout appears, in a sense, to have become a fashion and is sometimes used when the simpler pointer indication may have advantages. It ought to be remembered that the presentation of a measurement in numerical form does not mean that it is necessarily a more accurate measurement. It may give a more precise reading, but that is not to be confused with precision of measurement.

In some instances, a digital display can be a liability. Any slowly-varying quantity or a parameter which is only required to be known roughly is not a suitable case for the digital treatment. In the first instance the display may be a jumble of changing figures and in the second the display must be read rather than recognized. Clock faces, for instance, are very rarely read and are sometimes made without numerals: the position of the hands is taken in at a glance. Or again, a speedometer is usually an analogue indicator and one feels that any time spent reading numbers on a car dashboard would be better employed in looking where one is going, particularly while accelerating or decelerating, when the numerals on a digital indicator can be incomprehensible.

This is not, of course, a tirade against the superb digital instruments designed for precision voltage or frequency measurement — there is often no other way to present such readings. But a feeling persists that the humble moving-coil meter on the end of amplifying or impedance-changing electronics (or even displaying the results of measurements made by digital means, such as frequency) can still give a good account of itself.

Doubtless, many engineers who would consider themselves improperly dressed without an array of digital instruments in their equipment will disagree, pointing out that not only have we published a design for a digital speedo, but are now in the digital watch and clock business. We welcome their comments and, if they are sufficiently numerous and interesting to other readers, will publish a selection of them in a future issue.
This article describes the design and construction of a wristwatch having a liquid crystal digital display and a quartz crystal time reference. The author was originally tempted to consider the design of a watch with a conventional display, i.e., one with hands driven by a sub-miniature stepping motor. This was later ruled out, however, because it involved a lot of conventional watchmaking expertise quite beyond the limited resources of the author.

Design considerations
As a starting point, the following requirements were set down in order of importance:
(1) The watch should be accurate; at least as accurate as a more conventional watch of equivalent price. The timing reference must, therefore, be a quartz crystal which will result in a stability of better than 1 second per day.
(2) The finished watch should be a reasonably presentable piece of jewellery, i.e. it must not only look good, but also be a sensible size. The author had no intention of wearing a three-inch die-cast box on his wrist!
(3) The display must be easy to read under all reasonable conditions of use.
(4) The batteries must be easily obtainable and must last a sensible length of time. The absolute minimum acceptable battery life would be three months or, preferably, longer.

While compiling this list, the author was tempted to add a fifth requirement — that the watch should be easy to assemble — but later realised that this and requirement (2) are incompatible. It will become obvious during the course of this article that this design is very much a compromise because, in early 1974, there were very few suitable components available: all the major ones being imported from either the USA or Europe. For this reason the watch is quite difficult to construct and is, unfortunately, not suitable for economic manufacture in quantity. The watch described here can be built for around £50.

Logic type. Considerations of size and power consumption ruled out the use of s.s.i. or m.s.i. bipolar logic for the watch: the only alternative was i.s.i./c.m.o.s. and at that time only one manufacturer's product was available in this country. This was the two-chip system designed for wristwatches with digital displays by Solid State Scientific Inc. One of the chips is a silicon gate c.m.o.s. oscillator and divider packaged in a ten-lead flat-pack — the SCL-5425-AF — and the other is an aluminium-gate watch circuit with outputs suitable for driving liquid-crystal displays, packaged in a 30-lead flat-pack — the SCL5424-F.

Display. After deciding on a digital display, the author's first intention had been to use an i.e.d. display and to provide a "Display" button to reduce power requirements, so that the display would draw power from the battery only when required. The SCL-5424-F watch chip will not drive an i.e.d. display directly and so it was decided to interface it with CD-4009 c.m.o.s. inverters. At that time, however, c.m.o.s. was in short supply, and c.m.o.s. in flat-packs (necessary because of the size) was almost completely unobtainable, delivery times in excess of six months being quoted in some cases.

The alternative was to use a liquid-crystal display which would require no interfacing, since the SCL-5424-F watch chip was designed to drive this type of display directly. Again, at this time there was considerable difficulty in obtaining a liquid-crystal display suitable for a watch. There was quite a lot of literature from various manufacturers, but the only device which was actually obtainable was the LC-201135 from Brown, Boveri and Company. One important advantage gained from using this particular display was its ability to operate from a very low voltage (1.5V to 5V r.m.s.) and still provide a reasonable contrast ratio.

Crystal. The quartz crystal required by the SCL-5425-1F is a 32.768 kHz type. Motorola manufacture a suitable crystal in a sub-miniature vacuum envelope — the MTQ-32. They also make a sub-miniature ceramic trimmer capacitor for wristwatches; this is the MTT-02.

Batteries. It took the author approximately four months to obtain the main components for the watch, these being logic i.c.s., display, quartz crystal and trimmer capacitor. Batteries were now the only components outstanding on the shopping list and the author confidently expected this to be very easy since he had obtained the "Designers Guide to Battery Systems" from Mallory. This book describes a very comprehensive range of batteries including no less than six specifically designed for electronic watches.

At this time the author discovered that early wristwatch cases were available from wholesale watchmakers, so rather than try to make one (the die-cast box effect) it was decided to buy a suitable case ready-made. Before this could be done, however, the type and size of battery or batteries would have to be decided upon.

After visiting a number of large jewellers in central London, it was discovered that watch batteries are not available for sale over the counter; electronic watches requiring new batteries are, apparently, returned to the manufacturers to have their batteries replaced.

To have gone ahead and designed a watch using a battery from the Mallory range, but which proved difficult to obtain in practice, would have broken design rule (4). The author spent some time, therefore, visiting chemists,
jewellers and photographic shops in central London to determine which small mercury batteries were easily obtainable. The result of this investigation showed that the best battery for the watch would be the RM-312H mercury battery designed primarily for hearing aids. This is obtainable from most chemists and with a capacity of 36mAh it would appear to fulfil design requirement (4). Three of these batteries are used in the watch and they will have a life in excess of nine months.

**Oscillator design**

The timing reference in this watch is a quartz crystal oscillating at 32.768kHz. This frequency is popular with i.c. manufacturers because it is an exact power of two, \(Q^{15} = 32768\) thus simplifying frequency divider design. It is also a compromise between power consumption in the oscillator and the physical size of the crystal; lowering the frequency increases the size of the crystal, making it too large for a wristwatch, while increasing the frequency causes an unacceptable increase in power consumption. Future trends in watch design, however, are tending towards much higher frequencies (as high as 4MHz) to increase stability, with the use of silicon-on-sapphire c.m.o.s. integrated circuits to maintain extremely low power consumption.

The equivalent circuit for the Motorola MTQ-32 quartz crystal is shown in Fig. 1 which gives some typical component values. A typical \(Q\) of 50,000 would be obtained. The MTQ-32 is an NT-cut crystal and is mounted in an evacuated sub-miniature envelope.

There are two resonance conditions that a quartz crystal can exhibit – parallel resonance and series resonance. If it is assumed, for simplicity, that \(R_i = 0\), then the crystal impedance is given by:

\[
Z = \omega^2 L C_x \cdot \frac{1}{\omega^2 L C_x + \omega (C_a + C_p)}.
\]

The series resonant frequency \(f_s\) is defined at the zero impedance point, where \(\omega^2 L C_x = 1\), as:

\[
f_s = \frac{1}{2\pi \sqrt{LC}}.
\]

The parallel resonant frequency \(f_p\) is defined as occurring at the infinite-impedance point, where \(\omega^2 L C_x = \omega (C_a + C_p)\), and is

\[
f_p = \frac{1}{2\pi \sqrt{LC}} \cdot \left( \frac{1}{C_a + C_p} + \frac{1}{C_a} \right).
\]

Figs. 2(a) and 2(b) respectively show the variations of \(X_E\) and \(R_s\) with frequency for typical crystal. Unfortunately \(R_s \neq 0\) for practical crystals and the real condition frequencies are called resonance \(f_r\) and antiresonance \(f_a\). These two frequencies are defined as occurring when the crystal appears purely resistive, i.e. when \(X_E = 0\). In practice \(f_r\) is very close to \(f_s\) and \(f_a\) is very close to \(f_p\). Fig. 3 shows the practical oscillator circuit used in the watch – in this circuit the resonant frequency \(f_s\) lies very close to, but below \(f_p\). The MTO-32 crystal is designed to resonate at the specified frequency of 32.768kHz only when it has an external capacitance \(C_i\) of 10-12pF in value in shunt with \(C_p\).

\[
C_i = \left( \frac{C_a + C_p}{C_a} \right) C_x,
\]

where \(C_{in}\) is the input capacitance of the 5425 inverter, and is probably in the region of 5pF.

The inverter is biased into its linear mode by the series combination of \((R_1 + R_2)\) which should have a value of about ten times the crystal impedance at resonance \(f_0\) (typically \(Z > 10^9\) ohms) but should be lower than the various d.c. leakage impedances – the inverter input impedance, for example. A value of between 10 and 100\(\Omega\) for \(R_1\) satisfies these requirements. Unfortunately, however, physically small high value resistors (>10k\(\Omega\)) are very scarce; indeed, only a very few types of small resistor can be obtained in values above about 330k\(\Omega\). The Mullard CR25 range of carbon film resistors extends to 10M\(\Omega\) and this is the value of \(R_1\).

Resistor \(R_2\) controls the drive level to the crystal and contributes to the feedback network attenuation constant. Its value should be about ten times the crystal series resistance \(R_s\) to give reduced dependence of frequency on the supply voltage.

---

Fig. 1. Equivalent circuit of the MTQ-32 crystal. The values shown are typical and result in a typical \(Q\) of 50,000.

Fig. 2. The crystal reactance \(X_E\) plotted against frequency is shown at (a), while (b) is the resistance \(R_s\) against frequency.

Fig. 3. The practical crystal oscillator circuit, using a logical inverter as the maintaining amplifier. The trimmer is type MTT-02.
The trimmer capacitor VC₁ is adjustable between 6 and 35pF, giving a frequency adjustment of about ±1 Hz from the nominal 32.768kHz. This represents a timing adjustment of about ±2½ seconds per day. VC₁ is a subminiature ceramic trimmer, the MTT-02, manufactured by Motorola to complement the MTQ-32 crystal.

**Dividers**

The main watch integrated circuit (SCL-5424-F) requires an input of 64Hz and this is provided by the oscillator and divider integrated circuit, the SCL-5425-AF.

Fig. 4 is a logical block diagram of the SCL-5425-AF. It consists of an inverter between pins 1 and 2, to which are connected the frequency determining components to form the oscillator, followed by nine binary dividers. There are two outputs, one at the input frequency divided by 2² (output 1) and the other at the input frequency divided by 2³ (output 2). With an input of 32.768kHz, output 1 will be 32768 ÷ 512 = 64Hz and output 2 will be 32768 ÷ 128 = 256Hz.

The binary dividers used in the 5425 are master/slave toggle flip-flops whose output transitions occur on the rising edge (logic 0 to logic 1 transition) of the input clock. The logical construction of these toggle flip-flops can be seen in Fig. 4. The output from these divers has, of course, a 50% duty cycle (1:1 mark:space ratio). Resistive loads connected to the outputs would, under these conditions, dissipate an unacceptable amount of power, and the output pulse duty cycle is therefore reduced from 50% to 0.1% for output 1 and to 0.4% for output 2. These low duty cycles are generated by the output pulse flip-flops which are set by the first logic 0 to logic 1 transition of TF when TS is high and reset by the next logic 1 to logic 0 transition of TF (Fig. 5). Further changes in TF have no effect until TS has gone low and then high again, when another output pulse is generated. The net result of this is that the output pulse flip-flop gates one 15μs pulse of the input frequency at the repetition rate of the output frequency.

Output 1 is an uncommitted p-channel transistor and requires an external drain load of 100kΩ. Output 2 is a conventional c.m.o.s. inverter with an output “on” resistance of 1kΩ at a current of 300μA. This output of 256Hz is primarily intended for driving voltage up-converters to produce the 10 to 15 volts required by many liquid-crystal displays, and it is not used in this design.

**Main logic and display**

A block diagram of the SCL-5424-F watch is shown in Fig. 6. It accepts an input of 64Hz from the oscillator and divider I.C. and directly drives the liquid crystal display. The 64Hz input (pin 23) has an n-channel transistor connected down to V₆ with its gate connected to V_DD. It is intended to provide a load for
the p-channel current source transistor in the SCL-5425-AF. It is only effective, however, with a supply of 15 volts; with $V_{DD} = -4$ volts an external resistor of 100 kΩ is required, since at this voltage the transistor has a very high channel resistance. This value of load resistor might appear to be rather low as it results in a peak current of 40 μA ($I = 4 \times 10^{-3}$ A) but, since there is only one 15 μs pulse every 15 ms (duty cycle 0.1%), the mean current is only 40 nA. If this resistor is made too large, then the 64 Hz input pulse fall time will be too long and may cause incorrect operation of the divider stages following.

There are two "Reset" inputs (pins 29 and 30) which for normal operation are connected to $V_{DD}$ (0 V); through two 1 MΩ resistors (R1 and R2). When reset 2 (pin 30) is taken to $-V_{dd}$ (case potential) the hours increment at a 1 Hz rate. Connecting both resets 1 and 2 to $-V_{dd}$ causes the minutes to increment as for the hours. Connecting reset 1 to $-V_{dd}$ by itself, stops the watch and resets the divide-by-60 stage preceding the minutes counters. This is to enable the watch to be synchronized to a reference time source, for example, the Greenwich Time Signal.

Liquid crystal displays require a symmetrical a.c. waveform for correct operation. There must be no d.c. component present in this waveform or electrolytic effects within the cell will seriously damage the display, thus shortening its life. The SCL-5424-F watch chip, therefore, contains interface circuitry which produces the required symmetrical square waveform from the output of the seven-segment decoders: Fig. 7 shows one of these circuits together with the associated waveforms. The interface circuits consist, essentially, of two c.m.o.s. transmission gates connected so that the segment output can be either in phase

with the display common output, or 180° out of phase with it.

When all inputs to the segment decoding gate are at logic 1, then its output ($a$) is logic 0; transmission gate $S_1$ is on (conducting) and gate $S_2$ is off. Under these conditions the 32 Hz waveforms at pins 28 and 26 are, therefore, out of phase and the net result is that segment $E_1$ "sees" a symmetrical square wave of 8 volts peak-to-peak, which turns it on. When this segment is

Fig. 6. Block diagram of the SCL-5424-F.

Fig. 7. Display interface circuits.
not required at least one of the inputs to the decoding NAND gate will be at logic 0, its output is logic 1 and transmission gate $S_2$ will be off, while gate $S_1$ will be on. There is now no voltage across the cell for this segment, and it is, therefore, off.

Many liquid-crystal displays for wristwatches have a colon between the hours and minutes; the Brown Boveri display used in this design, however, has only a single point. The $P$ output of the watch chip (pin 22) is driven by a 1Hz waveform, through an interface circuit, which results in the display point flashing at a 1Hz rate. This serves two purposes; it gives the wearer confidence that the watch is, in fact, operating, and it can be used as a seconds timer if one is prepared to count the seconds.

**Power supply**

The SCL-5425-AF oscillator and divider i.e., being a silicon-gate device, will operate with a supply voltage between 1.2 and 10 volts; the main logic i.e. (SCL-5424-F), being an aluminium gate device, operates with a supply of between 3 and 15 volts, while the display requires a voltage of between 1.5 and 10 volts. Three 1.4 volt mercury cells, giving a supply of 4.2 volts, satisfy these requirements. The current consumed by the SCL-5425-AF is typically between 2 and 5µA at 1.4 volts; above this the current rises quite rapidly, until at 4 volts it will be about 15µA. This is far too high for prolonged battery life, so this i.e. is supplied by only one of the three cells. The main logic i.e. and display together consume a maximum current of 300mA (0.3µA) at 4 volts; the total current required by the complete watch will, therefore, lie somewhere between 2.3 and 5.3µA.

The capacity of the RM-312-H mercury cells used in this design is 36mAH or, as it is more usefully quoted in this application, 4µAyears. These cells will therefore last between nine and about eighteen months. For those constructors who would like to try to obtain one of the range of Mallory watch batteries, the WH1 cell is dimensionally the same as the RM312H but it has a slightly higher capacity of 45mAH (5µAyears). (Although the capacity of the RM312H cell is given in the data as 36mAH, this is for a drain of about 2 milliams and it may in practice have a slightly higher capacity. The capacity of the WH1 is quoted for a current drain of about 50µA.)

Fig. 8 shows the circuit diagram of the complete watch, omitting only the display for clarity. (Constructional and operational details will be presented in Part 2 of the article.)

**Vision cassette and cartridge recorders**

Thorn Electrical Industries have asked us to say that the price of the Radio Rentals Contracts Model 8200 V.C.R. is £335 plus v.a.t., or £132 plus v.a.t. per annum when rented. The firm's address is now APEX House, Twickenham Road, Feltham, Middlesex.

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**HF predictions**

Magnetic activity is still reaching disturbance level on more than fifty per cent of days. This feature developed rapidly in February 1974 and has persisted to date. The same feature was present in 1973 and terminated abruptly in July of that year.

Smoothed sunspot number is now very low with rate of change near zero. It should start increasing in a few months' time.

---

**Fig. 8. Complete circuit diagram of the watch.**
**News of the Month**

**Josephson faster than transistors**

A US patent covering the fabrication of a new type of electronic switch, whose performance far exceeds that of transistors, describes a technique used to grow a thin insulating layer — only 10 to 30 atom layers thick — that is the heart of a device called a Josephson tunnel diode. (The Josephson effect was described in 1962 and patented by B. D. Josephson in *Wireless World*, October 1966, entitled “New superconducting devices”).

It has been recognized for some time that the Josephson effect could provide an extremely fast logic switch that would require very little energy. However, a device based on these effects requires an insulating layer far thinner than has previously been used in electronic devices. The new process, accredited to James H. Greiner of the IBM Thomas J. Watson Research Centre, New York, uses a glow discharge somewhat like that in a fluorescent lamp. When a gas such as oxygen is present in the discharge, an insulating oxide is formed on the surface of the sample up to a certain thickness which is dependent on the oxygen pressure and other factors. As the oxide approaches this thickness, the discharge begins to remove oxide at almost the same rate as it is produced. Thus the oxide layer quickly approaches an equilibrium at the desired thickness.

The Josephson switch device is based on a phenomenon called electron tunnelling. Because electrons appear to behave as waves as well as particles, they can penetrate or “tunnel” through a barrier such as a thin layer of insulation which would be expected to stop them according to classical physics. The type of tunnelling explained in the new device occurs only at very low temperatures, where some metals lose all resistance to current flow and become “superconductors.” It was discovered in the early 1960s that two different types of tunnelling can occur through a thin insulator separating two superconductors. In one type there is no voltage drop and the insulator acts like a weak superconductor itself. In the other type, there is a slight voltage drop across the insulator. The insulator can be switched from the no voltage drop state by a small magnetic field. (Discovery of these effects was recognised by the 1973 Nobel Prize in physics.)

**1975 Spring trade shows**

The annual trade shows for the radio and television industry took place in London during May amid an air of surprising confidence and optimism from both retailers and suppliers. Perhaps also, the fact that this was the last show to be held in the London Hotels before the move next year to the Birmingham exhibition centre was the reason for the unusually large number of exhibitors.

It was very evident, from the range of hi-fi equipment introduced, that many manufacturers believe that one way of ensuring buoyant trade prospects is to move up-market to the very expensive (and very large!) end of the scale. A typical example of this philosophy was to be found at the Cumberland Hotel, where Sony (UK) Ltd were introducing a range of esoteric audio equipment of phenomenally high price. Among these were some loudspeakers priced at about £1,500 a pair and a new reel-to-reel tape recorder priced at around £1,200!

In total contrast, AEG Telefunken were showing some brightly coloured pocket-sized portable radios that were of quite modest price. Significantly, no new developments appeared on show for television, with the exception of an increase in the incidence of slot-mask tubes and the new 20 inch screen size.

Philips were showing a rather battered prototype Teletext decoder incorporated in a large screen television and since it did not seem to be fully functional, little interest was shown by those dealers present at the time.

At the end of the five days, most exhibitors reported a steady trade with dealers buying products, but in many cases, in a cautious way. So much new technology was shown, that this will form the subject of a brief report in the next issue of this journal.

In conclusion, the overall impression gained was of a very high morale and a determination to fight back against the trends of the current decline in consumer sales.

**VAT muddle**

The introduction of the 25% VAT rate on certain so-called luxury goods appears to have resulted in a good deal of confusion not only among component suppliers and kit companies but also among advertisers using magazines such as Wireless World which are aimed at a combined audience of professional engineers and enthusiasts.

As a result we have asked the Customs and Excise for clarification on a number of common questions. The question of VAT liability requires that a distinction be made between those who are VAT registered and those who are not.

For those who are registered, VAT is recoverable, so no change will result in accounting procedures. The changes affect the non-registered purchaser who
is either a business with a turn-over less than £5,000 per annum, or is a private purchaser. The types of goods and services subject to the high rate of VAT are defined in Notice 742 from Customs and Excise and are quite clear in all instances except where relative to components and borderline products having a possible consumer application.

Taking components first; all those components forming parts of higher rated goods are subject to 25% VAT. Those that have been specifically excluded by the VAT Liability Office may be rated at the higher rate, if they are fitted as a spare part by an engineer performing a service classified at the higher rate. Examples of such components would be those that are rated to a higher specification than that normally used in, say, audio amplifiers. Hence a high power thyristor is excluded and rated at 8% unless it is used by a service engineer as a spare part in a domestic audio amplifier.

A telephone enquiry to the VAT Liability Office elicited the information that there are still many "grey" areas where final decisions on rating have yet to be made. They informed us that local VAT offices are issuing conflicting information and that enquiries concerning electronic and electrical goods not specifically mentioned in Notice 742 should be addressed via the appropriate trade association who will then negotiate with the London VAT Liability Office. It is only this office that is in the position of making final decisions on specific exclusions.

The VAT officer went on to say that in the intervening period of uncertainty, manufacturers are expected to make, their own decisions regarding the rating of their components, in instances where Notice 742 has provided insufficient information.

Finally, two specific examples of liability in relation to advertising. Notice 742 states "Appliances . . . are generally regarded as being for domestic use if they are . . . advertised in those periodicals intended for the public." We are informed by the VAT Liability Office that this is only one of the criteria applied and that where the only advertising outlet for a professional product is through a publicly available journal than this would not result in the high rate being applied. For example, public address equipment is tentatively rated at 8% (subject to discussion with the trade associations resulting in a permanent decision). Advertising in a magazine such as Wireless World would not result in the high rate being applied. Similarly professional discotheque equipment may be advertised in the Melody Maker and yet be rated at 8%.

All this note serves to show is that in areas of doubt final clarification can only be obtained from the central VAT Liability Office, preferably through a trade association.

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**Antenna system for AEROSAT**

A complete L-band aircraft antenna system of a type suitable for airliners using the projected AEROSAT aircraft/satellite communication system has been developed under contract to the Ministry of Defence in close co-operation with the Royal Aircraft Establishment and completed in time to be installed in one of RAE’s Comet 4 aircraft for the AEROSAT system trials. The trials, based on the Azores, are organised by the European Space Research Organisation and started at the end of February. NASA’s ATS6 satellite is being used to assess performance of the L-band system, which is of particular significance since in this geographical area the access angle from the aircraft to the satellite is low, resulting in appreciable noise temperature and accentuating possible multipath problems.

The antenna system consists of two antenna groups with two switching and phasing units with their associated cabling. The antenna groups are virtually flush mounted on each side of the aircraft fuselage at about 40° from the zenith. The switch units are installed in a convenient location inside the fuselage.

An antenna group consists of six slot-dipole elements, three transmitting and three receiving which give the required beam coverage with good multipath signal rejection. The slot-

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**Brema on VAT**

The following text of a telegram to Members of Parliament was issued by the British Radio Equipment Manufacturers’ Association on 22nd April.

“To all members of Parliament from Lord Thorneycroft, President of British Radio Equipment Manufacturers’ Association. The recent budgetary proposals to increase VAT to 25 per cent will have the immediate effect of causing serious and permanent redundancies in this industry. At a special meeting of the Association today member companies estimate that permanent and unavoidable lay-offs of staff would exceed 6,000 at least 20 per cent of direct labour employed by the industry. These figures in addition to the 5,000 redundancies which occurred in the calendar year 1974, and exclude serious position which will undoubtedly occur in component and associated industries. Position is grossly aggravated by the retro-active effect of VAT proposals on rental side of industry. We wholeheartedly endorse representations already made by members of the National Television Rental Association. We emphasise that enforced reduction in production capacity now cannot fail to increase the cost of imports when the market next improves, with serious consequences on balance of payments position.”

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**Is circularly polarized TV coming?**

The heading to this item was the title of a paper by M. S. Siukola presented in the IEEE Transactions on Broadcasting, Vol. BC-21, No. 1, March 1975. According to the paper, there is on the horizon a new approach to the way TV signals are radiated. The technique is circular polarization, and while this has been in use for several years in f.m. broadcasting, it is now being viewed for TV as an...
effective antidote for ghosting, spotty coverage, multipath, poor reception on whips and rabbit ears, misoriented antennas, co-channel and adjacent-channel interference — all ills that have plagued broadcasters and viewers since TV's inception. The conclusions reached in the paper indicate that the principal deterrent would seem to be the cost to the broadcaster, but this may well be completely compensated by the more solid coverage, increased viewer enthusiasm for the enhanced reception and the additional viewer audience that circularly polarized TV is expected to provide within existing coverage areas. Present receiving antennas would not be made obsolescent, but viewers would certainly benefit most by acquiring circularly polarized types. The designs for receiving antennas as well as transmitting types are available and the antennas can be built now.

Association of research contractors formed

The Association of Independent Contract Research Organizations has been formed to represent and promote the resources of seven constituent founder members. Contract research is becoming increasingly important in its scope for serving industrial, commercial, institutional and governmental clients as a world-wide business. The seven major independent organizations based in the UK serve home and overseas markets. Between them they employ 2,250 staff of whom 800 are of professional consultant status. The Association aims to communicate on behalf of its members with international and intergovernmental bodies, with central governments in the UK and overseas, with industry — representational bodies and industry groups — and with commercial organizations able to benefit from contract research services. Further information can be obtained from D. McCa. Craig, 7 Catherine Place, London SW1E 6EB.

Queen's awards to electronics

Worthy of note are the following companies who were included in those honoured by the Queen earlier this year for their contribution to Britain's technological development and exports: Beckman Instruments for export achievements during the last three years; Mullard for outstanding export success and development; Micro Consultants for outstanding work in the field of ultra high speed analogue-to-digital conversion; EMI Sound and Vision Equipment for export achievement of the brain diagnostic EMI-Scanner; Marconi-Elliott Avionic Systems for technological innovations in the sophisticated navigation and weapon aiming system NAVWASS, supplied for Jaguar aircraft; KEF Electronics for export achievement of loudspeakers; Sinclair Radionics for outstanding export achievement and for technological innovation in electronic calculators; British Aircraft Corporation for export achievement and for technical innovation in high performance flight radomes by the Reinforced and Microwave Plastics Group. Our congratulations go to all concerned.

Transistor assembly automated

A recently developed transistor assembly system that can assemble a variety of transistor types by means of pattern recognition techniques comprises a minicomputer and image processors and will ultimately have 50 wire-bonding machines with visual functions to determine chip positions of transistors fed into the machines. The system has been installed at Hitachi's Takasaki Works, in Japan.

To recognise the position of transistor chips, an artificial eye is needed to replace the human eye. For this purpose, a microscope and a TV camera are mounted on each wire-bonding machine, so that the image signal from the camera is analyzed by a combination of the image processor and the computer to give a high-speed position recognition rate of 0.2 seconds per chip average. Position data is fed back from the computer to the appropriate wire-bonding machine, whose servomechanism can stretch the gold wire between emitter and base electrodes on the chip and the corresponding outer leads. It is claimed that the production rate has more than doubled since replacement of traditional methods.

World markets decline

An extract from the 1975 first quarter and stockholders meeting report issued by Texas Instruments indicated that the decline in demand for their electronics products experienced in 1974 was generated by the longest, deepest recession since recovery from World War II. The major difference between this and past recessions is the almost concurrent decline of all major economies. In 1970, while the US economy was declining, Japan and Europe continued to grow. In contrast, late 1973 industrial production turned down simultaneously in Japan, Europe and the US.

In the US real growth has dropped for the past five quarters. High unemployment, inventory liquidation and sluggish capital spending suggest they (TI) will not reach the bottom before the third quarter of 1975, with at best a modest upturn this year and a slow recovery in 1976. Japan appears close to the bottom of its recession, spring wage negotiations.
influencing the rate of recovery. West Germany probably will have a moderate upturn in the second half of the year. France will have minimal real growth in 1975. In Italy, industrial production is falling still, but the trade balance is improving and inflation is down compared with 1974. The UK economy is expected to remain sluggish well into 1976.

**Direct-drive a.c. motor**

Matsushita, who are represented in the UK by Symot, have announced a direct drive a.c. motor for use in record turntables. The unit, which is called the FF-2000 is basically a linear motor, which drives a light-weight aluminium platter, by eddy currents. Underneath the platter is a geared magnetic speed sensor which forms part of a feedback system for speed regulation. The motor is electronically controlled and requires an a.c. supply of 18V, 40mA. Performance of the FF-2000 is claimed to be equal to or better than the existing d.c. direct-drive motors used in the Technics range (also manufactured by Matsushita). We understand that these motors will cost around £12 in production quantities, and will be available at the end of this year.

**Ceefax, Oracle — now Tifax**

Texas Instruments announced at a press conference given in May that they have now completed the design of a modular data processor which will decode the Teletext transmissions broadcast by the BBC and IBA. Although no product has yet appeared, samples of the first generation processor will become available later this year, followed by full production in 1976.

What makes the difference in the "Tifax" decoder, from those already designed by such companies as Decca and GEC, is that the Tifax unit is being offered as a p.c.b. module carrying a relatively small number of dedicated l.s.i. circuits.

The projected cost of the "Tifax" module is expected to be around £50, reducing to £10-£15 over a few years. Power consumption is said to be 3W drawn from a regulated +5V ±0.25V supply with a maximum ripple of 10mV pk-pk. Interface to the receiver is by direct connection to the R.G.B. video drive stage from 15V, 20mA current sink open collector outputs. The brightness display can be varied using three integral common base transistors. The video input requires a signal amplitude of 2-3V pk-pk, negative going sync, with an input impedance of typically 10kΩ.

New microwave distance measuring equipment, the Tellurometer MRA5, indicates directly in metres and centimetres, operation being either fully automatic or manual at the choice of the operator. Measurement can be accomplished in less than 20 seconds.

**Microwave Conference overwhelmed**

The organizers of the fifth European Microwave Conference, which takes place September 1-4, 1975, at the Congress Centrum, Hamburg, have announced the list of papers that have been selected by their technical programme Committee and paper review board for this annual event. The response to the conference call for papers has been overwhelming. Over 350 papers from 30 countries were submitted and from those, a total of 112 will be presented at Hamburg. The papers will be grouped into 20 sessions and the morning sessions will be preceded by a total of nine specially invited state-of-the-art survey papers to be given by a number of the world's leading microwave experts. Papers will be presented in English and are limited to 15 minutes duration. A book of abstracts for the conference was published during May. Further information can be obtained from the organizers, Microwave Exhibitions and Publishers Ltd, Temple House, 34-36 High Street, Sevenoaks, Kent TN13 1JG.

**TV deliveries down again**

Deliveries to UK distributors of UK-made and imported colour television receivers reached 139,000 in March, a 20% decrease over March 1974 (188,000), according to the latest statistics compiled by the British Radio Equipment Manufacturers' Association. This brought the year's total to 475,000, a fall of 23% compared with the same period in 1974 (619,000). Total monochrome television deliveries for March were 63,000, a fall of 22% compared with March 1974. BREMA members delivered 65,000 audio stereo systems in the month, a fall of 10% compared with March 1974 (72,000). This brought the year's total to 192,000, comparable with 202,000 for the same period in 1974. Deliveries of radio receivers reached 351,000 for the month, bringing the year's total to 1,032,000, compared with 1,345,000 in 1974, a fall of 23%.

**Briefly**

IEA plus Electrex. The International Instruments, Electronics and Automation Exhibition running at Olympia since 1957 and the International Electrical Exhibition (Electrex) organised at Earls Court since 1953 are to be held as a combined event, short title IEA-Electrex, in 1976, at the new National Exhibition Centre, Birmingham.

VAT late extra. An itemized list of components on which specific agreement has been reached between HM Customs & Excise and the Electronic Components Board is now available. It is recognized that there may be some individual products to which the application of these definitions is not entirely straightforward. In such cases, an individual ruling will be given by Customs & Excise, the facts being initially reported to the Electronic Components Board.

**Teletext**

We plan to publish in the near future a short series of articles on the Teletext television information system, culminating in a design for a decoder for use with domestic receivers. Teletext is a unified version of the BBC's CEEFAX system (Wireless World, May 1973, p.222) and the ORACLE system developed by the IBA (July 1973 issue, p.314). Test transmissions were started by the BBC in September 1974 on BBC1, while a group of independent television companies (London Weekend, ITN and Thames) will be starting them in July 1975. The Teletext broadcasting standard was outlined in News of the Month, November 1974 issue.
Active notch filters

Design theory behind the development of discrete frequency rejection circuits

by Yishay Nezer, B.Sc.

We often need to separate a wanted signal from periodic interference. This may happen, for example, when a whistle or a power-line hum is disturbing a radio programme. In simple cases a filter which has zero transmission at one discrete frequency and unity transmission at all other frequencies is sufficient. In contrast to a practical low-pass or high-pass filter, an almost ideal notch filter can be realized with only one section; moreover it can be voltage tuned or even automatically track the interference.

The major class of notch filters, both passive and active, is of the second order and has the following transfer function:

$$G(s) = \frac{s^2 + \omega_0^2}{s^2 + \frac{1}{Q_0} s + \omega_0^2}$$

where $\omega_0$ is the rejection frequency and $Q_0$ is the figure of merit of the filter, which is given by $Q_0 = \frac{\omega_0}{\Delta \omega}$ where $\Delta \omega$ is the rejection bandwidth defined by the 3dB attenuation points. Some typical frequency response curves are plotted in Fig. 1 with $Q_0$ as the fixed parameter.

Many passive notch networks are known. This article will deal mainly with RC networks, since the use of coils is inconvenient, particularly at low frequencies. The best-known RC notch network is the symmetric twin-tee illustrated in Fig. 2(a), for this network $\omega_0 = 1/RC$ and $Q_0 = 1/4$. The function is, of course, realized only if the network is fed by a voltage source and subjected to an infinite load.

Another well-known notch network is the Wien bridge. This network is characterized by $\omega_0 = 1/RC$ and $Q_0 = 1/3$. The left side of the bridge shown in Fig. 2(b) is composed of equal resistors and capacitors and, in order to obtain an infinite null, the other two resistors must satisfy the relationship $r_1 = 2r_2$. The notch response, however, can be achieved even if the corresponding components in the reactive side of the bridge are not equal. The rejection frequency will then be $\omega_0 = 1/\sqrt{R_1 R_2 C_1 C_2}$ but the ratio $r_1/r_2$ will no longer equal two. An important special case occurs when $R_1 = 2R_2$ and $C_1 = C_2/2$; we then have $r_1 = r_2$.

Fig. 1. Normalized phase and magnitude response curves of a notch filter for several values of $Q_0$. 

![Diagram showing normalized phase and magnitude response curves of a notch filter for several values of $Q_0$.]
A drawback of the above two networks is that in order to vary the centre frequency and still maintain the infinite null, two or three closely matching ganged variable components must be used. Several RC bridge networks are known in which a single component is sufficient to control the rejection frequency. However, their practical significance is limited because the frequency response becomes severely asymmetric as the rejection frequency is varied.

A more acceptable variable network was proposed by Hall. It is shown in Fig. 2(c). This network can be tuned by means of a single potentiometer and the tuning law is \( \omega_0 = 1/RC \sqrt{a(1-a)} \) which in theory spans the whole frequency range. In practice the tuning range is quite limited due to the extreme non-linear dependence of \( \omega_0 \) on \( a \). However, this network has unity gain on both sides of the null frequency, irrespective of the tuning.

However, unlike the twin-tee and the Wien bridge it is asymmetric on a logarithmic frequency scale. This follows from the fact that the transfer function of this network is not given by expression (1) but contains an additional real pole and real zero.

A similar potentiometer tuned null network based on the twin-tee was proposed by Andreev.

All the networks discussed so far are characterized by low selectivity. In fact, no passive RC notch network, however complex, is capable of achieving \( Q_0 \) higher than 0.9. If the notch filter must be passive, a relatively high \( Q_0 \) may be achieved by including an inductance as in the bridged-tee network shown in Fig. 3. In order to achieve a complete null this network must satisfy the two conditions:

\[
\omega_0^2 = C_1 + C_2/LC_2 \text{ and } \omega_0^2 = 1/rRC_1C_2
\]

The figure of merit will then be \( Q_0 = 2 \omega_0 L/r \), i.e. proportional to the quality factor of the coil.

**Active notch filters**

As has been mentioned above, passive RC notch filters suffer from a low selectivity. A theoretically unlimited selectivity can be obtained by the use of active notch filters. These can be built by various active realizations of the transfer function given by expression (1). Simple active circuits are based on passive null networks in which the selectivity is raised by means of negative feedback.

One such scheme is shown in Fig. 4 and the effect of feedback can be explained as follows: When the feedback loop is open the network is simply a passive null network with a passband gain of \( A_P \), represented by curve (a) in Fig. 5. When the feedback loop is closed, the network tends to maintain a voltage gain of \( A_P (1 + A_o) \). However, it fails to do so where the forward gain is low, i.e. in the vicinity of \( \omega_0 \). As a result, the response curve is compressed as shown in curve (b) and the rejection band is narrower. As an additional benefit, the active filter can now be cascaded without being subjected to loading.

The calculated transfer function of the active notch filter is:

\[
G(s) = \frac{A_P}{1 + A_P \cdot s^2 + \omega_0 \omega_2/(1 + A_Q Q_0 + \omega_0^2)}
\]

A different realization is shown in Fig. 6 which relies on a single, less-than-unity gain amplifier. It can be seen that there are two feedback paths in the configuration, a positive unity-gain feedback which renders the effective gain of the amplifier equal to \( K/(1-K) \) instead of \( K \), and a negative feedback which subtracts the output voltage from the input. If \( K/(1-K) = A_o \), this method is equivalent to the former and the transfer function is:

\[
G(s) = \frac{s^2 + \omega_0^2}{s^2 + \omega_0^2(1-K)/Q_0 + \omega_0^2}
\]

in which the selectivity is multiplied by \( 1/(1-K) \).

**Practical circuits**

The simplest amplifier for the above method is the emitter follower. How-

![Fig. 2. Three RC null networks: (a) the symmetric twin-tee, (b) the Wien bridge, (c) a potentiometer-tuned network.](image)

![Fig. 3. Bridge-tee RCL null network, the selectivity of which depends on the Q factor of the coil.](image)

![Fig. 4. Basic active configuration for enhancing the selectivity of passive notch filters.](image)

![Fig. 5. Frequency characteristics of the network in Fig. 4 (a) open loop, (b) closed loop.](image)
ever, it is not very suitable because in order to prevent the null network from being loaded by the relatively low-input impedance, the resistors must be relatively low and the capacitors must be correspondingly large.

It is obvious possible to replace the transistor by an f.e.t. and use smaller capacitors, but the increase in the selectivity would be limited due to the smaller gain usually associated with the f.e.t. The bootstrapped source follower benefits from high-input impedance and also a gain closer to unity and is shown in Fig. 7 together with the network of Fig. 2(c).

The networks discussed above have ideally an infinite attenuation at the notch frequency. Practically, the attention is limited by the tolerances of the components and is typically 40dB for 1% tolerance. This figure may be exceeded by trimming and is ultimately limited by stray capacitance.

The Wien bridge is attractive owing to its simplicity. However, it is not a three-terminal network, and cannot be activated directly.

The circuit in Fig. 8(a) is a Wien bridge built around an operational amplifier. In spite of being active, the factor of merit is only ½ instead of ½ in the passive bridge (it does not belong to either of the schemes shown in Figs. 4 and 6) yet, being a three-terminal network, its selectivity can be improved as shown in Fig. 6. Since the output impedance is already zero an additional buffer amplifier is unnecessary, so that we only have to decrease the gain to below unity by a voltage divider and close the feedback loop at r2. The network Fig. 8(b) then contains an equivalent amplifier whose gain and output impedance are:

\[ K = \frac{R_4}{(R_1 + R_4)} \text{ and } R_3 \| R_4 \text{ respectively.} \]

Accordingly, the latter value must be subtracted from r2 or must be much smaller. Alternatively the voltage divider can be buffered as in Fig. 8(c).

If we use \( V_0 \) as the output of the filter instead of \( V_{out}' \), an advantage results with respect to the network of Fig. 6, in that the passband gain is unity instead of K. However, as the internal amplifier's gain is \( K = \frac{R_4}{R_3} \times R_e \), the factor of merit will be

\[ Q = Q_0(1 - K) = (1 + R_2/R_3)/2 \]

and the rejection bandwidth can be varied by means of either \( R_3 \) or \( R_e \). The null frequency can be varied, for example, by \( R_3 \) and the notch depth can then be adjusted by means of \( R_e \).

If it is desired to vary the rejection frequency over a wide range, it is best to vary simultaneously resistors \( R_1 \) and \( R_2 \).

If the tracking is good, the null will be maintained throughout the tuning range without further adjustments.

**High Q notch filters**

At a frequency \( \omega = \omega_0(1 \pm \epsilon) \) close to the resonant frequency, that is \( \epsilon < 1/2Q \), the response of a notch filter can be approximated by two...
straight lines with slopes ±2Q, and the filter can be used as a frequency discriminator. If Q is large, very small frequency deviations can be observed. On the other hand, if a high Q notch filter is used to reject a power-line hum, for example, a slight deviation of the frequency from its nominal value will suffice to render the attenuation excessively low. In this case a filter with an infinite attenuation over a band of frequencies would be desirable. However, such a filter cannot be realized. It is then possible either to lower the Q of the filter or to stagger-tune two or more filters in cascade and obtain a frequency response as in Fig. 9. A more elaborate solution is to use an interference tracking notch filter as suggested in the last paragraph.

When dealing with high Q notch filters, the increasing sensitivity of the notch symmetry and depth to the tolerance of the passive components becomes a serious problem. A practical solution is to use stable capacitors and trim resistors for the required notch frequency and depth. It has been seen that the most suitable network from this standpoint is the Wien bridge, but for ultimate stability the state variable filter (see below) is required owing to its extreme stability.

Another problem in realizing high Q notch filters is the roll-off in the open loop gain of operational amplifiers at high frequencies. It is found that there is a limitation on the maximum possible Q which is inversely proportional to the rejection frequency. This limitation may cause asymmetric frequency response (Fig. 10), even if the values of the passive components are accurate. With the increase in Q, the size of the "hump" increases to the appearance of oscillations. If the notch frequency is preset this can be rectified by adding an RC phase leading section at the input of the amplifier of Fig. 6 and experimentally adjusting the time constant. If the notch frequency must be variable, the only remedy is to use a wide bandwidth amplifier. However, at fairly high frequencies high Q inductors are available, and a passive filter such as the one in Fig. 3 may be preferable.

**Simulated inductance**

The circuit of Fig. 11(a) is an active bridge similar to that shown in Fig. 8(a), but has a series resonant circuit in one of its arms. If \( R_1/R_2 = r_1/r_2 \), the circuit will behave as a notch filter whose rejection frequency and factor of merit are

\[
\omega_0 = \sqrt{L/C} \quad \text{and} \quad Q_0 = \omega_0 L/R_2
\]

(6)

In order to avoid using an inductance, the series connection of \( R_2 \) and \( L \) can be replaced by the circuit of Fig. 11(b) which is equivalent to a coil whose value is \( L = C_2 R_2 \alpha (1 - \alpha) \) in series with a resistor \( R_2 \). The resulting notch filter has a rejection frequency and factor of merit given by:

\[
\omega_0 = \sqrt{C_1 C_2 R_2^2 \alpha (1 - \alpha)} \quad \text{and} \quad Q_0 = \alpha (1 - \alpha) \sqrt{C_2/C_1}
\]

(7)

respectively; the tuning law is thus exactly the same as that of Fig. 2(c).

This network can also be tuned by means of \( C_2 \), which may consist of a small trimming capacitor if the simulated inductance is made appropriately large.

**State variable filters**

A unique active notch filter may be realized by the so-called state variable method. This method is based on a multiple feedback network containing integrators and adders. In spite of the rather large number of operational amplifiers, the number of capacitors needed for the realization of any arbitrary transfer function is minimal.

The basic building block shown in Fig. 12 simultaneously provides three transfer functions of the second order; these are high-pass, band-pass and low-pass function, given by:

\[
V_H(s) = K s^{1/2} \left( s^2 + \frac{\omega_0^2}{Q_0} s + \omega_0^2 \right) \quad \text{and} \quad V_L(s) = K s^{1/2} \left( s^2 + \frac{\omega_0^2}{Q_0} s + \omega_0^2 \right) \quad \text{and} \quad V_I(s) = K s^{1/2} \left( s^2 + \frac{\omega_0^2}{Q_0} s + \omega_0^2 \right)
\]

(8)

The resonant frequency \( \omega_0 \) is determined by the time constants \( T_1, T_2 \) of the integrators, and is given by \( \omega_0 = 1/\sqrt{T_1 T_2} \). The factor of merit is \( Q_0 = 1 / K \sqrt{T_1 T_2} \), where \( K = 1/(1 + R_2/2R_1) \).

This configuration does not provide complex zeroes and in order to obtain
the symmetric notch response given by expression (1) we must sum the high-pass and low-pass outputs. The filter obtained contains four operational amplifiers but has the following characteristics: if \( \tau \) varies, the rejection frequency is varied while the bandwidth \( \Delta \omega = \omega_0 / Q_0 \) remains unchanged; if \( \tau \) and \( \tau \) vary simultaneously, the rejection frequency varies linearly while the factor of merit \( Q_0 \) remains unchanged; if \( K \) is varied with the aid of the resistor \( R_0 \), the rejection frequency remains unchanged and the rejection bandwidth and gain alone will change; the network is quite insensitive both to the values of the passive components and to the gain of the amplifier. Its stability approaches that of passive filters.

The transfer function of a conventional integrator is: \( G(s) = 1 / s \) \( \tau \), and the variation of \( \tau \) can be obtained by varying \( R \) or \( C \). If we connect an amplifier with a gain \( K \) in series with the integrator, the transfer function changes to \( G(s) = K / s \tau \); i.e., \( \tau \) is decreased without altering \( R \) or \( C \). If an analogue multiplier is substituted for the amplifier, an integrator is obtained, in which the time constant \( \tau \) is dependent on a control voltage. A notch filter with a constant \( Q \) and rejection frequency directly proportional to the control voltage can thus be built from two such integrators.

It has already been mentioned that the maximum \( Q \) factor which can be obtained in an active filter realization is limited – for a given resonant frequency – by the bandwidth of the operational amplifier. The state variable method is no exception to this rule. Design considerations resulting from these limitations were discussed by Thomas.

**Bandpass filter synthesis**

A common disadvantage which is probably shared by all accepted active bandpass realizations is that the bandwidth is inversely proportional to the midband gain; in other words an increase of \( Q \) is accompanied by a proportional rise in gain, which is often undesirable.

A different realization of bandpass filter is given in Fig. 13(a) based on the equation:

\[
A(1 - \frac{s^2 + \omega_0^2}{s^2 + \omega_0^2 + \omega_i^2}) = A \frac{1}{sQ} \frac{\omega_i}{\omega_0^2 + \omega_i^2}
\]

(9)

which results in a bandpass transfer function whose midband gain is independent of \( Q \). The configuration operates as follows: at frequencies which are remote from \( \omega_0 \), the notch filter transmission is unity, and the input to the differential amplifier is zero. At frequency \( \omega_0 \) the notch filter transmission is zero and the input to the amplifier is unity. As equation (9) shows, the resulting bandpass filter has the same \( Q \) as the notch filter. Now it is easy to realize notch filters in which the selectivity is controlled by a single resistor, such as the one shown in Fig. 8. If such a filter is incorporated, a constant-gain variable bandpass filter is obtained.

**Interference-tracking notch filter**

It has previously been mentioned that a sufficiently narrow-band notch filter may not be effective in rejecting interference which drifts in frequency. Such interference could in principle be tracked by a phase-locked loop, and the output of the loop, which is proportional to the frequency, then applied to a voltage-controlled notch filter. A drawback of such a method is that it is an open-loop system and any residual interference at the output due to imperfect tracking is not corrected for.

In contrast to the phase-locked loop, which is a signal-tracking oscillator, a signal-tracking band-pass filter can be easily built. It is similar to the phase-locked loop and consists of a voltage-controlled band-pass filter, a phase detector and a low-pass filter. The closed loop then centres the band-pass filter on the signal by maintaining an (ideally) zero phase shift between the interference and the output of the filter. This bandpass filter can be converted to a signal-tracking notch filter as shown in Fig. 13(b), in which the filtered interference is subtracted from the original input. However, unlike conventional notch filters, the input to this filter must contain a minimum of interference for locking to occur.

**References**


**Books Received**

Dictionary of Data Processing by Jeff Maynard is designed mainly as a source of reference for those interested in, or using, computers and data processing equipment. Over 4000 terms are defined, as well as acronyms and abbreviations, in alphabetical order. The final section of the book lists British and American standards relating to data processing. Price £3.90, Pp.269. Butterworth & Co. Ltd, Borough Green, Sevenoaks, Kent T158 SPH.

Energy and Humanity edited by M. W. Thring and R. J. Crookes, as a former US Secretary of Commerce has pointed out, a finite world with finite energy resources cannot support an exponential growth rate. This book presents the problem as it exists now, with a view to what might be the situation at the turn of the century. An attempt has been made to cover the existing sources of energy along with their associated problems and to assess what might become available in the future. Much of the material has been drawn from an international conference on energy and humanity, held by the SIRS in September 1972. Price £3.50, Pp.195. Peter Peregrinus Ltd, P.O. Box 8, Southgate House, Stevenage, Herts SG1 1HQ.

Recent, updated additions to the D.A.T.A. books cover range discontinued thyristors, m.s.i., i.c.s., memories, diodes, linear i.e.s., microwave tubes, digital integrated circuits, and thyristors. As usual these books provide information on components from most of the producers in the world. Outline diagrams are provided together with pin configurations, and an index of manufacturers is also incorporated making these volumes an invaluable source of information. London Information Ltd, Index House, Ascot, Berks SL5 7EU.

The International VHF-FM Guide by Julian Baldwin and Kris Partridge. This booklet is a useful source of information for the radio amateur who wishes to operate through the VHF repeater networks both at home and abroad. Price 25p + 5p postage and packing. Pp. 38. J. F. C. Baldwin, 50 Aldbourne Road, Burnham, Slough, SL1 7NJ.
I was most interested to see that the position of the ceramic filter in Mr J. B. Dance’s F.M. tuner (March issue) was after the limiting amplifier. Surely if the filter is to aid tuner selectivity then it must be placed before those parts of the tuner which have non-linear transfer characteristics, i.e., before the limiting amplifier.

Assuming that the input signal level is sufficient to cause limiting (as it should be for any usable signal) then any spurious signal components emerging from the r.f. unit outside the desired i.f. bandwidth will be subjected to intermodulation by the limiting process. Any intermodulation products which appear within the i.f. bandwidth will not be removed by subsequent filtering, but will affect the audio output. Consider, for example the case where one is trying to receive a weak signal in the presence of strong signals in adjacent channels. The output of the r.f. section may contain levels of the adjacent channels of similar or greater amplitude than that of the required signal. In this design the limiter may act on these adjacent channel signals so that these subsequently affect the audio output. Placing the i.f. selectivity before limiting will reduce the levels of unwanted signals entering the limiter, and reduce audible effects due to them.

Intermodulation in the limiter may explain the poor performance observed by the author on an r.f. unit with a high level of local oscillator output.

Note that some filtering after the limiter may be necessary to avoid harmonics of the 10.7MHz i.f. signal affecting the operation of the second mixer.

J. E. Marshall,
Bilborough,
Nottingham.

Mr Dance replies:

I am grateful to Mr Marshall for raising an important point. The NE563 limiter has a bandwidth of 22MHz (typical) and a gain of 60dB. The filter placed between the limiter output and the mixer will reject any noise which lies outside the required passband and will also attenuate any harmonics generated in the limiter.

If one wishes to have a tuner which provides optimum reception of weak signals in a crowded band, then additional filtering is required before the limiter. However, the front-end specified in my article contains a double tuned 10.7MHz circuit before the emitter follower output stage and I felt that the overall selectivity was adequate for normal domestic reception.

I did experiment with an additional FM-4 filter between the front-end and the limiter input. This involved the use of a matching resistor (270 ohms) between the front-end and the filter and a 330 ohm resistor on the output side of the filter. An active circuit is required between the filter output and the limiter, since the input impedance of the latter is quoted as 135 ohms. An active circuit providing gain will also improve the a.m. rejection. However, it seemed that the published circuit was satisfactory for domestic reception and I felt its simplicity would appeal to many readers.

With reference to Mr Marshall’s penultimate paragraph, I did try an extra FM-4 filter with the front-end which gave unsatisfactory results, but I still did not obtain audio output of satisfactory quality.

Dr A. Tip of FOM-Instituut Voor Atoom-En Moleculairefysica, Amsterdam, has kindly sent me details of his unpublished work with a more complex system which he uses to receive many UK transmissions. Unfortunately only a few brief details can be mentioned here.

Dr Tip employs a BFY90 amplifier by his aerial, and a Valvo FDIA front-end which feeds a 3028A variable gain cascode stage. Four ceramic filters are used, followed by a LM733 stage (set for a gain of 100); the latter feeds the NE563 limiter. The a.g.c. output from the NE563 is fed to a transistor amplifier which provides a.g.c. for the 3028A stage; the amplified a.g.c. also feeds a signal strength meter which is unaffected by the setting of the 563 muting circuit.

In conclusion, may I mention that the NE563 masks are being modified in California? The new NE563 devices will have a different oscillator impedance and should operate more reliably with the Taiyo CR-9.8 ceramic resonators mentioned in my article; the parallel 2.2kiloohm resistor and 5pF capacitor should not be required with the new NE 563. In addition, Signetics expect the new device will produce an output with less distortion, but no figures are available at the time of writing. In their current data Signetics suggest a value of 5.1 kilohm for R4 of my Fig. 1; this will give a slightly improved signal-to-noise ratio and a reduced bandwidth, but the difference is not very noticeable.

J. B. Dance,
Alcester, Warwickshire.
optimize on/off ratio on the one hand, distortion and input impedance on the other.

As a postscript, a useful configuration when, for example, equalization is to be switched, is shown in Fig. 2. In this circuit, using low impedance termination of the feedback networks at both ends, the required equalization time constant, \((R_1 + R_2)C_1\) or \((R_1 + R_2)C_2\) in Fig. 2, is selected by turning the f.e.t. (T1 or T2 respectively) off, and all others on. Thus all feedback paths except that required are shunted to ground. If \(R_2\) and \(C\) are of the order of 1000kΩ and \(T_1\) and \(T_2\) are T1S3L, attenuation of the unwanted f.b. is \((25 + 100k)\)/25, greater than 70dB. Distortion is low because the selected f.b. path uses an f.e.t. in the "off" mode, and contributions from f.e.t.s in the "on" mode are at very low level.


**COMPUTER POWER**

It is your privilege as editor to don the mantle of Cassandra from time to time; and we may have cause to be grateful to you for warning us (leader, May issue) of the latent threat of computer power. Readers of your editorial may however care to join me as I hasten to bury my head in the sand. Or, as it might be, paper; since waste paper is the most obvious product of today's computers. What can be discovered in this mountain of paper is that a computer with a remote terminal and interactive conversational language can take all the arithmetic grind out of hitherto laborious calculations in radio and electronics. And this generally involves only a minimal knowledge of computer programming — rather less, I suspect, than is currently mastered by schoolchildren. Indeed, readers who have no direct access to such computing facilities might well brine some and daughters to compute for them.

There is of course an additional bonus in speed and accuracy. For example, given the radio frequency range required together with the intermediate frequency, one can calculate component values for superheterodyning reaching to an accuracy much higher than the tolerances of actual components in less than five minutes. A similar time is required to plot the actual tracking error curve. If one then requires information on where the tracking curve moves to at the limits of specified component tolerances, another five minutes will suffice. Most of this total of fifteen minutes is spent by the system in organising itself and in printing — actual computing time may be only two or three minutes.

The story is very similar if one wishes to perform calculations on coupled tuned circuits — the complex arithmetic takes much less time than printing the (unnecessarily accurate) results.

Any reader who has problems with tedious calculations of this kind is welcome to contact me.

D. P. C. Thackeray, Department of Chemical Physics University of Surrey, Guildford, Surrey.

**ELECTRODYNAMICALLY INDUCED E.M.F.**

"Cathode Ray's" article and Mr Meade's letter (February issue) have prompted some questions concerning a well known problem in school-boy physics:

"If an aeroplane flies a level course through a vertical component of the earth's magnetic field, then its wings act as a flux cutting conductor and an e.m.f. is induced in them. Can this e.m.f. be measured and, hence, the speed of the aircraft deduced?"

The answer is, of course, no, because as "Cathode Ray" stated, connecting a voltmeter to the ends of the conductor (in case the wings) would form a closed loop embracing a constant magnetic flux.

Not being well versed in the theory of electromagnetic induction, we would like to ask the following questions:

"Can the induced e.m.f. be measured if the meter leads are screened with a high permeability material?" and, naturally, "If not, why not?"

D. C. E. Todd and N. G. S. Taylor, Brunel University, London.

**DIRECTORY OF AUDIO COURSES**

The Audio Engineering Society is preparing a Directory of educational facilities and institutions in its field of interest, which embraces sound recording and reproduction, instrumentation, sound reinforcement etc. This information will appear in the Journal of the AES and be included in a career book eventually to be prepared by the Society.

As the UK representative of the AES Education Committee, I have been asked to act as a focal point for the collection of such information from schools, colleges and universities in the British Isles.

To simplify the listings, I can supply copies of a short questionnaire and I would urge all organisers of courses related to audio engineering to apply to me for copies as soon as possible.

John Borwick, Senior Lecturer, Recording Techniques University of Surrey, Guildford, Surrey.

**INSTRUMENT READ-OUT IN BRAILLE?**

I have recently become interested in the problems of teaching science to blind pupils. Present techniques offer analogue derived information in tactile or audible form. For example in typical science experiments one might wish to indicate the height of a liquid column or the reading of a meter. The liquid level would be located by a light probe whose audio tone would alter on discovering the surface; the meter pointer would be located either by the light probe or feeling the pointer in an instrument made specially for the blind. The actual reading is then taken from an adjacent Braille scale.

Of necessity these procedures are slow, clumsy and very approximate. It occurs to me that in the case of meter readings digital techniques could be exploited with considerable advantage. Many readouts are available in digital form — voltmeters, ammeters, frequency meters, multimeters, balances, calculators — and in many cases the electronics involves binary processing of numbers which in b.c.d. form is decoded for a visual readout such as a seven-segment device. Could not the b.c.d. be decoded direct to Braille format and used to operate a tactile display? It should be possible to devise an electromechanical Braille readout as only four pins are required as shown in the diagram. My guess, based on no direct experience, is that blind pupils would find the reading of electrical quantities more direct and so more meaningful. In the case of balances the accuracy of the weighing would certainly help science teaching. If it is proved possible to use this idea with electronic calculator chips, perhaps a new era in teaching maths to the blind might follow also.

Recent progress in artificial vision, reported in your April issue, raises an even more powerful possibility, namely direct excitation of the optic nerve with numerical data. Whether this should be b.c.d., seven-segment, Braille, or some other format is a matter for speculation. Braille has already been used successfully (WW, April p.157) for text by this means.


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Binary coded decimal compared with numbers in Braille.
Intended mainly for hiss reduction in magnetic-tape recording machines, this noise reduction unit can be switched to decode commercially-available Dolby B-encoded cassette tapes, Dolby B-encoded f.m. radio transmissions (current in the USA), or to encode blank tapes from any source. As an alternative it can be used in trading some of the noise improvement for reduced distortion at peak recorded levels. Part 1 in the May issue gave background to the Dolby system and part 2 gave circuit and constructional details together with some suggestions for circuit options and alignment procedure. This part shows how to set-up the kit version design without using additional equipment and gives calibration procedure.

3 — Kit alignment and calibration

by Geoffrey Shorter

Constructors who build a Dolby-B processor without using the full WW kit have the option of using the power supply included in the circuit of Fig.12 or of using an alternative one, for instance one built into existing equipment. Component values for the circuit of Fig. 12 have been optimized to provide an overload margin of 16dB (equivalent to 1200Wb/m on open-reel) for a 15-volt supply, but voltages between 15 and 24 volts could be used provided component voltage ratings are chosen appropriately. The main requirement is that supply ripple be less than 200µV r.m.s. Current consumption at 15 volts is 20mA per processor; with IC1 and IC2, it is 30mA. The voltage regulator IC2, whose output is 15 volts ±5%, is essential if the meter calibration oscillator of Fig. 14 is used. Input to the regulator should be not greater than 25V and not less than 18.25V.

Kit setting-up procedure

The procedure for setting up the kit design is a little more elaborate than the basic alignment instructions because it is designed to eliminate necessity for additional equipment i.e. a.c. millivoltmeter and variable-frequency a.f. oscillator. It therefore includes a facility for generating a 5kHz circuit alignment tone, as well as a 400Hz calibration tone. Two meter amplifiers, and a 580mV source (1kHz oscillator) to calibrate the meters, are included to obviate the need for an a.c. millivoltmeter.

In using the in-built meter scale in setting up, it is better to use close-tolerance resistors in an attenuator so that all measurements can be made at one meter reading (0dB). Errors in meter reading are minimized by this technique, and errors due to an inaccurate scale eliminated.

Right-channel meter calibration

The unit is aligned using part of IC1 as a meter calibration oscillator. The amplifier section of IC1, based on pins 10, 11 and 12 is first used as shown in Fig. 14. In this mode the amplifier is wired as an astable multivibrator switching between the 15V supply rail and 0V, with a mark-to-space ratio of about 1:1 and a frequency of around 1kHz. The real voltage swing is a little less due to saturation voltages, but is highly repeatable from one sample to another.

Typical performance

Noise reduction: better than 9dB weighted
Clipping level: 16.5dB above Dolby level (measured at 1% third harmonic content)
Harmonic distortion: 0.1% at Dolby level (typically 0.05% over most of band, rising to a maximum of 0.12%)
Signal-to-noise ratio: 66dB (20Hz to 20kHz, signal at Dolby level)
Approximate voltage readings (AVO 8)

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Connect resistor R5 (3.9MΩ) from the pin at R51 to pin 2 or the L2' position.
-Wire R56 (10kΩ) in parallel with R47 (1MΩ) across the pins at R47 position.
-Form an attenuator with R59 (110kΩ 2%) and R68 (10kΩ 2%) in series, Fig. 14, earth the end of R61 by connecting to pin 3 of L2' and connecting R60 to pin 1.
-Solder one end of R58 (330kΩ 2%) to R25 (330kΩ) to their pins. Take the other end of R58 to the junction of R49, R61 (R155 remaining floating). Switch on.
-Adjust R6 (Fig. 15) until the r.h. meter reads 0dB. Switch off.
-Remove R55, R58, R59, R60, R61 and do not alter the setting of RV4.

Circuit alignment

The now-calibrated r.h. meter is used to set the gain and f.e.t. bias controls of both left and right processors with the help of a 5-kHz oscillator, Fig. 14, adapted from the 1-kHz oscillator circuit by using arrangement (b).
-Solder C30 in position, removing and replacing the p.c.b.
-Solder L1 on to pins 1 and 2 of the L2' position. Gently screw in the core.

Right-channel circuit alignment

-Connect R65 (10kΩ 2%) between the R55 pin and test point 1 (TP1) on the sub-board.
-wire the oscillator pin, marked "osc." to the sub-board pin marked R' (input to processor).
-Set RV2 (oscillator level) fully anticlockwise. Check that no plugs are connected into the sockets.
Set RV2,102 fully anticlockwise. Switch on.
- Select the auxiliary position for SW3. Set the balance control RV9 to mid-position and the input level control RV10 fully clockwise.
- Ensure the calibration tone switch SW5, the noise reduce switch SW4, and the 19-kHz filter switch SW4 are in the off position (out), and the check tape switch SW6 is in the normal position (out).
- Check that the f.e.t. gates have previously been shorted to ground by two looped links.
- Turn the law control RV1 fully clockwise to pinch-off f.e.t.
- Switch SW3 to record and adjust RV3 until the meter reads 0 dB (equivalent to 17.5mV at TP1). Switch off.
- Transfer the end of R61 from TP1 on the sub-board to TP2 and switch on. Meter should read within ±1dB of the previous, 0-dB reading. Note actual reading *. Switch off.
- Solder R59 (15kΩ 2%) and R63 (6.8kΩ) in series with R61 (i.e. between the R65 pin and TP2), decreasing meter sensitivity by 10dB. Switch on and check meter reading reduces by roughly two thirds.
- Switch on noise reduction, SW4 and adjust RV2 (gain) to bring back meter reading to that noted above at *. Switch off.
- Cut the f.e.t. gate short for the right-hand channel with wire cutters and short-circuit R63, increasing meter sensitivity by 2dB. Switch on.
- Adjust RV1 (law) until meter reads as noted above, at *. Switch off.
- Re-apply f.e.t. gate short and replace R63. Switch on and check meter still reads as above, at *. Switch off. Remove gate short.

**Encode/decode matching check.**
- Switch SW1 to play and switch noise reduction off, SW4.
- Short-circuit R63, leaving R61 and R42 connected. Set RV1,104 fully clockwise. Switch on.
- Adjust 5-kHz oscillator output level control RV5 until meter reads 0dB (equivalent to 44mV at TP2). Switch off.
- Switch noise reduction on, SW4. Short-circuit R62 and R63 so that only R61 is in circuit. Switch on. Meter should read 0dB to within ±1dB. Switch off.

**Left-channel circuit alignment.** Now repeat this process for the left channel, starting from the point of connecting R62 between the R65 pin, (not R63), and the test point — now to be TP101 — on the sub-board. Note that the right channel meter, being calibrated, is still used in setting the left channel, and that TP101 is to be read for TP1, TP102 for TP2, RV101 for RV1, RV102 for RV2, and that the left-channel f.e.t. gate-shorting loop is now implied. The "osc" pin is to be connected to the point L' on the sub-board at the appropriate time.

---

Fig. 18. First part in setting up procedure for-kit version (left) shows arrangement used in calibrating the right-channel meter. For aligning the noise reduction circuit the meter calibration oscillator is changed to a 5kHz oscillator, using L1 temporarily in the L4 position (centre). Its output, via the "osc" pin, is taken to the processor input (R for the right channel). To calibrate the 400Hz oscillator, L4 is put in its normal position, the i.c. oscillator disabled, and the oscillator output taken from TP1 or TP101 (right).

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Amplitude response with and without 10kHz filter.
After repeating for the left channel, switch off. The gain and law adjustments are now complete.

- Remove the f.e.t. gate shorts, R₈₁, R₈₂ and L₂, inserting L₂ into its normal (final) location.

400Hz oscillator calibration

- Solder one end of R₃₆ to its pin and connect the other end to TP1.
- Short pins 1 and 3 at the L₂' position, and remove the wire from osc pin to point L'. Switch on.
- Switch Sw₂ to record, press the noise reduce switch Sw₄ off and switch on the 400Hz calibration tone oscillator, Sw₃.
- Adjust RV₃ (oscillator level) until the right-channel meter reads 0dB. Switch off.
- Transfer the end of R₆₁ from TP1 to TP101 and switch on. Adjust RV₁₀₃ until the r.h. meter reads 0dB.
- Repeat this procedure because of a slight interaction between RV₃ and RV₁₀₃. Switch off.

Left-channel meter calibration

- Disconnect R₆₁ from TP101 and connect the free end of R₁₅₅ to TP101 and switch on.
- Adjust RV₁₀₃ to obtain 0dB at the left-channel meter, being careful not to disturb RV₄. Switch off the calc. tone oscillator. Switch off.

19kHz filter adjustment

- Wire R₁₃₅ permanently onto the main board, replace R₆₁ with R₅₁ and connect free end to TP1.
- Connect an f.m. stereo tuner to the auxiliary input and with the aux-tuner links wired in, switch on and tune to a BBC stereo test transmission.*

Alternatively, if a high accuracy (±50Hz) 19kHz oscillator is available, connect its output to point R' on the sub-board.

- With zero a.f. modulation,* adjust the record level control RV₁₅₆ to give a 0dB meter reading. Switch the 19kHz filter on, Sw₄.
- Adjust L₂ for minimum reading on the right-channel meter. Do not adjust L₁ or L₁₃₁. Increase record level for sharper null near tuning point.

Repeat for the left channel starting by transferring end of R₆₁ from TP1 to TP101 and adjusting L₂ for minimum reading. (In using a 19kHz oscillator, connect to point L' on the sub-board and transfer R₅₁ lead from TP1 to TP101 before adjusting L₁₃₁.)

Calibration

To ensure interchangeability of all Dolby B-encoded tapes and of Dolby B-equipped machines, the voltage levels in the processors must be related to flux levels on the tape. A certain amplitude level is used that bears a fixed relationship to the noise reduction parameters and to conditions between encoder and decoder. The level chosen corresponds with a flux on open-reel tapes and cartridges of 185mWb/m, with 200nWb/m for cassette tapes, with a deviation of 37.5kHz on f.m. transmissions, and with a voltage level at the processor output of 380mV r.m.s.

This level, often called Dolby level, should not be taken to imply an operating level. If the level-setting meters in the unit are to be used as modulation-depth meters, a mark may be made on the meter to indicate the reference level. Whilst setting this level equal to 0VU on meters can often lead to reasonable modulation depths, this is not always the case: for cassette recorders it is best set at +3VU.

The 400Hz oscillator and tapes recorded with a 400Hz tone to the above level are used in calibrating units, once the circuitry has been set up. When playing or recording the standard flux level, the 580mV level is set by adjusting the play calibration potentiometers during play, and the record calibration potentiometers during recording.

Playback-only decks and units. As the signal levels on encoded tape cassettes are to be related to those in the decoder during playback only, the 400Hz oscillator is not required and calibration is achieved with a calibration cassette, containing the reference flux.
An alternative to the kit design is this single-channel processor, using the circuit of Fig. 12 but excluding power supply, alignment and calibration circuitry. (Track diagram will be given in a subsequent issue.)

Switch noise reduction off.
- Play calibration tape. Set play gain control on tape deck to 0VU on deck meter, if possible, or to mid-position otherwise.
- Adjust play cal. control for 580mV on meter or Dolby level indication, depending on meter used.

Playback gain controls on the recorder in the signal path en route to the processor input should not now be disturbed.

Switchable encode/decode processors.

Playback calibration
- Switch to play and switch off noise reduction. Connect millivoltmeter to point G if meters not built-in.
- Play calibration tape. Set play gain control on tape deck if fitted to 0VU on deck meter, if possible, otherwise to mid-position.
- Adjust play cal. control for 580mV indication.

This completes playback calibration and the play gain controls on the tape deck should not be altered. Adjust listening level with the output level control following the decoder output (as in Fig.13).

Record calibration
Start by setting record gain control on tape deck to mid-position, if fitted. (If combined with playback gain, do not adjust.)

- Switch to record.
- Fit blank tape (as recommended by maker or for which bias is correctly adjusted) and feed in 400Hz at points from external or internal oscillator. (If unit has been built into cassette machine and 400Hz input is via line input socket, adjust record level control so that meter reads 580mV, or Dolby level.)
- Record on tape for a few seconds, rewind and playback, switching to play on the noise reduction circuit as well as on the deck. Note whether meter shows about or below 580mV, or Dolby level.
- Make small adjustment to record cal. controls in appropriate direction and record 400Hz tone again, observing meter reading on playback. Repeat this procedure as many times as necessary to obtain correct reading.

This completes record calibration for tapes. If the circuit of Fig.13 or similar has been adopted, recording level is adjusted with record balance and level controls on the noise reduction unit, the level being judged by the tape deck's normal meters.

When the noise reduction unit is connected to a three-head machine with a simultaneous monitoring facility the tape signal may be monitored in its encoded form by operating the check tape switch.

Simultaneous encode/decode circuits.
Constructors with three-head machines having a simultaneous monitoring facility can use single-processor boards permanently wired in the encode and decode modes. If provision for encoded f.m. transmissions is required switching must be arranged so that encoding does not take place during recording. A monitor switch can be provided at the input to the decoder, to switch from tape, via a play cal. potentiometer, to source i.e. a connection to the encoder output via a 580-30mV attenuator, Fig.19.

Playback calibration procedure is as above, but record calibration is simplified.
- Set record level controls on tape recorder to mid-position. Set monitor switch to tape.
- Record on blank tape, operating the calibration tone switch or injecting a 400Hz tone from an external oscillator.
- Adjust record cal. control so that meter reads 580mV, or Dolby level.

FM calibration.
If you wish to set the controls for encoded f.m. transmissions, currently being transmitted by stations in the USA, an approximate calibration can be achieved by tuning to a local station, switching to f.m. or Dolby f.m. and setting the f.m. cal. control to give meter readings similar to those obtained when playing pre-recorded tapes. More accurate adjustment can be obtained if a station can be received which transmits the 400Hz calibration tone, identified by a characteristic warbling, or alternatively by using an f.m. generator. In this last-mentioned case, modulation frequency should be set to about 400Hz with a peak deviation of 37.5kHz (not including pilot tone).

Change of time-constant for encoded f.m. transmissions
There are two commonly used pre-emphasis time constants, 50µs and 75µs. Under certain conditions, these values can lead to reduced modulation at low and medium frequencies or severe amplitude distortion at high frequencies. In the USA the FCC has approved Dolby Laboratories' proposal of using 25µs for encoded transmissions, and to receive such broadcasts it is necessary to alter the de-emphasis time constant. In the circuit of Fig.13 this is achieved with components R_x and C_x values being given in the component list on page 259 (June) for the change from 75 to 25µs and for a change from 50 to 25µs (not yet authorized in 50µs countries). When recording such broadcasts the encoding function of the noise reduction unit is clearly not required and the "Dolby f.m." switch position automatically switches off the encoding function. Application of the Dolby B system to f.m. broadcasting is discussed in two articles in the Journal of the Audio Engineering Society, June 1973, pp.351-62, and briefly in the July 1974 issue of Wireless World, page 237.
Using the unit

The calibration procedures described theoretically apply to the one tape speed used during calibration. Whether calibration will hold for different tape speeds depends on the design of the deck, so check calibration when speed is changed. The calibration tape available can be used at 4.75 and 19cm/s, as well as 9.5cm/s. (For 38cm/s tape speed, where the noise spectrum is wideband, applying the B-type system may result in the remaining mid and low-frequency noise becoming more apparent). When the brand of tape is changed it is usually necessary to readjust the record cal. controls, the play cal. setting remaining unchanged. The characteristics of cassette tapes are more critical, and changing brand will normally require adjustment of bias (and equalization when using Cr02 tapes).

When the unit is connected to the normal input and output points of a tape recorder, the recorders own input and output controls from part of the calibrated system. The settings used during calibration should not be disturbed, input and output level controls being provided on the noise reduction unit, and it is a good idea to mark the tape recorder control settings.

The amplitude response of the tape recorder must be flat and its gain unity, measured between point G of the processor in record and play, to ensure correct operation, so that the signal voltage in the decoder is the same as that at the encoder (to within 2dB). If there is a bandwidth restriction between encoder and decoder, e.g. if the response of the recorder does not extend up to at least 10kHz, a non-complementary situation arises, unless of course the encoder input bandwidth is similarly limited.

In using the unit don’t forget that it will only reduce noise generated after the encoder and before the decoder. If the input signal is noisy in itself, or is made noisy by poor circuitry prior to encoding, this noise will be reproduced unaltered along with the signal. In some cassette decks, the line inputs are attenuated prior to amplification by a sometimes noisy microphone pre-amplifier.

As the sensitivity of the processor is of the order of 30mV, a line input amplifier is not required when the circuits are built into a tape recorder, and the input signal should be taken directly to the input gain control via a switch, or socket with switch, to disconnect the microphone pre-amplifier. It’s a good idea too to make sure any automatic level limiter operates only in the microphone input and not in the line input.

Complete kits for the Wireless World Dolby B noise reducer are available through the address given below.
The two-channel design features:
- a weighted noise reduction of 9dB
- switching for both encoding (low-level h.f. compression) and decoding
- a switchable f.m. stereo multiplex and bias filter
- provision for decoding Dolby f.m. radio transmissions (as in USA)
- no equipment needed for alignment
- suitability for both open-reel and cassette tape machines
- check tape switch for encoded monitoring in three-head machines

The kit includes:
- complete set of components for a stereo processor
- regulated power supply components
- board-mounted DIN sockets and push-button switches
- fibreglass board designed for minimum wiring
- solid mahogany cabinet, chassis, two meters, front panel, knobs, mounting screws and nuts.

Price is £43 inclusive.

A single-channel printed-circuit board, with f.e.t. costs £2.50 and £8.63 with all components inclusive (excluding edge connector, £1.37 extra). Selected field-effect transistors cost 68p each inclusive, £1.20 for two and £2.20 for four.

Calibration tapes are available, costing £1.94 inclusive for 9.5cm/s open-reel use and for cassette (specify which).

Send cash with order, making cheques payable to IPC Business Press Ltd., to:
Wireless World noise reducer
General sales department
Room 11, Dorset House
Stamford Street
London SE1 9LU
Allow three weeks for delivery.
A 50 MHz oscilloscope

3 — E.h.t. oscillator, power supply and tube circuit

by C. M. J. Little, B.A.

Department of Electronics, Southampton University

The requirements of the c.r.t. are a negative supply of 1kV with a current capability of 2mA at maximum brilliance, and a positive supply of 3kV, at a current of 50uA. The e.h.t. supply must be stabilized in order to avoid changes in X and Y plate sensitivities with brilliance. Some oscilloscope designs use mains-derived e.h.t. and rely on the large current capability of the supply to avoid changes in voltage with loading. There are two major disadvantages of this system. First, the need for large-value smoothing capacitors, and second, the danger of lethal electric shock. This design uses the alternative, which is a transistor inverter operating at about 20kHz. Smoothing is easy at this frequency, and feedback stabilization can be used.

E.h.t. generator

The circuit is shown in Fig. 11. Tr82 and Tr83 form a current-switched class D oscillator. This type of circuit produces a sine wave, but has the high efficiency usually associated with square-wave inverters. The transistors act as switches with, ideally, zero voltage across the transistor for half the cycle. The waveforms are illustrated in Fig. 12. L7 provides a constant current at the frequency of oscillation, and is uncritical as to exact value. The criterion is that its impedance at the working frequency should be large compared with the static resistance of the oscillator (supply voltage divided by supply current). The output voltage of this type of oscillator is very dependent on load.

The transformer T2 resonates with its stray capacitance at about 18kHz. The secondary winding steps up the voltage to about 1.5kV peak, which is rectified to give -1kV d.c. and voltage trebled to give about +3kV d.c. The loading on the negative peaks is much greater than that on the positive so the trebler gives a higher voltage than one would expect. The feed to the stabilizer is taken from C12c in order to avoid including the extra

![Fig. 11. The e.h.t. generator.](image-url)

![Fig. 12. Waveforms in the e.h.t. class D oscillator.](image-url)
pole $R_{232}$ and $C_{123}$ in the feedback loop. The error amplifier is an integrated operational amplifier with current multiplication provided by $T_{R8}$ and $T_{R4}$. The amplifier is used in non-inverting mode and the negative e.h.t. is compared with the 18V rail. $R_{243}$ and $C_{122}$ provide a dominant pole at 30Hz. The gain at 50Hz is high enough to eliminate mains hum from the e.h.t. voltages. If a 741 type of amplifier is used instead of a 709, $R_{246}$ and $C_{121}$ may be left out. $C_{124}$ is probably unnecessary and could be left out, as the effect of $C_{123}$, $T_{R80}$ and $T_{R81}$ will be equivalent to a large capacitor at this point.

C.r.t. and blanking

The circuit associated with the cathode ray tube and the blanking amplifier are shown in Fig. 14. Before considering these, I would like to make some comments regarding the c.r.t. used in the instrument.

Of all the parts in an oscilloscope, the c.r.t. is the most critical, and the choice of tube will affect the performance that it is possible to obtain, and also the circuit techniques used. For these reasons I chose a currently available c.r.t. of modern design, instead of trying to find a surplus tube that might be satisfactory.

The c.r.t. specified has many modern features, such as a flat rectangular screen, built-in parallax-free graticule, spiral p.d.a. (post deflection acceleration) and high deflection sensitivities. It also has a second grid so that retrace blanking may be applied at earth potential. This greatly simplifies the blanking amplifier.

The penalty to be paid for this high performance, of course, is cost. The c.r.t., Mumetal shield, and base will cost about £40. For those readers who would otherwise hurriedly stop reading this article, I will now give some suggestions for possible use of surplus tubes.

---

**Fig. 13.** The square-wave amplitude calibrator.

**Fig. 14.** The tube circuit and controls with (inset) the base connexions of the Brimar D13-476GH/26 tube. Base type is B12F.
Surplus p.d.a. tubes sometimes come on the market, and it is possible that a suitable one may be found. To aid bargain hunters, I have included, in Table 1, extracts from the manufacturers application data giving the main details of the c.r.t. The most important characteristics are the X and Y plate sensitivities, and it is necessary to use a tube with, at worst, sensitivities of 2/3 of these figures. If this is not adhered to, complete redesign of the amplifiers will be necessary. E.h.t. and other voltages are not usually so critical. If a tube is found that does not possess a second grid for blanking, but is otherwise satisfactory, a circuit similar to Fig. 15 may be used.

An additional winding on the e.h.t. transformer is used to provide a floating supply of about 1.2kV. This is used to add a sufficient negative voltage to the output of the blanking amplifier to come within the required range of brilliance control voltages on the grid. The blanking amplifier needs an additional inverting stage to maintain correct polarities. Apart from these ideas I can provide no other information on other c.r.t.s, and any constructor who tries some other tube must make his own decisions.

To return to Fig. 14, D91 provides a negative voltage of 80V w.r.t. the cathode for the brilliance control. D92 is a safety diode to ensure that the grid cannot go positive with respect to the cathode. L4 provides an axial magnetic field for alignment of the horizontal trace with the internal graticule. R275 adjusts the current in this coil, and the connections are reversed if the correction is in the wrong direction.

Tr100 is the blanking amplifier and is an unsaturated switch, with D90 preventing saturation. The beam is unblanked when Tr100 is on and grid 2 on the c.r.t. is at the same potential as anode 1, i.e. 0 volts. About 60 volts is necessary to fully blank the trace.

Calibrator
The amplitude and time calibrator is shown in Fig. 13. It is a conventional multivibrator which switches a constant current to and from a ladder attenuator. This provides a square wave of good rise and fall times with an amplitude that is independent of supply voltage variations. R241 and R240 enable the frequency to be set to 1kHz, and the mark-space ratio to unity. D73 provides temperature compensation for Tr92.

Power supply
The mains power supply (Fig. 16) has been kept as simple as possible, consistent with a reasonable degree of stabilization. The main points are the large
Reference circuits — Circards 23

Set 23 of Circards, covering reference circuits, is now available. (Because of space limitations the introductory article has been omitted from this issue.)

Titles in the set are

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“Circuit designs — 1, Collected Circards” brings together the first ten sets of Circards, introductory articles to each of the subjects, and ten pages of additional circuits. The hardback A4 book contains 168 pages, in which 120 cards are rearrayed and so that each is laid out on one page. A brief introduction precedes the articles, which were previously published in Wireless World, and each of the ten subjects is followed by an up-dating page. Corrections have been incorporated where appropriate.

“Circuit designs” is obtainable through leading bookstalls at £10 per copy. In case of difficulty order direct by sending remittance for £10.40 (includes postage and packing) to the address given later, making cheques payable to IPC Business Press Ltd. Advertisement appears on page 27.

Circards are a new method of collating and presenting data about circuits in a compact and easily retrievable way. The sets of 203 x 127 mm (8 x 5 in) double-sided cards are designed for easy filing in standard boxes and for easy access at the desk or at the bench, where transparent plastic wallets keep the cards in good condition.

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15 pulse modulators
16 current-differencing amplifiers — signal processing
17 c.d.s — signal generation
18 c.d.s — measurement and detection
19 monostable circuits
20 transistor pairs
21 voltage to frequency converters
22 amplitude modulators.
23 reference circuits

Hermetic plastics i.c.s.

A process introduced by RCA is claimed to have overcome the problems of obtaining a hermetic seal in a plastic-packaged integrated circuit at no increase in cost. A high degree of hermeticity has previously been obtainable only with relatively expensive ceramic or glass packages and seals, the package bearing the brunt of atmospheric attack. In the new “trimetal” process, the chip itself is sealed.

At the equivalent of the final oxide step in ordinary i.c.s, a silicon nitride layer seals the junctions, a mask being used to gain access to contacts. Platinum is then sputtered over the wafer and sintered locally to provide platinum silicide. Layers of titanium and platinum are now laid down, the platinum layer being etched to provide interconexions, which are then gold-plated. (Titanium is used to provide an easy bond between platinum and silicon nitride.) All this avoids the use of aluminium interconnecting runs, which are prone to attack by moisture. The new process is applied to six RCA linear i.c.s, including the CA741G and 747G which take the name of “Gold Chip” devices.
Electronic circuit calculations simplified

2 — Resistive circuits (continued)

by S. W. Amos, B.Sc., M.I.E.E.

Two resistors in parallel. Every experimenter knows that the effective value of a resistor is reduced by connecting another resistor in parallel with it. But by how much is it reduced? To find out we can, of course, use the following well-known expression for the resistance of two resistors \( R_1 \) and \( R_2 \) connected in parallel:

\[
\text{effective resistance } R_{\text{eff}} = \frac{R_1 R_2}{R_1 + R_2}
\]

With this equation we can determine the product of individual resistances.

As an example a 33-kilohm resistor in parallel with a 47-kilohm resistor has an effective value given by:

\[
R_{\text{eff}} = \frac{33 \times 47}{33 + 47} = 19.4 \text{ kilohms}
\]

In practice, however, problems concerning resistors in parallel are usually presented differently. Often we need to know what value of resistor must be connected in parallel with a given resistor to obtain a desired lower value of resistance. It is tedious and unnecessary to repeat the above calculation several times in order to obtain the answer. Instead, the above expression can be rearranged as shown below to give the required information directly:

\[
\text{resistor to be added } R_2 = \frac{R_1 R_{\text{eff}}}{R_1 - R_{\text{eff}}}
\]

As an example, if it is desired to reduce a resistor of 27 kilohms effectively to 22 kilohms, the resistance which must be connected in parallel is given by:

\[
R_2 = \frac{27 \times 22}{27 - 22} = 120 \text{ kilohms}
\]

The above expressions show that the value of a resistor is effectively halved by connecting an equal-value resistor in parallel with it. If the added resistor has twice the value of the original resistor, the net resistance is two thirds that of the original resistor. If the added resistor has three times the resistance of the original, the net resistance is three quarters of the original. In general, if a resistor with \( n \) times the resistance of the original resistor is connected across the original, the net resistance is \( n/(n + 1) \) that of the original resistor.

Another useful rule which can be deduced from the above expression is that in order to reduce the effective value of a resistor by 10%, the parallel resistor must have a value 9 times that of the original. For a 5% reduction, the parallel resistance must have 11 times the resistance of the original and for a 1% reduction the parallel resistance must be 99 times the original resistance. These added resistances are respectively approximately 10 times, 20 times and 100 times the original resistance — in each case 100 times the reciprocal of the percentage reduction required. Thus to effect a 2.5% reduction, the added resistance should be 100/2.5, i.e. 40 times the original resistance. In general to reduce the effective value of \( p \% \) the added resistance should have a value of \( (100 - p)/p \) times the original and if \( p \) is small compared with 100, it can be neglected in the numerator of the fraction so that the added resistance is approximately \( 100/p \) as deduced from the numerical examples earlier.

Current divider. The current \( I \) which flows externally to a parallel combination of resistors divides between them in the inverse ratio of their resistances. Thus in Fig. 16 the current \( I_1 \) flowing in \( R_1 \) is given by:

\[
I_1 = \frac{R_2}{R_1 + R_2} \cdot I
\]

and \( I_2 \), the current in \( R_2 \), is given by:

\[
I_2 = \frac{R_1}{R_1 + R_2} \cdot I
\]

It is useful to regard a parallel resistor combination as a current divider because such a combination is often used as the basic negative feedback circuit in a current amplifier, and this approach enables the resistor values needed to give a wanted value of current gain to be readily calculated. Alternatively it enables the current gain of an amplifier to be deduced from inspection of the resistor values used in the feedback circuit.

For example, in the simplified circuit diagram of Fig. 17, \( R_1 \) and \( R_2 \) are effectively in parallel, the current path through \( R_1 \) being completed by the input resistance of \( T_r \) which is normally small compared with \( R_2 \). The emitter current of \( T_r \) splits at the junction of \( R_1 \) and \( R_2 \) and the fraction of it which is returned to \( T_r \) base as a negative feedback signal is, as shown by the above expressions, given by \( R_1/(R_1 + R_2) \). Normally \( R_1 \) is large compared with \( R_2 \) and the fraction is thus approximately \( R_1/R_2 \). The current gain of the amplifier is equal to the reciprocal of this fraction i.e. \( R_2/R_1 \). The values of \( R_1 \) and \( R_2 \) should thus be chosen to give the desired value of current gain. Now \( R_2 \) is one of the components used for biasing \( T_r \) and this consideration imposes limitations on its value: a likely value...
for \( R_1 \) is 100 ohms. \( R_2 \), on the other hand, can be made almost any value and, to give a current gain of 100, should have a value of \( 100 \times 100 \) ohms i.e. 10 kilohms. Only the essential signal-frequency components are shown in Fig. 17: in a practical circuit additional components may be necessary, e.g. for stabilising the mean emitter currents of the transistors.

**Two resistors in series.** The effective value of two resistors \( R_1 \) and \( R_2 \) connected in series is the arithmetical sum of the two thus

\[
R_{\text{eff}} = R_1 + R_2
\]

Thus is a small resistor is connected in series with a large one, the effective resistance is slightly greater than the larger. (If the resistors are connected in parallel, of course, the effective resistance is slightly less than the smaller of the two.)

Two resistors connected in series are often regarded as a potential divider because the voltage \( V_{\text{out}} \) across \( R_2 \) (Fig. 18) is a certain fraction of that (\( V_{\text{in}} \)) applied to the combination: the fraction is determined by the resistor values according to the expression

\[
V_{\text{out}} = \frac{R_2}{R_1 + R_2} V_{\text{in}}
\]

which is similar to the expression for the current in \( R_1 \) in a parallel resistor combination.

Potential dividers are often used for biasing transistors and Fig. 19 shows a typical circuit using an f.e.t. The potential divider \( R_1 R_2 \) impresses a particular value of voltage on the gate of the f.e.t. and the external source resistance \( R_e \), then determines the drain current of the transistor. This bias circuit is preferable to that in Fig. 3 (June) because it gives better d.c. stability i.e. it defines the mean drain current more accurately in spite of manufacturing spreads in transistor parameters and variations in parameters with temperature.

As mentioned earlier one of the features of f.e.t.s is that their input resistance is very high. Thus there is virtually no gate current and the resistors \( R_1 \) and \( R_2 \) are required solely to provide a particular voltage for the gate. The only current in \( R_1 \) and \( R_2 \) is therefore that which flows through them from the supply (the bleed current) and this can be made any desired value. Normally the bleed current is made very small because this enables the input resistance of the circuit to be kept high. To illustrate this let us assume that there is a 15-V supply and that the transistor is required to take 2mA mean drain (and therefore source) current. For such a current the transistor will be assumed to require a gate-source bias voltage of -1.5V. The gate voltage can be given any desired value up to 15V: 3V is a convenient value. The mean source voltage must then be 4.5V to give the required gate-source bias. Thus the voltage across \( R_1 \) is 4.5 and the current required in it is 2mA. Ohm’s law gives the value of \( R_1 \) as 2.2 kilohms.

The potential divider must give a gate voltage of 3 and the supply voltage is 15. Let us assume a very low value of bleed current, say 10µA. This is then the current in \( R_2 \) and the voltage across \( R_2 \) is 3. Thus, from Ohm’s law, the value of \( R_2 \) is 300 kilohms. The current in \( R_1 \) is also 10µA and the voltage across it is 12 so that its value, also from Ohm’s law, is 1.2 megohms. \( R_1 \) is four times \( R_2 \) and provided this ratio is maintained the gate voltage always has the same value of 3: the particular values chosen for \( R_1 \) and \( R_2 \) determine the bleed current.

In practice we are often concerned about the input resistance of the circuit. The input resistance of the f.e.t. itself is practically infinite but the input resistance of the circuit depends on \( R_1 \) and \( R_2 \), which are effectively in parallel at signal frequencies. For the values of \( R_1 \) and \( R_2 \) calculated above the effective input resistance of the circuit is given by:

\[
\text{input resistance} = \frac{\text{product of } R_1 \text{ and } R_2}{\text{sum of } R_1 \text{ and } R_2}
\]

\[= \frac{1200 \times 300k}{1200k + 300k} = 240 \text{ kilohms}
\]

With f.e.t.s it is possible to achieve much higher input resistances than this. For example, suppose we require an input resistance of 5 megohms: such a value might be required to terminate a capacitor or piezo-electric microphone or pickup cartridge. A useful way of calculating the values of \( R_1 \) and \( R_2 \) to give this value of input resistance as well as the required value of gate voltage is as follows. Let the ratio of the gate voltage to supply voltage be \( a \). This \( a \) is also therefore the step-down voltage ratio of the potential divider. Then the resistor values to use in the potential divider are given by:

\[
R_1 = \frac{\text{input resistance}}{a}
\]

and

\[
R_2 = \frac{\text{input resistance}}{1-a}
\]

For the example in question \( a = 3/15 = 0.2 \)

\[
R_1 = \frac{5}{0.2} = 25 \text{ megohms}
\]

\[
R_2 = \frac{5}{1-0.2} = 6.25 \text{ megohms}
\]

\( R_1 \) is four times \( R_2 \) which ensures the required gate voltage. A calculation of the resistance \( R_1 \) and \( R_2 \) in parallel shows that this is 5 megohms as required.

The circuit of Fig.19 is also used to bias bipolar transistors but the design usually proceeds along entirely different lines from those described for f.e.t.s. This is primarily because the bipolar transistor has a significant input (base) current: as a result the input resistance is low and, because this shunts both resistors of the potential divider, there is no point in using high-resistance components for \( R_1 \) and \( R_2 \). Indeed there is a good reason for using low-value resistors, namely that d.c. stability of the circuit is dependent on the value of these resistors, increasing as the resistor value is decreased. The resistors should therefore be made as low as possible subject to keeping the bleed current acceptable: in battery-operated equipment the potential divider should preferably not take as much current as the transistor itself.

A good starting point for the design is thus to decide on a value for the bleed current. Suppose the transistor is to take 1mA mean collector current. Then it is reasonable to let the potential divider take 0.1mA. Let the required base voltage be 3V as before. Then a simple application of Ohm’s law tells us that \( R_1 = 3/(0.1 \times 10^{-3}) = 30 \) kilohms. Now \( R_1 \) carries the base current of the transistor in addition to the bleed current of 0.1mA. We can take the base current as \( 1/\beta \) of the collector current. If \( \beta = 100 \) then the base current is 0.01mA and the total current in \( R_1 \) is 0.11mA. If the supply voltage is 12 then there are 9V across \( R_1 \) and the value of \( R_2 \), is, from Ohm’s law, \( 9/(0.11 \times 10^{-3}) \) i.e. 82 kilohms.

Finally we need to calculate the emitter resistor value. If the transistor is a germanium type the emitter potential is very nearly equal to the base potential and there is thus 3V across \( R_2 \). Since the current in \( R_2 \) is 1mA, the value of \( R_2 \) is given by \( 3/(1 \times 10^{-3}) = 3 \) kilohms. If, however, the transistor is a silicon type there is an offset voltage of approximately 0.7V between base and
Fig. 20. Simplified diagram of a two-stage voltage amplifier with emitter follower output in which the gain is determined by the negative feedback circuit $R_1/R_2$.

Emitter potentials and the voltage across the emitter resistor is only 2.3V, giving $R_2$ as 2.3 kilohms.

If the base current had been ignored $R_1$ would have been calculated as 90 kilohms instead of 81 kilohms. The difference is not great and to obtain approximate estimates of resistor values it is often permissible to neglect the base current.

A potential divider is often used as the basic element in the negative feedback circuit of a voltage amplifier and the values of the two resistors enable the gain of the amplifier to be set at a particular wanted value. Alternatively, the gain of a voltage amplifier can be deduced from the values of the resistors in the negative feedback circuit. For example, in the simplified circuit diagram of Fig. 20 $T_{R1}$ and $T_{R3}$ are common-emitter stages and $T_{R1}$ is an emitter-follower stage which provides the amplifier with a low output resistance. $R_1$ and $R_3$ constitute the potential divider which defines the voltage gain of the amplifier. The step-down voltage ratio of the potential divider is $R_2/(R_1 + R_2)$ but $R_1$ is usually large compared with $R_3$ and the step-down ratio is therefore approximately $R_2/R_1$.

The voltage gain of the amplifier is equal to the reciprocal of this i.e. $R_1/R_2$. Thus if voltage gain of 200 is required $R_1$ must equal 200$R_2$. $R_3$ is used to provide bias for $T_{R1}$ and its value is to a large extent determined by this consideration: a likely value for $R_3$ is 100 ohms. $R_2$, on the other hand, can be given almost any value and, to give the desired value of voltage gain, should be $200 \times 100$ i.e. 20 kilohms.

Only the essential signal-frequency components are shown in this circuit diagram: in a practical circuit additional components are necessary, e.g. for stabilising the mean collector currents of the transistors.

### Announcements

A one day conference in Cybernetics is being organized at Chelsea College, University of London, by the Cybernetics Society on September 1st, 1975. Topics will include artificial intelligence, pattern recognition, cybernetic medicine, systems theory and other topics relating to cybernetics. Further details about the conference or offers of papers can be obtained from the conference organizers: F. Inman c/o The Cybernetics Society, Chelsea College, University of London, Pulteney Place, London S.W.6 or Dr. C. M. Eisost, Cybernetics Dept., Brunel University, Uxbridge, Middlesex. The Society also holds monthly meetings in London with speakers from various fields in cybernetics and other related topics.

Anyone interested in attending the meetings or becoming a member of the Society should contact Mr. Kevin Clifton at the Cybernetics Society address or by telephone 01-736 1244, ext 229.

Norse Audio Systems Ltd recently launched the Radionette range of audio and television equipment in the UK. The range, which is manufactured in Norway, includes colour television receivers, music centres, tuner-amplifiers, record decks, speakers, and transistor radios. Radionette (a subsidiary of the Tandberg organization) have said that the product range will be backed-up by a first class after sales service.

### Vision network switcher

Logic-controlled network-switching equipment, recently installed at Yorkshire Television's Leeds studios, provides very simple routing of signals from 34 inputs (studios, telecine, v.t.r., etc) to three main external outputs and several internal monitoring and recording stations.

The photograph shows the control panel of the switcher, which contains the relevant buttons for two of the external outputs, the other being on another panel. The top two banks (of 68 buttons) each consist of 34 buttons to select signal source, coupled with "on air" lamps. When switching codes, set up in memory by the selector buttons, are activated by output "take" buttons, the "on air" lamps illuminate. Outports are to the Emley Moor YTV transmitter, the national network and the YTV Belmont transmitter, the selectors for this being on a separate panel. The bottom bank of buttons select inputs for previewing or prelistening and the left-hand set controls feeds to internal monitoring stations, etc.

In essence, the control panel sets up in memory the switching configuration required, resulting logic states being used to control single-i.e. cross-point switches (reeds, in the case of audio cross-points).

The equipment was supplied by Crow of Reading, the cross-points originating with Sandar Electronics — a Norwegian company.
Noise — confusion in more ways than one

4—Noise figure and the design of front-ends

by K. L. Smith
University of Kent at Canterbury

The quest for low equipment noise temperature has formed a large part of research and development effort in recent times. The maser, parametric amplifier and other aspects of low-noise technique are covered in this concluding article. Noise figure is still commonly used to characterise performance and this idea, together with the earlier discussion of noise temperature, is considered to show that basically they are saying the same thing.

Getting noise levels down at the front end of modern equipment has been a success story. In this section I will present an outline of some of the strides that have been made, but will not attempt a detailed description of actual front-end hardware, or specific devices and techniques. If you have any special interest requiring a little more detail, I have mentioned a few references for you to follow up.

The basic mechanism of noise generation in active devices is the shot effect, found, of course, in semiconductors as well as in thermionic emission devices. A serious limitation in microwave receivers has been the crystal mixer. Until the advent of low-noise r.f. stages such as the ruby maser and parametric amplifier, the crystal mixer was the front-end component. The crystal has a direct current flowing when operating and therefore shot noise is produced. This makes it appear “hotter” than an equivalent resistor under the same conditions. It also has a conversion loss. If you glance at equation 9 (Part 3), with $L$ now standing for the mixer conversion loss and $T_J$ somewhat above room temperature and related to the crystal noise temperature, then you will see that crystal mixers do not enhance the requirement for low effective input noise temperatures or microwave receivers.

One awkward point arises because of a traditional definition. The crystal noise temperature, $T_c$, is defined as the effective temperature at the output of the mixer stage when the input is terminated with a matched source resistor at $T_o$ (290K). So $T_c$ includes the source contribution at $T_o$. From this, taking care to account for the source contribution, we can write down an analogous equation to (9) for the $T_c$ of the superhet with a crystal mixer front end and an i.f. amplifier whose effective input noise temperature is $T_{i,f}$.

$$T_c = L + T_o + LT_{i,f}$$ (11)

The first term is the crystal noise temperature referred to the front end or input terminals; the second term is the subtraction of the standard source temperature, assumed in the definition of $T_c$, and the third term is the i.f. amplifier input temperature referred forward to the front-end terminals. For good noise performance, a low $T_c$ and small $L$ is required. A manufacturer's catalogue shows the following for the 1N23C point-contact X-band mixer diode: $T_c>290=2L=6DB$, which is a loss of four times.

Putting these values into (11) gives

$$T_c = 4 \times 2 \times 290 - 290 + 4 \times 116 \cong 2500 K.$$ 

I have assumed a low-noise i.f. amplifier, a temperature of $T_{i,f}=116 K$. The noise temperature of 2500K is not a very good performance. From the same catalogue, a modern Schottky-barrier mixer diode type P1906F would give an effective input noise temperature of 740K if used in the same receiver.

Other hazards exist which I have ignored in the above discussion. An important one is the noise generated in the local oscillator. The noise sidebands from the oscillator mix to produce an appreciable output at the i.f. frequency. A balanced mixer should be used to reduce noise from this source. Another dodge is to lock the local oscillator to a low-noise stable frequency generator, such as a crystal oscillator and multiplier chain (which, incidentally, reduces the drift, frequency jumping and the f.m. noise of the local oscillator). The noise performance of the i.f. amplifier is critical with direct conversion receivers. The presence of the loss factor $L$ makes this so, as you can see from (11).

The first stages of the amplifier require transistors specially selected for their good noise performance and the optimum matching conditions between mixer and i.f. input circuit is important. It is possible to obtain optimum performance empirically by switching on and off a gas-tube noise source coupled into the front end, attenuated if necessary, while adjusting the mixer, local oscillator, and i.f. couplings for minimum $T_c$, as monitored by observing the changes on a power meter at the output of the system. The calculation of the optimum conditions for any given case, including the effects of parasitic reactance, etc. is extremely difficult. If you wish to follow this up, E. G. Nielsen14 wrote an interesting article on the topic.

You may recall the publicity, about a decade ago, in connection with the Goonhilly station system. The “Signifi-
...cant" results that the new maser amplifier was going to make possible were indeed achieved, after the usual teething troubles. We had visions of large Dewar flasks surrounded by liquid-nitrogen-produced vapour clouds and all the other complexities of the cryogenics. The cost and complexity of the maser hardware and cryogenics, and the relatively narrow bandwidth obtainable has meant a decline in their use, and the much more convenient parametric amplifier has taken over. In spite of this, the maser still offers the ultimate in low-noise performance, because not only are the "working bits-n-pieces" cooled to about 4K with liquid helium, you will sometimes see negative temperatures mentioned! Of course, the physical temperature is never below absolute zero, but in some ways the maser acts as if it were.

The action of the solid-state maser depends on materials with paramagnetic ions in a crystal lattice, such as the chromium\(^{3+}\) ion in ruby. The spin and orbital motion in these atomic particles are quantized, which means that the only energies allowed them by the laws of quantum physics are very definite values. Any change in energy of particles means a jump from one energy level to another with the absorption or emission of radiation of characteristic frequency. This is pictured in a diagram such as that in Fig. 16; (a) shows some of the chromium ion levels in ruby. The lowest level is the only one of interest for maser work.

In this ion, there are three available electrons which contribute to the paramagnetic splitting. The three electrons give four possible energy levels and the field of force in the crystal lattice splits the levels into two pairs 11.46- GHz apart (the energy difference is measured in "frequency" - this being the frequency of the radiation involved in any jump between the levels), see Fig. 16 (b). A long time ago a fellow named Zeeman in 1896 found that applying an external magnetic field to atoms or molecules with levels which have the same energy (the jargon for this is that they are "degenerate", but this does not mean that they suffer from anything nasty) separates them, or as we say lifts the degeneracy by an amount depending on how strong the magnetic field is.

Sure enough, this happens with our chromium ions in ruby, see Fig. 16 (c). The low energy levels of the billions of particles in the crystal lattice are occupied according to the quantum laws, and the absolute temperature \(T\). The lowest levels are crowded while higher levels are only sparsely filled. The whole point of maser action is to invert this population distribution - by pumping, so that when dropping back to a low energy level, the electrons involved give energy to the passing signal wave, boosting its amplitude. The waves passing through the crystal (we are discussing a travelling-wave maser) need time to interact with the excited atoms, so a careful design is made of a slow wave structure for this purpose.

The correct populations can only be produced at very low temperatures - hence the liquid helium involved. The magnetic field to split the energy levels to just the right value for the frequency of the signal and pump is often applied by the use of a "persistent current superconducting magnet" - a case of exploiting the very low temperature available and the resulting superconductivity in some materials. Fig. 17 (a) illustrates this and (b) shows diagrammatically the parts of the travelling-wave maser.

We can sum up all of this by noting that the signal wave in the maser extracts energy in phase from the excited ions in the crystal. The presence of the signal stimulates the emission - hence the name of the device, Microwave Amplification by Stimulated Emission of Radiation, as you probably already know. The extremely low effective noise temperature of this amplifying mechanism arises from the requirement to use liquid helium temperature, together with a further reduction of this already low temperature according to \(T_{\text{in(masr)}} = T_{\text{amb}} / I\), where \(T_{\text{amb}}\approx 4.2K\) (liquid helium) and \(I\) is the inversion ratio of the maser, usually about three and is the number of times the upper energy level is more densely populated than the lower, as a result of the pumping action.

Naturally, the signal has to be got into the amplifying part of the maser, and out again. Equation 9 shows that the effect of any attenuation in the feeder at the input, will very convincingly degrade the noise performance.

\[
T_s = (L-1)T_m + \frac{LT_{\text{amb}}}{I}
\] (12)

If \(T_m\) is a mean temperature of around 100K and \(L\) is a loss of only 0.5dB (1.122 times), then \(T_s = (11.2 + 56)K\). That is, the loss contributes about 11K and the maser itself about 1.5K. Herein lies a big difficulty with such amplifiers: how to get the signal down into the crystal in the Dewar, through the couplings and hardware of the front-end feeder system, without introducing prohibitive losses. Taking the signal out is all right, it has been amplified by the maser gain,

---

Fig. 17. Inconvenience of maser amplifiers is mainly because of the complex cryogenic system required. Also, getting the signal in and out and the pump power in to the crystal is troublesome. Show here is a typical example of a maser with rutile for the slow-wave structure and interacting element (a). Magnet could be a self-sustaining superconducting one. Velocity of the wave in rutile is very much less than the velocity in free space, because of the high dielectric constant. One method of feeding in the pump power that has been proposed is to use a side radiating horn, as shown at (b). (A. Fletcher).
perhaps 45dB or so. A good article on masers was presented in Philips Technical Review in 1965 if you can get hold of a copy. The practical description of the Goonhilly Down maser is discussed fully.

**Parametric amplifier**

The source of gain in the parametric amplifier is an entirely different physical mechanism to that in the maser system, although a pump oscillator is again employed. Most of the literature proudly states that Lord Rayleigh, and even Michael Faraday, spoke of the paramp principle way back in the last century. This is so, but the electronic realization of the principle for low-noise amplification is recent (the 1950s). Basically a reactive parameter, such as a capacitance, is varied and this feeds energy into a signal wave. Being a virtually noiseless parameter (reactance), the usual Johnson and current noise contributions are reduced. There is some loss in the active device, which is usually a varactor diode, and in the feeder hardware. Cooling will therefore lower the effective noise temperature. Noise contributions also arrive in the pump frequency oscillations and can degrade the performance.

Three frequencies are usually involved in a paramp; the signal frequency \( f_s \), the pump frequency \( f_p \), and the idler frequency \( f_i \). The idler frequency is the mixing product of \( f_s \) and \( f_p \), i.e. \( f_i = f_p - f_s \). Much of the design work in parametric amplifier projects is involved with the correct design of the resonant structures for \( f_s \) and \( f_i \), and in ensuring the isolation between them. A circulator is required at the signal front end, because the varactor diode is a two-terminal device. (Tunnel diode amplifiers also have this drawback.) You are correct if you immediately assumed that isolators, circulators, switches or any other such lossy hardware in the input feeder are bad for low-noise performance, equation 9 or 12 operates again.

Fig. 18 gives some idea of the electrical arrangement of a pumped paramp of the “two-tank” variety. The input signal arriving at port 1 of the circulator is diverted out of port 2 into the signal-tuned circuit \( C_L \) in the paramp. The pump is, varying the capacitance of the varactor and energy is fed into the signal oscillations by this action. The amplified signal is passed back to port 2 of the circulator. (You could look upon all this as a signal passing into the paramp on the input transmission line and there reflected with a reflection coefficient greater than unity back down the line.) The circulator passes the amplified signal out of port 3 to the load. The circuit \( C_L \) resonates at the idler frequency and filters \( F_s \) and \( F_i \) reject the idler and signal frequencies respectively so keeping the various signals and oscillatory powers in their places.

There are various ways in which the realization of the scheme shown in Fig. 18 can be carried out. Fig. 19 shows just one possibility. The pump is much higher in frequency than the signal and is usually supplied via a waveguide, although microstrip techniques are increasingly being employed.

A theoretical analysis of Fig. 18 enables the gain, noise temperature and bandwidth of the amplifier to be derived. If you are interested in some of the theoretical argument, you will find a very good discussion in reference 16. One factor of great importance is \( \gamma \), which is defined as

\[
\gamma = \frac{C_{\text{max}} - C_{\text{min}}}{2(C_{\text{max}} + C_{\text{min}})}
\]

This is a kind of goodness factor for the varactor diode and indicates the amount of capacitance variation obtainable by pumping. Another diode parameter is the cut-off frequency, \( f_c = 1/2\pi CR_m \), which must be way above the frequencies involved in the amplifier. Noise is contributed directly by the diode loss resistance \( R_d \) and also \( R_L \) in the idler circuit. It is in cooling these that the improved performance at low frequencies is obtained. Quoting now the noise temperature expression (see reference 16)

\[
T_n = T + \frac{R_f}{R_p} + A_f(1 + \frac{R_d}{R_s})
\]

where \( R_f/R_p \) is often called the over-coupling ratio and \( T \) is the physical temperature. \( A \) is dependent on the gain, but is normally close to unity, especially at high gain. This means that if we make \( R_f \rightarrow R_p \), then the effective noise temperature of the paramp is

\[
T_{\text{effective}} = \frac{T}{f_s/f_i}
\]

(13)

Under these conditions, we can see that if the pump frequency is high, so that \( f_c < f_i \), then the effective input noise temperature of the parametric amplifier can be made less than the physical temperature (i.e. by the fraction \( f_i/f_c \)). Lest you think “Ah! let’s push up the idler frequency to millions of gigahertz – and get noise temperatures around absolute zero”, life is not so kind. Remember \( f_c \) for the diode, and other losses limit the possibility of an unlimited rise in the pump and therefore idler frequency. The noise temperature given by (13) is oversimplified and as the gain drops when \( f_c \) goes up the noise temperature will start to increase again.

This would indicate that there is a minimum noise temperature for an optimized paramp with a given diode, packaging, and so on. It can be shown that

\[
T_{\text{effective}} = \frac{2f_i}{\gamma f_c}
\]

As would be expected, a large \( \gamma \) and high \( f_c \) for the diode is the best way to a low \( T_n \) at any given signal frequency \( f_c \).

**Handling the idea of noise figures**

The noise figure, \( F \), has already been mentioned and you may recall from part 1 that H. T. Friis and D. O. Hohn were instrumental in getting the concept off the ground. All the equations to calculate \( F \) are derived from the basic definitions in a similar way to the expressions for \( T_n \). If we went through it all again, it would mean a duplication of effort. The main point now is to derive the relationships between \( F \) and \( T_n \) so that all the earlier equations can be written in terms of \( F \), if required. Friis defined \( F \) as the signal to noise ratio at the input of a network when the source resistance is close to 290K, divided by the signal-to-noise ratio at the output. Because signal-to-noise ratios at the output of amplifiers, receivers and so on, are always smaller than those at the input, then \( F \) is always larger than one.

From the above verbal definition you will see that

\[
F = \frac{S_i}{P_{\text{in}}} \quad \text{and this tides up to} \quad \frac{S_{\text{out}}}{P_{\text{out}}} \frac{P_{\text{in}}}{S_{\text{in}}}
\]

If the network has a power gain of \( G \),
then \( S_i = G_A S_i \) and \( P_{N_0} = G_A P_{N_i} + P_{N_0} \), by precisely the same argument that was made in part 2 (first paragraph). This means that

\[
F = \frac{S_i (G_A P_{N_i} + P_{N_0})}{P_{N_i} G_A S_i} = 1 + \frac{P_{N_0}}{G_A P_{N_i}} \quad (14)
\]

Immediately from the equation \( P_{N_0} = G_A k \), \( T_e B \), (part 2) and \( P_{N_i} = k(290)B \), so that from equation 14

\[
F = 1 + \frac{T_e}{290} \quad (15)
\]

This is the relationship connecting \( F \) and \( T_e \), we set out to find.

I indicated at the beginning of this series that D. O. North also defined a noise factor. His definition dispensed with signals right from the start. You may like to see that his definition is just the same as that just given in (15), although North originally suggested 300K for the standard temperature. North's definition went like this: “\( F \) is the ratio of the total noise power output from a system when its input termination is at 290K, to that part of the output which arises from the input termination only”.

In symbols, this is

\[
F = \frac{P_{N_0} \text{ (when } T_i \text{ is } 290 \text{K})}{G_A k(290)B}
\]

but \( P_{N_0} = G_A k(290 + T_e) \), see part 2, p.169, and therefore

\[
F = 1 + \frac{T_e}{290}, \text{ as before.}
\]

Some early ideas connected with this topic was discussed in Wireless World by L. A. Moxon12.

We are now in a position to write down any noise expression in terms of \( F \) by using the fact that \( T_e = 290(F - 1) \) by transposing (15). You can see this from the following examples.

Substituting for \( T_e \) in the equation on p.171 (part 2) gives

\[
F = \frac{F_e - 1}{G_1} \quad \frac{F_e - 1}{G_2}
\]

From equation 6

\[
T_e = \frac{T_{hot} - A T_{cold}}{A - 1} \quad \text{or}
\]

\[
290(F - 1) = \frac{T_{hot} - A T_{cold}}{A - 1}
\]

Substituting for \( T_e \) gives

\[
F = \frac{T_{hot} - A T_{cold}}{290 \quad A - 1} + 1 \quad \text{which is}
\]

\[
F = \left( \frac{T_{hot}}{290} - 1 \right) - A \left( \frac{T_{cold}}{290} - 1 \right)
\]

The quantities in the brackets are excess noise ratios. Very often \( T_{cold} = 290 \), so that

\[
F = \frac{T_{hot} - 1}{290} \quad A - 1
\]
If A is made 2 by choosing the value of $T_{\text{hot}}$ to make it so, then
\[ F = \frac{T_{\text{hot}}}{290} - 1 \]

The last equation in part 2 shows this to be equal to $20\log R$ for a noise diode, so that we have the interesting result $F = 20\log R$, a well-known expression.

For the next example, we can look at (8), which when substituted gives
\[ F = 1 + \frac{(L-1)T_{r}}{290} \]

Similarly from (9)
\[ 290(F-1) = (L-1)T_{r} + L(290(F_{R}-1)) \]
\[ F = 1 + \frac{(L-1)T_{r}}{290} + L(F_{R}-1). \]

Considering equation (11), we can now see why we had some argument about $T_{c}$, because the original derivations were in terms of $F$. We should expect to see a simplification when "going backwards" to $F$, to the well-known equation for crystal mixer performance. Now $T_{c} = 290(F-1)$ and $T_{f} = 290(F_{f}-1)$, also $T_{0} = 290K$ all by definition, so that equation 11 gives
\[ 290(F-1) = L(T_{x} - 290) + L(290(F_{f}-1)) \]
\[ F = L(t_{x} + F_{f}-1) \]
where $t_{x} = T_{x}/290$. This is often seen quoted in discussions about crystal mixers.

And so we could go on. Notice that there is often a "1" for the first term on the right-hand side in noise figure equations. This arises because of the contribution of the source at 290K. When the division by 290 is made to obtain the ratio that is $F$, the first term is "1". This means (by subtracting the "1" from both sides) that the expression $F-1$ occurs frequently, hence the growth of the term excess noise figure for this. Because I am making a case for a decline in the use of $F$ and an increase in thinking in terms of temperature, I will not labour the point any further.

If you would like to give more prominence to $F$, and convert all the formulae, for instance, then you can go ahead. But, in this series of articles, I have attempted to show that the mysteries of "temperature" are only imaginary and that so long as care is taken to realise that the term has been extended to mean more than physical hotness, then the idea is valuable in noise discussions.

I suggest that the noise figure is not so clear as temperature, and is tied to "290K", which is sometimes manipulated behind the scenes for the unawary, and this makes a mockery of "$F". $F$ has a habit of being what anyone wants it to be (see the satirical article The Art of Noisemanship by J. C. Green, also The Noise Figure Muddle by Seymour B. Cohn). As you might see, a lot of confusion can arise. If you have not yet embarked upon a course of advanced study in electronics, perhaps I have been able to give you an inkling of the interesting concepts concerning equipment limitations, meanings of temperature, etc., that you are likely to meet as part of your studies. If you are already doing some work in communications engineering, then a few of the points covered in these articles may show a little of what is "behind the scenes". Mastering the technique of handling $F_e$, and especially $F_{op}$ or $F_{net}$ gives you a direct route to the overall signal-to-noise ratio — which when all is said and done, is the vital parameter in modern communications systems.

**References**


**Correction to part 2 (April issue)**

An error occurred in the discussion of noise bandwidths in Appendix B, page 173. The two equations for $G_1$ at the top of the page are correct, but $L$ and $C$ have subsequently become transposed. There is a symmetry about $L$ and $C$, so the final result is correct, but the statement
\[ R = \frac{\omega C}{wL} - 1 \]
and all $L$s and Cs should be interchanged from then on. (The error is obvious if you look at the statement $B_{dB} = R/2\pi C$, which cannot be true because it is dimensionally incorrect. Replace C with L and it is then alright.)
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A custom-designed MOS LSI digital processing IC controls the auto-polarity dual-slope integration A to D converter. The circuit built around this IC uses a MOSFET op-amp input buffer with 0.1% metal-film resistors. The result is excellent accuracy and stability with a very high basic input impedance.

The instrument reads to ±1999 and has a basic accuracy on the 1 V DC range of 0.3% ± 1 digit. Four 8 mm LED displays provide excellent legibility and angle of view. Battery operation allows complete independence of mains supply.

The Sinclair DM2 has all the capability you need. Just take a look at its features and compare them with higher-priced multimeters. You'll find the DM2 is their equal in virtually everything – except price!

Features of the Sinclair DM2

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<th>Functions giving 22 ranges</th>
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<td>DC volts – 1 mV to 1000 V</td>
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<tr>
<td>AC volts – 1 mV to 500 V</td>
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<tr>
<td>DC current – 0 to 1 A</td>
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<tr>
<td>AC current – 0 to 1 A</td>
</tr>
<tr>
<td>Resistance – 1 to 20 MΩ</td>
</tr>
</tbody>
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Easy to use
Automatic polarity, bush-button selection for all ranges and modes from a single input terminal pair.

Easy to read
Big, bright 8 mm LED display gives a quick, clear reading.
3½ digit display
Display reads from 000 to 1999.

Overload indicator
 Protected
 Separate fuses for current and resistance circuits.

Accurate
Dual slope integration High stability.

Rugged construction
Tough metal casing takes the roughest treatment – try standing on it!

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### DC Volts

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<th>Accuracy</th>
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<tr>
<td>1 V</td>
<td>0-3% ± 1 Digit</td>
<td>&gt;100 MΩ</td>
<td>1 mV</td>
</tr>
<tr>
<td>10 V</td>
<td>0-5% ± 1</td>
<td>10 MΩ</td>
<td>10 mV</td>
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<tr>
<td>100 V</td>
<td>0-5% ± 1</td>
<td>10 MΩ</td>
<td>100 mV</td>
</tr>
<tr>
<td>1000 V</td>
<td>0-5% ± 1</td>
<td>10 MΩ</td>
<td>1 V</td>
</tr>
</tbody>
</table>

Maximum overload - 350 V on 1 V range
1000 V on all other ranges.

### AC Volts

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<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Input Impedance</th>
<th>Frequency Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 V</td>
<td>1-0% ± 2 Digits</td>
<td>10 MΩ/40 pF</td>
<td>20 Hz-3 KHz</td>
</tr>
<tr>
<td>10 V</td>
<td>1-0% ± 2</td>
<td>10 MΩ/40 pF</td>
<td>20 Hz-3 KHz</td>
</tr>
<tr>
<td>100 V</td>
<td>2-0% ± 2</td>
<td>10 MΩ/40 pF</td>
<td>20 Hz-3 KHz</td>
</tr>
<tr>
<td>1000 V</td>
<td>2-0% ± 2</td>
<td>10 MΩ/40 pF</td>
<td>20 Hz-1 KHz</td>
</tr>
</tbody>
</table>

Maximum overload - 300 V on 1 V range
500 V on all other ranges.

### DC Current

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Impedance</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 µA</td>
<td>2-0% ± 1 Digit</td>
<td>10 KΩ</td>
<td>100 nA</td>
</tr>
<tr>
<td>1 mA</td>
<td>0-8% ± 1</td>
<td>1 KΩ</td>
<td>1 ρA</td>
</tr>
<tr>
<td>10 mA</td>
<td>0-8% ± 1</td>
<td>100 Ω</td>
<td>10 ρA</td>
</tr>
<tr>
<td>100 mA</td>
<td>0-8% ± 1</td>
<td>10 Ω</td>
<td>100 µA</td>
</tr>
<tr>
<td>1000 mA</td>
<td>2-0% ± 1</td>
<td>1 Ω</td>
<td>1 mA</td>
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</table>

Maximum overload - 1 A (fused).

### AC Current

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Frequency Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 mA</td>
<td>1-5% ± 2 Digits</td>
<td>20 Hz-1 KHz</td>
</tr>
<tr>
<td>10 mA</td>
<td>1-5% ± 2</td>
<td>20 Hz-1 KHz</td>
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<tr>
<td>100 mA</td>
<td>1-5% ± 2</td>
<td>20 Hz-1 KHz</td>
</tr>
<tr>
<td>1000 mA</td>
<td>2-0% ± 2</td>
<td>20 Hz-1 KHz</td>
</tr>
</tbody>
</table>

Maximum overload - 1 A (fused).

### Resistance

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Measuring Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 KΩ</td>
<td>1-0% ± 1 Digit</td>
<td>1 mA</td>
</tr>
<tr>
<td>10 KΩ</td>
<td>1-0% ± 1</td>
<td>100 µA</td>
</tr>
<tr>
<td>100 KΩ</td>
<td>1-0% ± 1</td>
<td>10 µA</td>
</tr>
<tr>
<td>1000 KΩ</td>
<td>1-0% ± 1</td>
<td>1 µA</td>
</tr>
<tr>
<td>10 MΩ</td>
<td>2-0% ± 1</td>
<td>100 nA</td>
</tr>
</tbody>
</table>

Overload protection - 50 mA (fused).

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Resistance comparator
—with linear fractional off-balance indication

by D. Griffiths Ph.D.

Imperial College, London

A simple d.c. op-amp design featuring linear reading of percentage unbalance for any arbitrary value of reference resistor without recalibration in the range 100Ω to 10kΩ.

It was not very clear what the design objectives were at the beginning of this effort; they never seem to be in my experience. It appeared that we needed either (a) to display the variation of the reciprocal of the unknown resistance $R_x$ about some nominal value $R_f$ with zero output when $1/R_x = 1/R_f$ or (b) to give a strictly linear output proportional to the fractional deviation of $R_x$ up to 100% above and below $R_f$ with zero output when $R_x = R_f$. The further proposal that accurately-calibrated percentage deviations should be shown on plug-in or arbitrary value of reference resistor $R_f$ without recalibration, did not seem to simplify matters.

In the event both requirements (a) and (b) were met in a single circuit design which also gave the proposed constant calibration of the fractional unbalance indication. An accuracy of order 0.1% is achieved with full scale indications of unbalance between 1% and 100% in the range 100Ω to 10kΩ.

The design may be useful on small production runs where resistors have to be hand-trimmed within certain readily-seen percentages of different awkward values; the components used in the original bread-board can then be the master references during trimming. The original application however was with temperature sensors which could be of widely different values although all had similar temperature coefficients of resistance when expressed as a percentage change; this circuit then gave the same loop gain to a process controller when used with different sensors.

As is often the case it was difficult to get away from the first lines of attack that occur to one at the initial doodling stage; I felt that something in the ratio transformer or Wheatstone bridge line would be required. After all, item (a) involved $1/R_x$, i.e. a conductance, and invited a ratio arm transformer technique; but it was not evident how to maintain a calibrated fractional unbalance indication with arbitrary values of $R_x$ at least with a reasonably simple circuit (watch forthcoming Letters' columns). On the other hand alternative (b), which involved direct display of resistance variations, ruled out a Wheatstone bridge scheme by the wide range of linearity said to be required.

The requirement (b) can be looked at thus: suppose we keep a constant and known voltage across $R_f$, then if the current through $R_f$ also flows through $R_x$, the difference in the voltage drops across $R_f$ and $R_x$ will only depend on the fractional relation of $R_f$ and $R_x$ and not on their actual values. It then starts to look like an exercise in op-amp techniques, as indicated in Fig. 1. (Requirements (b) can also be met by interchanging $R_f$ and $R_x$.)

In the prototype the voltage across $R_f$ was set at 1 volt, a higher value bringing power dissipation problems in the test resistors. A proposed system accuracy of order 0.1% implied keeping drifts and computing errors to less than 1mV when referred to this 1-volt level. Past experience suggested that a d.c. technique with type 741 op-amps would just suffice to achieve this, especially in view of the comparatively low values of resistance to be compared. (The use of an a.c. carrier scheme would enable $V_f$ and the power in the test resistors to be set many orders of magnitude smaller but the circuitry would then be a good deal more complex.)

It just remained to convert these ideas into hardware. The functions required are: (1) to hold the voltage $V_f$ across $R_f$ constant at 1 volt; (2) to use a differential amplifier to monitor $V_{in}$ and $V_f$, and scale the unbalance appropriately.

Consider first how to hold $V_f$ constant. In an op-amp follower-with-gain circuit, drawn in the usual way in Fig. 2, the ideal action of the negative feedback is to maintain the inverting input terminal (−) at the same voltage as the non-inverting input (+). This will be so, irrespective of the value of $R_p$, provided $R_f$ is not so large as to cause the output of the amplifier to saturate at either full positive or negative excursion. Calling...
R_1 = R_x and R_2 = R_y, we can redraw Fig. 2 as in Fig. 3, where the voltage across R_x will be held at V_{ref}, for an ideal amplifier. This is the first step in a realization of Fig. 1.

The second function to be achieved is that of causing the voltage across R_x to appear referenced to the common line, so it can be compared subsequently with the voltage appearing across R_y; this is the function of a differential amplifier. Such an amplifier can have its input stage operating in the inverting mode, whereupon the common-mode rejection ratio (c.m.r.r.) of the amplifiers does not limit the achievable c.m.r.r. of the circuit. On the other hand, inverting stages usually have a much lower input resistance than follower stages, though in differential applications the latter are limited by the c.m.r.r. of the op-amps used. The choice is easily made here for R_x has a constant voltage across it which lessens the c.m.r.r. demands on the amplifier and we require a high input resistance to minimize the loading of R_x; if R_x = 10kΩ, an amplifier input resistance of 10MΩ will effectively lower this resistance by 0.1%.

Fig. 4(a) shows a high input resistance differential amplifier, while Fig. 4(b) is intended to show how the common-mode rejection occurs by emphasizing the bridge action through drawing it in a more "Wheatstone" way. If R_x = R_y, then in Fig. 4(b) both points A and B move up and down in voltage together (i.e. V_{AB} = 0), thus points C and D will experience equal voltage excursions of one half of this magnitude if E remains at ground potential. This absence of output from the op-amp is consistent with its zero input, V_{CD} = 0 as required.

In Fig. 4(a) the gain to differential input signals is -1 if all the bridge resistors are of equal value. This can perhaps be seen most easily by thinking of the differential input V_{AB} as riding on top of the common-mode signal V_{CM} and remembering that ideally V_{CM} does not affect the voltage at E. Thus one can consider V_{CM} = 0 when thinking only of the differential signal and note that this makes point D at ground potential. Resistances R_x and R_y are then evidently seen as a unity-gain see-saw amplifier with phase inversion.

In the prototype circuit the maximum voltage across R_y had to be limited to 1 volt and to reduce the effects of voltage offset drifts the voltage across R_x and R_y was multiplied by three before passing it to the remaining op-amps. The differential amplifier of Fig. 4(a) can give gain by the addition of three resistors as shown in Fig. 5.

![Fig. 4 (a) high input resistance differential amplifier, (b) diagram emphasizing bridge action of (a).](image)

If R_x = R_y, the voltage gain of the input stage is +1/(1+2R_y/R_x). At first glance one might expect the factor of two to be outside the bracket but inspection of Fig. 5 shows that

\[ V_{AB} = V_{in} + V_{CM} \]
\[ V_{CM} = V_{in} \left( \frac{R_x + R_y}{R_x} \right) - V_{in} \]
\[ = V_{in} \left( 1 + \frac{2R_y}{R_x} \right) \]

Compared with many other differential amplifiers that of Fig. 5 has two outstanding advantages. First, the gain can be adjusted by varying a single resistor, R_x. Secondly, setting of the gain is quite independent of trimming for best c.m.r.r. It should be noted that if R_y does not equal R_x exactly, then this only affects the gain and does not reduce the common mode rejection of the circuit.

The addition of a standard virtual-earth summing amplifier to the circuits of Figs 3 and 5 finishes the design, the complete circuit of which is shown in Fig. 6. A saturated 741 o-amp with ±15V supplies will have an output of 13 to 14 volts and gives only a modest overload to the meter which requires ±10V at the output of A_0 for full-scale deflection. The 2.2 μF feedback capacitor reduces needle jitter below the visible limit with a 90-mm scale length meter.

One might imagine at first sight that a 1% change in V_{ref} (and hence V_f) would cause a full-scale change in meter reading when the output stage is on the 1% f.s.d. setting. (If this were so it would place a stringent demand on the stability of V_{ref} and make the circuit rather unattractive.) This misconception can arise by visualizing a 1% change in the voltage across R_y and seeing this as giving rise to a 1% unbalance at the input of the summing amplifier. However, this overlooks the changed voltage across R_x.

It is evident that if R_x exactly equals R_y then the output meter will register zero ideally for any value of V_{ref}. Suppose next that R_x = (R_y + ½%,) with V_{ref} = 1V and the output meter indicates +½% (i.e. half scale on the 1% range). If now V_{ref} increases by 1% say, then it is the extra voltage drop across the extra ½% of R_x (compared with R_y) that is the unbalanced input to the summing amplifier and will come through to the output; the new output reading would be \((0.5 \times 101/100\%) = 0.505\%\). Thus it is the span of the final meter deflection which is affected in direct proportion to possible changes in V_{ref}.

Fig. 6 shows that V_{ref} is derived here from the integrated power regulator output; the prototype used MVIR15 devices from RS Components. These resistors of devices have impressive output stability and modest price. Under constant load and in normal laboratory conditions their voltage drift is not

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visible on a 3½-digit multimeter on its 20-V range after a few minutes from switch-on, i.e. the output is steady to better than 10mV. In view of the insensitivity of this comparator circuit to changes in reference voltage, these regulators are more than adequate for supplying the reference voltage here.

Next we describe the procedure to adjust currents too. Earthing the 0.1mV wire through a large resistor may give significant and rather than adequate voltage drop within 10mV.

The zero stabilization is obtained by closing the 10kΩ bridge until the zero voltage is found at the output, and the point marked zero is made to read -100% (marked "to check e.m.r.t."). (5) Select R₁ and R₂ to be exactly the same value (around 2kΩ), and adjust one of the 3kΩ summing resistors to A₄ until the output meter reads zero on the 1% f.s.d. range. (6) See three paragraphs below.

If two resistors of known equality are not available for use in step (5) the comparator can still be used to set up two resistors to the necessary degree of equality. One of the nominally equal pair is made trimmable and adjustment carried out until interchanging these two resistors in the positions R₁ and R₂ causes no change in the output meter reading; it is of no importance if the reading is non-zero at this point since step (5) has not yet been completed.

The prototype on the 1% range had a zero stability of about 0.005% during a normal day and a drift from cold switch-on of about 0.04%.

The linearity of output was checked on the 100% and 10% ranges with a 3½-digit d.v.m. on the output of A₂ and was found to be better than 0.1% (±1 digit) with R₃ of 20kΩ, 1kΩ and 5kΩ and full-scale unbalance on either side of equality. This good linearity makes scaling of the output ranges easy and the final item of the trimming procedure, step (6) is: With R₃ = 0 and the output stage on the 100% range, then for any value of R₄ (10kΩ to 10kΩ) the meter is made to read -100% by trimming the voltage divider giving the nominal 1 volt reference. Accuracy of the scale of the remaining ranges depends almost entirely on the accuracy of the f.h. resistors around A₄ since the open-loop gain of a 741 at zero frequency is typically about 10⁶.

And what, you may ask is the fudge which is ill-concealed by the 1kΩ resistor forlornly stuck between the + terminal of A₁ and the point marked V_ref in Fig. 6? Well, it is tied up with the 100-kΩ and 10-kΩ limits which were so casually mentioned earlier without explanation. The lower limit is set by A₁ running out of the enthusiasm for supplying current much greater than 10mA, at least to the 0.1% accuracy required here. The prototype was satisfactory for R₄ down to 70Ω and lower values could doubtless be accommodated by buffering the output of A₁ with a power transistor. The upper limit arises from errors due to the bias currents in the increasing source resistance presented by A₂, by R₃ and R₄ in parallel. Making the source resistance of V_ref about 2kΩ (as seen by A₂) distributes this offset equally for R₄ = 100Ω and 10kΩ: on the prototype this error amounted to ±0.07%, causing high values of R₄ to read low by this amount. Substituting a superbeta or f.e.t. input op-amps for the 741 would enable the upper resistance limit to be pushed up.

![Diagram](https://www.americanradiohistory.com/image.jpg)
At the low frequencies, where a cyclical change of current and voltage can be recorded and where the wavelengths are many orders of magnitude greater than the transverse dimensions of a conductor or component, it is possible to insert a current- or voltage-measuring instrument into the circuit to determine these quantities. The instrument itself has either a negligible or precisely known (internal resistance) effect on the circuit and one knows, say in the case of an ammeter, that it is indicating the total current flowing at that point.

Consider, however, the case of a hollow metal waveguide: to d.c. this is a single conductor and one has difficulty with the concept of a "potential difference". Furthermore, as shown earlier in this series, the microwave current flows in circulating loops confined closely to the metal surfaces, and even were it possible to measure this current directly one could not say that this was the total current associated with an overall potential. It thus becomes conceptually inconvenient to think in terms of current and voltage as well as impractical to measure, and so these more fundamental quantities are lumped together and it is the total, time-average power which is measured.

Compared with the accuracy to which d.c. or low-frequency power can be measured, the absolute determination of microwave power is rather poor. Typical day-to-day accuracies using standard commercial power meters lie between 5% and 10% and even national standards are only in the region of 0.5% accurate. The fact that high-quality tracking and communication systems have evolved in spite of this limitation demonstrates that it is only on comparatively rare occasions that a highly accurate knowledge of power is necessary. Even then, it is usually required during the development of other instruments. In the majority of cases the important factor is the power difference between one space or time and that in another space or time, i.e. relative power. Using standard laboratory test gear relative microwave can be measured to within a small fraction of 1%.

As yet there is no practical, absolute method of microwave power measurement; that is, one which will indicate microwave power directly without calibration or transfer from some more primary effect. Although varied principles and techniques have been proposed and demonstrated over the years, the most practical way of determining microwave power has been by using some element to absorb the power and then observing the resulting heat dissipation. Devices used can conveniently be placed in one of two categories:

- in which the dissipated power produces a change in the electrical resistance of the absorbing element (called a bolometer), or a voltage difference (called a thermocouple);
- in which the rise in temperature of the dissipating element is measured and calibrated to read power (called a calorimeter).

These two groups are by no means exhaustive and there exist both variations and completely different methods of indicating microwave power. However, this series is essentially concerned with the real world of microwave electronics and, in this context, the above categories are the only ones of practical use.

**Bolometric devices**

Bolometers are temperature-sensitive resistances taking the form of either a thin resistive wire or film, called barreter, or small-bead thermistors. Calibration of these elements is carried out by low-frequency substitution; that is, the change in resistance with microwave power is noted and identified with the d.c. power necessary to produce the same change. Here one starts off with the chain of accumulating errors in the microwave power-measuring system by asking: does the
d.c. power represent the total incident microwave power? The answer is that it need not do so and several requirements have to be met before the answer can be changed to "yes" with confidence.

These requirements are that the cross-sectional dimensions of the wire or thermistor should be similar to the skin depth at the operating frequency, so that the d.c. and a.c. densities are similar; and that the physical length of the device should be as small as possible to minimize the very significant inductive reactance.

As a consequence, the barreter is usually constructed from a length of silver-plated platinum wire having a section perhaps 1mm long of the silver etched away, exposing the resistive platinum wire to the microwave field. This wire is very thin, typically 0.002mm in diameter, and is mounted inside a sealed cartridge or on a dielectric support of suitable dimensions for mounting in a waveguide or coaxial monitor. Barreters have a positive temperature coefficient of resistance, and a resistance/power sensitivity shown in Fig. 1(a). They have a very short thermal time constant of several hundred microseconds, useful for fluctuating signals, but this can produce errors when measuring the average power of a pulsed waveform because of the tendency to respond to the signal peaks.

Most popular of the general run of power-sensing elements is the bead thermistor. Composed of semiconducting, sintered metallic oxides, the bead is about 1/4mm diameter with two very fine wire contacts and has a negative coefficient of resistance. Its resistance/power sensitivity, shown in Fig. 1(b), is much greater than that of the barreter, operating temperature can be higher, it is more rugged and has a thermal time constant several times longer.

Whichever type of bolometer is used, most power-measuring instruments come in two units. One is the bolometer mount, consisting of the microwave input connection, either waveguide or coaxial, in which the wire or thermistor forms an absorptive termination, plus a relatively large thermal mass. The second is the meter with associated circuitry, connected to the mount by flexible cable carrying d.c. or low-frequency bias signals.

Within the meter the basic circuit is the balanced bridge with automatic feedback shown in Fig. 2 in which the bolometer element forms one arm. A suitable resistance, usually 100 to 200 ohms, is selected for the element to balance the bridge and is achieved by passing low-frequency current through the element. Audio frequency power is usually chosen for ease of measurement and amplification and, in the case of the thermistor, Fig. 1(b) shows that the current required for a single bead lies between 1 and 15mA.

At this stage, with no microwave power present, the bridge is balanced and the meter reading will be zero. With microwave power present, however, the thermistor will be hotter, its resistance will decrease and, in order to maintain a balance, the audio power must decrease by an amount equal to the microwave power, which can then be identified and displayed. There are many refinements adopted to preserve accuracy and stability in commercial instruments, but the balanced bridge remains the heart of the circuit. The dynamic range of these instruments is generally 40dBA with good sensitivity, switchable meter ranges lying between 10 microwatts full scale and 10 milliwatts full scale.

Within the bolometer mount itself lie the sensing elements, usually thermistors, a large thermal mass to help reduce fluctuating external temperatures and also a couple of other thermistors to sense changes in ambient temperature. These latter are not connected to the microwave circuit but form part of another bridge within the power meter and help to distinguish between ambient temperature variations and changes in microwave power. With a large thermal mass, this may appear trivial but, when measuring a few microwatts of power, a 25% meter drift can easily occur with an uncompensated mount.

To be accurate the sensing elements in the microwave circuit must, ideally, form a perfectly matched termination to the transmission line. In the case of standard 50-ohm coaxial line, Fig. 3 shows a mount and method of thermistor attachment. Two beads are used and are mounted between the inner and outer conductors to give the type of electrical circuit shown in Fig. 4. C1 is a d.c. blocking capacitor to eliminate spurious effects from the source and of the circuit and, with a value of 1 to 2nF, has a very low reactance at the microwave input frequency.

For convenience in biasing and matching, two thermistors are used and are connected in series as far as the audio frequency substitution circuit is
concerned. They are biased to a value of 100 ohms each. Capacitor $C_2$ is similar in value to $C_1$ and thus also presents a low impedance to the input signal, the effect being to cause the thermistors to appear in parallel at r.f. Overall impedance presented to the 50-ohm transmission line is thus 50 ohms and helps towards achieving a good match. The compensating thermistors are electrically isolated from the microwave circuit but are in close enough thermal proximity to experience, identically with the detection thermistors, any variation in ambient temperature.

With a range of waveguide mounts also available, the thermistor power meter is a sufficiently accurate and reliable instrument and, for many years, has been the backbone of engineering power measurements.

### Thermocouple power meter

A strong challenge to the position of the thermistor is being made by the thin-film thermocouple and, with the improvements made in recent years, this device now offers several advantages such as very much greater temperature stability and higher burnout levels. The coaxial mount construction is very similar to that of the thermistor except that, instead of fine wire supports, the favoured technology is that of thin film deposition on as thin as possible a substrate.

![Thermocouple power meter diagram](image)

**Fig. 4.** Thermistors which terminate coaxial line appear in series to the d.c. substitution and bridge circuit but in parallel to the high-frequency microwaves. A large thermal mass together with twin compensating thermistors substantially eliminate ambient temperature fluctuations.

An example is shown in Fig. 5, the thermocouple being formed from evaporated bismuth and antimony with the hot junction lying in the gap between centre and outer conductors; the large semicircular contact pads are of gold.

Thermocouple resistance, and hence match, are controlled by the bismuth and antimony film thickness. Incident microwave power absorbed at these hot junctions raises their temperature, giving rise to a thermoelectric voltage which is then amplified and displayed in terms of microwave power.

The voltage generated in this fashion can be as low as several hundred nanovolts and thus requires careful circuit design to avoid odd thermocouple effects creeping in from dissimilar metal junctions and other sources of noise. Amplification is carried out after first chopping the d.c. signal and the signal is then synchronously detected. Unlike the bolometric power element, instrument calibration is carried out by direct comparison with a national substandard and not by d.c. substitution.

Recently an improved form of thermocouple sensor has appeared on the market* using a silicon p-n diffused region as one arm of the thermocouple with a gold contact to act as the cold junction and a resistive tantalum nitride contact as the hot junction. The complete element is less than 1mm square and 0.005mm thick and offers a better r.f. match because of lower resistance, a faster response time and greater long-term stability than the normal bismuth-antimony element.

### Calorimeters

In contrast to the devices so far mentioned, in which microwave power is measured by some calibrated change in an electrical property, the calorimeter relies upon dissipating the incident power within some absorbing medium and then measuring the associated rise in temperature. The microwave power can then either be calculated from a knowledge of the temperature rise versus time and the thermal mass, or a calibration can be made against known quantities of d.c. or low-frequency input power.

Commercially available calorimeters are bulky and expensive devices and have measurement rise times of several minutes. Consequently they are usually kept in the standards room of the user as a calibration reference for the more general-purpose bolometric instruments. Higher-order standards tend to be individually designed and vary from country to country and are always being improved. Thus, although the operating principles are the same, there are many different types of calorimeter. Instead of enumerating these, it will probably be of greater interest to describe briefly one of the instruments which has recently been developed at the National Physical Laboratory for use as a national standard.

### NPL calorimeter

This device is a twin calorimeter and operates in the system shown in Fig. 6 between d.c. and 6,000MHz and for input power levels of between 10mW and

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The instrument uses two coaxial transmission lines of 50-ohm characteristic impedance, each terminated in a matched, power-absorbing, resistive load. Microwave power is fed into one of the lines and the resulting rise in temperature difference between the two matched loads is measured by the thermopile. This unit consists of 800 junction pairs of copper-constantan. The loads and thermopile at the end of the calorimeter are shown in Fig. 7. After the heating effect is noted the microwave power level can be determined by measuring the amount of d.c. power which must be applied to the load to produce the same effect. The symmetry of the calorimeter, together with the double-packed thermal insulation, helps to eliminate the effects of room-temperature fluctuations.

An interesting technique used in the development of this NPL calorimeter and in building up a thermal equivalent circuit was the determination of thermal resistances by localized heating with a laser beam. Instead of supplying heat to the calorimeter and noting the temperature response in different areas with thermocouples, these areas were painted black and irradiated with a laser beam, so producing an output from the thermopile. By calculating the local thermal capacities from a knowledge of dimensions and using the laser technique to measure the local resistances, a very comprehensive equivalent was obtained.

In practice, to speed up the reading time the calorimeter is operated in a feedback loop wherein the output from the thermopile is fed via an amplifier and frequency-compensating network back to the second input of the calorimeter. The load at the end of this line is thus heated to the same temperature as that terminating the microwave input and the microwave power can be determined from the substituted d.c. input power necessary to produce the same power in the feedback loop.

With corrections made for losses in the input lines, power lost in heating the thermopile, load mismatch error and d.c. instrumentation errors, the absolute accuracy in measuring the microwave power lies between 0.2% and 0.5% with a measurement time of several minutes.

So, even for a national standard, the accuracy with which microwave power can be measured is not as good as that for d.c. or low-frequency power. The uncertainty of the d.c. calibration power in the calorimeter, for instance, was an order of magnitude loss. Even so the general purpose commercial instruments are many times worse than this in absolute accuracy and it is worthwhile noting the prime causes of error.

**Power measurement errors**

There are three main sources of error:

- instrumentation error,
- d.c. substitution error,
- mismatch error.

Finally, the largest single source of error and one which can only be partly controlled by the instrument manufacturer is that due to mismatch. The preceding article in this series pointed out the significance of having an obstacle or load impedance which was different from the characteristic impedance of the transmission line. It was seen that such a difference, or mismatch, caused some of the power to be reflected back again to combine with the incident pattern and produce a standing wave along the line.

Unless the quantity of reflected power is known the power meter will read only the amount actually dissipated in its sensing element and not that incident. Here, though, lies a major problem, in that the actual power reflected depends on two things, one being the impedance of the sensing element and the other the impedance of the microwave source plus any other discontinuities in the line. And, being in general complex quantities, their phase controls the manner in which they combine, a factor which is also influenced by the length of transmission line separating them. This phase relationship is usually an unknown quantity so that, unless elaborate and inconvenient tuning procedures are intro-
duced, the day-to-day practice is for manufacturers to design components to as low a v.s.w.r. as possible and for the user to accept the resultant error.

By way of illustration and to recap on previous subject matter: the ratio of the electric field magnitude reflected from a mismatch to that incident is termed the voltage reflection coefficient (ρ) and this reflected signal gives rise to a transmission-line standing wave pattern as shown in Fig. 1 of Part 9. The ratio of the electric field maximum to minimum of this pattern, which can be directly measured, is termed the voltage standing wave ratio (v.s.w.r.), and is universally used as an indication of the degree of mismatch of a component. A perfect match is when \( S = 1 + \rho / 1 - \rho \) remembering that, in general, ρ is the modulus of a more complete reflection coefficient containing phase information.

Now when two or more sources of mismatch are present the amount of power that is actually reflected from any one of them depends upon the way in which the reflected waves combine, that is, upon their relative phase, and this is the quantity usually unknown. It is, however, possible to define a worst case and a best case limit to the resultant mismatch from just a knowledge of the v.s.w.r. involved. Take the case of a microwave generator having v.s.w.r. \( S_1 \) and power monitor of v.s.w.r. \( S_2 (S_2 > S_1) \), then the worst combination would be if one had a v.s.w.r. of \( S_1 \), \( S_2 \) and the other unity, and the best case would be if one were \( S_2/S_1 \) and the other unity.

Taking some practical values of \( S_1 = 1.30 \) and \( S_2 = 1.50 \), which are typical for general test equipment up to J-band (12.4GHz), the worst and best cases result in values of 1.95 and 1.15 corresponding to values of \( \rho \) of 0.32 and 0.07 respectively. The power reflected is proportional to \( \rho^2 \) and so, in this case, the power meter error will lie between -10.2% and -0.5% depending upon the way in which the mismatches combine. A reduction of the power monitor v.s.w.r. to, say, 1.2 would have a significant effect on the error, reducing the uncertainty range to between -4.8% and 0.2%. The way in which other variations in match can effect the power measurement error is shown in Fig. 8 in which the above two cases are plotted as points A and B. It can readily be seen how easy it is to introduce quite large errors into microwave power measurement and how important it is to minimize the mismatch loss of microwave components.

A final point concerning Fig. 8: the percentage error introduced by combining the v.s.w.r.s on this basis is that compared to what would be delivered to a power monitor having an impedance equal to that of the transmission line.

Measurement of frequency

In general the direct measurement of frequency is basically a measurement of time but, because of the manageable size of wavelengths in this region of the spectrum, frequency can also be determined by a measurement of length. An example of this latter method is the slotted line used for v.s.w.r. measurements previously discussed in Part 9.

By moving the sliding carriage, the attached probe samples the periodic standing wave pattern in the transmission line which repeats itself every half wavelength. The position of the carriage is indicated by a calibrated venier scale like that used in vernier calipers or sometimes by a clock gauge. In either case position can, be measured to about 0.1%, but this accuracy is somewhat degraded when applied to measuring wavelength because of the error in finding the identical probe positions on different cycles of the pattern. This method is comparatively laborious and is only used nowadays either as a teaching aid or in those cases of dire emergency when one's own frequency counting system has broken down and one can't borrow a replacement from someone else.

A second frequency-measuring instrument, and the most widely used of all, is the wavemeter. Many designs exist but all are based on noting the response of a three-dimensional microwave cavity at its point of resonance, this point being adjustable. A popular method is shown in Fig. 9, which illustrates a cylindrical cavity into which slides an adjustable spindle. The cavity is loosely coupled to the main transmission line so that a small amount of microwave power can enter.

With the spindle withdrawn completely from the cavity a waveguide mode can exist, and the cavity will appear as an electrically resonant circuit with a resonant frequency determined by its diameter. As the spindle is inserted the resonant frequency is reduced from this upper limit and becomes a function of the spindle length, \( L \), and the cavity now supports a hybrid mode consisting of the original waveguide one and a TEM mode due to the coaxial section formed by the spindle.

Finally, as the spindle penetration becomes greater, the fringing capacitance between the end of the spindle and the base of the cavity starts to influence the resonant frequency, which starts to decrease quite rapidly. The Q factor of this type of microwave resonant circuit is a function of the ratio of cavity to spindle diameters and can be in the vicinity of 1,000.

A practical realization of the instrument is shown in Fig. 10, in this case for use with miniature coaxial connectors. The right-hand component is inserted in series with the coaxial line carrying the frequency to be measured so that the microwave power enters the left-hand connector, say, and leaves via the right-hand one, which might be terminated in either a crystal detector or a power meter. The hole by which a sample of the power can be coupled out can be clearly seen between the two connectors.

On the left of Fig. 10 is the other part of the wavemeter, which contains the cavity, spindle and frequency readout.
Fig. 9. Popular version of the wavemeter cavity in which the resonant frequency is governed by the spindle penetration. Cavity is coupled electrically to the main transmission line and the frequency is indicated by the point at which the cavity absorbs power at resonance.

Fig. 10. Commercial wavemeter operating from about 5 to 18GHz, showing the series-mounted section and coupling hole on the right. Bolt-on unit on the left comprises the tunable cavity and frequency readout.

and which bolts onto the other component. Thus, when measuring an unknown frequency, the large drum carrying the scale is slowly rotated, thereby turning also a micrometer thread carrying the spindle and varying its penetration into the cavity. When the point is reached where the cavity is resonant at the transmission line frequency, it will absorb energy from the main line and a sharp dip in output will be observed from the detector or power meter.

The instrument is calibrated from a frequency standard and the advantage of this type of readout is that it enables a large, finely graduated scale to be used. In this case the unwound scale length is about 2m and the measurement accuracy is ±0.1%. The larger type N connector in the photograph is an additional facility and enables the resonant condition to be identified by connecting a detector to monitor the power absorbed into the cavity.

Progressing in complexity (and cost), one comes to the frequency counter type of instrument which is a true frequency meter in that it actually counts the cycles of a periodically varying waveform. For microwave frequencies the counter usually consists of two sections: a low-frequency part containing a crystal-controlled reference oscillator, digital counter and digital display and a high-frequency section containing a transfer oscillator, harmonic selector and r.f. input.

The transfer oscillator consists of a conventional low-frequency oscillator circuit operating at, say, 100MHz and it could either be a highly stabilized one or tunable by several tens of MHz either side of the fundamental, depending upon the approach adopted by the manufacturer. Whichever it is, the output from the oscillator is fed to an harmonic generation circuit producing usable outputs up to, say, the 100th harmonic.

Taking the tunable version as an example, the oscillator output plus harmonics would be fed to a tunable mixer along with the input signal to be measured. Harmonic selection circuitry then allows the harmonic nearest to the unknown frequency to be selected, to producing a low i.f. from the mixer which is displayed on an integral c.r.t. The fundamental oscillator can then be tuned to a frequency which gives a zero beat between the two mixer inputs, at which point the unknown frequency is now a known number of harmonics up on the fundamental and is displayed on a digital readout. In arriving at this correct display it is necessary to know the fundamental oscillator frequency and this is counted directly by shift register in the low-frequency section. Here a reference oscillator accurately times the opening of a sample gate while the number of cycles passing is counted.

Accuracy of these counters is ± one count in the low-frequency section, in this case at a nominal 100MHz plus crystal stability. This latter is typically 1 in 10⁸ per week or 2 in 10⁻⁷ per second with high stability options giving 1 in 10¹⁰ per day. Most instruments also possess a switchable a.f.c. loop which eliminates zero beat error in the case and also enables f.m. signals to be counted.

At present, commercial instruments are available with transfer oscillator plug-ins which enable frequencies of up to 40,000-MHz to be directly measured. But, in principle, the technique can be applied to higher frequencies still.

Acknowledgment Many thanks to my old friends at the Sanders Division of Marconi Instruments Ltd for the photographs used in Figs. 3, 5 and 10 and also to Dr Alan F. Fanton at the NPL for details of the 6GHz calorimeter.

Sixty Years Ago

ONE of the earliest methods of viewing response curves, now the province of cathode-ray and pen recorder instruments, was described in the issue of Wireless World for July 1915. Designed by Dr. J. A. Fleming (Sir Ambrose, of diode fame) the instrument was named the "Duddell" and was reminiscent of the Duddell oscillograph, in that it was "all done by mirrors".

A long, narrow mirror, mounted with its long axis horizontal, was connected by a cord to the spindle of a rotary potentiometer (a new device invented for this instrument) and was tilted as the pot was turned. The wiper of the pot derived a voltage which was applied to the device under examination (detector, valve, etc.). A mirror galvanometer, whose light spot was directed on the long mirror, was deflected by the dependent variable signal, such as anode current, the two together forming X and Y axes of the display, projected either onto a screen or photograph plate. Alternatively, the pot could be replaced by a variable capacitor, when the instrument could be used to plot resonance curves.

Examples of photographs obtained in this way were shown in the article and included valve characteristics, hysteresis curves of iron wire and resonance curves. An instrument which uses a similar principle was described by H. J. N. Riddle in the issue for November, 1971.
Amateurs and emergencies

From time to time amateur radio finds itself firmly at the centre of the world stage — unfortunately most often in connection with major natural disasters that disrupt normal telecommunication services in affected areas. Magazines from Australia and New Zealand reflect two such events: one last Christmas, the other in 1931. 

*Electronics Australia* describes how when the cyclone struck Darwin on Christmas Day one of the first links to be established, was a mobile station in a car 13km outside Darwin operated by VK2BNN, his wife VK2BYL and VK8JT. They made contact with amateurs in Victoria more than 1,600 miles away, providing one of the few channels for police and emergency traffic. On Boxing Day an emergency s.s.b. net was established on 14.111 MHz with VK3AUP, Melbourne acting as control. VK8CW at Alice Springs acted as a relay station when required. By December 27 the net had become a nationwide system with participating amateurs in Cairns, Townsville, Rockhampton, Mackay, Mt Isa, Brisbane, Lismore, Armidale, Sydney, Canberra, Cooma, Melbourne, Adelaide, Perth and Alice Springs. 

The New Zealand journal *Break-in* reports the death of James Mills, ZL2BE of Hastings who for many years was a leading figure of amateur radio in that country and whose activities attracted world interest in February 1931 when a major earthquake shattered the towns of Napier and Hastings. James Mills was one of the few amateurs having a rotary generator that could be run from car batteries and was able to make contact on 3.5 MHz, first with other amateur stations and then with the official New Zealand Government station ZLW at Wellington, handling a very large number of emergency messages. Later a relief expedition was organized and was accompanied by ZL2BO who set up a station there. The events generated enormous goodwill towards amateur radio and the country still maintains an Amateur Radio Emergency Corps which is frequently called upon to help; two recently reported cases involved a search for lost hikers who got into difficulties crossing the Roaring Meg River and a coastal rescue when a trawler went ashore off New Plymouth. 

In the United Kingdom the "Raynet" or Radio Amateurs' Emergency Network of the RSGB maintains preparedness for emergency operations through local controllers and groups and by regular exercises in conjunction with the British Red Cross Society, the St John Ambulance Brigade and the police. They are readily available to provide communications assistance on request from the user organisations in conditions where there is a real risk to human life, in the belief that Raynet is a way in which radio amateurs can use their knowledge as a service to the community. Fortunately, the occasions on which Raynet is called upon in earnest are relatively rare; most "on air" activity is during simulated emergencies. Nevertheless it is a service that believes in being ready and willing.

**Hourly propagation forecasts from WWV**

For amateurs and short-wave listeners a source of hourly propagation data is the American standard frequency transmissions from WWV (Colorado) and WWVH (Hawaii) on 2.5, 5, 10, 15, 20 and 25 MHz. These now include at 14 minutes past each hour an indication of solar flux (in the form 72 plus 0.6K) and geomagnetic activity (0 to 9 K scale).

In general for good h.f. conditions the higher the solar flux figure the better (for the next few years this is unlikely to exceed 100); conversely a low K figure (preferably 2 or under) is a good sign. High K figures indicate a significant influx of solar particles, usually resulting in weaker signals; increased fading and noise: over 4 usually indicates a solar storm; 3-4 unstable or unsettled conditions. Unfortunately at the present time the standard frequency transmissions heard most strongly in the UK are seldom those from WWV or WWVH.

There is growing evidence to suggest we are now very close to the end of the present sunspot cycle, with its oddly distorted decay during 1972. Although it is unwise to make long term predictions about future sun-spot activity it looks increasingly as though the next cycle may have a relatively low maximum.

The ZB2VHF beacon station at Gibraltar is now in operation on 144.145 MHz beaming signals towards the UK.

**Look no batteries!**

The energy crisis has made numbers of amateurs look quite seriously into the question of how radio communication could be maintained completely independently of mains supplies or primary batteries. Some recent experiments have been based on solar batteries, wind generators and pedal- and hand-operated battery chargers. *QST* for example reports that a man on a Jacked-up bicycle can generate 100 to 120 watts of power by driving the generator at over 1,100 r.p.m. using the high gearing provided by a 27-inch wheel (generators for this purpose were dropped by the RAF to the French resistance during World War II).

But for sheer ingenuity a prize must surely go to a Dutch amateur J. M. H. Wagenmans, PA0HWE who claims to have built a million watt transceiver powered entirely by the action of the Morse key! He does this by linking the moving arm of a Morse key to the core of a moving-coil loudspeaker so that the movement produces an electrical output which is rectified and stored in a 40,000µF capacitor to power a transistor crystal oscillator to a d.c. input of 1.5 mW. Brass pounding with a vengeance; although a nagging doubt remains, despite the photographs and circuit details in the April issue of the Dutch journal *Electron*: April issues of amateur journals are notorious for elaborate technical spoofs, though it is difficult to fault this idea!

**In brief**

The RSGB has stated that it appears that the 25% VAT rating applies to all amateur radio equipment and components... The recent use of the callsign GB2IARU at Tonbridge School to mark the 50th anniversary of the formation of the IARU is believed to be the first time a four-letter callsign suffix has been authorised for amateur radio operation in the UK. During 1976, the bicentennial year of the constitution of the United States, American amateurs will be able to use callsigns beginning with an "A" instead of "W" or "K": all prefixes will have two letters before the district number, ranging from AA1 to AL7 and including AC4 which was formerly used by amateur stations in Tibet and thus one of the most eagerly sought after prefix of all time... American amateurs are concerned that a Dallas consulting firm has filed a proposal with the FCC for a new television channel that would result in the elimination of the 50MHz amateur band... A 10GHz beacon station at a temporary site on the Isle of Wight has been heard at distances up to 65km, the beacon uses an 80mW oscillator on 10,100 ±1MHz with an omnidirectional aerial with a gain of about 11dBi. Callsign is GB3IOW... Among forthcoming mobile rallies are: Longleat near Warminster on June 29 by Bristol RSGB Group; Upton by Worcester society on July 6; Cornish RAC Rally, Camborne (provisional) on July 20 (details G3NKE); Polegate Steam Engine Rally on July 20 with exhibition station GB2SS.

PAT HAWKER, G3VA
75 Years of magnetic recording

5 - 'A diversity of applications

by Basil Lane

Assistant Editor, Wireless World

Shortly after the end of World War II, events took an unexpected turn for magnetic recording. Up to that time, although other uses had been suggested for this versatile storage method, the technology had not developed to the stage where they could be practically realized. However, the war effort had resulted in new electronic techniques becoming available and at the same time magnetic recording itself came of age. From that date forward, the number of applications for magnetic recording were to multiply.

Curiously, for the historian, the task of recording events in the recent past becomes more difficult the closer one approaches the present day. There could be many reasons for this, but in the case of magnetic recording it is because technology from 1945 advanced at such a rate that new developments followed one another at an incredible pace. This makes it difficult to say at times who was first in the field with a particular idea. One can only hope to describe from contemporary reports what happened.

A typical example is the computer. Where now we can hardly think of a computer without also thinking of the magnetic storage methods used, the earliest computers were without such an advantage. Suddenly, everyone seemed to be working on the idea of using magnetic tape as a storage medium. In an historical broadsheet put out by 3M a few years ago\(^5\) it was suggested that flexible storage systems were one of the most significant steps taken by computer designers, and in the days just after the war a variety of memory devices were proposed including delay lines filled with gin!

Just about the earliest computer with a magnetic store was the ARC (Automatic Relay Calculator) made at Birkbeck College, London in 1947. Built for the British Rubber Research Association, it had a nickel plated drum store with a capacity for 236 numbers each of 21 binary digits. Shortly after, the experience gained in developing this computer was used to develop the SEC (Simple Electronic Computer), the first all-electronic computer with a drum store.

Also in 1947, Eckert and Mauchly built a machine called BINAC for the Northrop Corporation, this computer using a mercury delay-line memory of 512 words and being the first to use a magnetic tape input and output.

Although magnetic drum and disc scored for short term storage with fast, random access, there was an increasing requirement for long term serial data stores of greater versatility than the punched paper tape and card type. Thus in the early 1950s several machines appeared with a tape storage system, one such machine being the IBM 701. This was the first of the IBM machines to use a combined drum and tape store, supplemented with Williams tubes to give a total memory capacity of 2048 words.

Most of the development from that time on, concentrated on the improvement of the magnetic media, not only to reduce the error rate due to drop-outs but also to increase the capacity of the media to record digital bits. Thus the capability rose from 100 bits per square inch in 1947 to 1500 in 1965\(^6\) and as high as 6500 bits per square inch today — although there are at the moment few, if any, machines that are capable of recording such a bit density. Although plated metal discs had been used from time to time, the real development of this form of recording medium did not start until Zaponi\(^7\) developed a reliable plating technique in 1952. It is interesting to note that 46 years earlier, P. O. Pedersen had patented\(^8\) a plated magnetic carrier though this was of course intended for audio recording.

In modern computers a variety of storage media is used, with magnetic methods still paramount. Disc is used for random access and usually is manufactured as a pack of discs in a standard format. The magnetic layer may be acicular Fe\(_3\)O\(_4\), plated nickel-cobalt or other similar magnetic metals and the writing and reading are achieved by flying heads located in the appropriate position in the disc pack by high speed actuators. Magnetic tape is still strongly favoured for long term storage where access time is not so important and will be either ferric oxide, cobalt-doped ferric oxide or chromium dioxide, all of the latter having been developed from the early sixties.

The most popular material used in magnetic tapes is still the leader, this being gamma ferric oxide. Since the

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Fig. 2. The TR-22 video recorder by RCA, this was the first transistorized video recorder to enter service, May 7, 1961.
early days of its use as the basis for a coating formulation for magnetic tape, it has been improved and the coating formulation itself developed almost beyond recognition. However, it was inevitable that some challenge to the supremacy of ferric oxide should come from other materials and the demands placed on magnetic tape by the invention of the video recorder and the data recorders for computers provided the impetus for this to happen during the 1960s. From around 1956, cobalt doping of ferric oxide had been studied as a method of improving the short wavelength response of tape and by 1967 an experimental tape had been made. However, the problems associated with this type of magnetic material (e.g. pressure instability of magnetisation) prevented it from becoming as popular as more conventional coatings.

Another line of research had been undertaken by Swoboda who had been concentrating on the problem of synthesising chromium dioxide, a promising material for tape since it had a high remanence, which offered a very good short wavelength performance. In early 1961 this substance was produced successfully, though it was not found to perform as well before it made its debut as a magnetic tape. As a matter of interest, it took ten years for Cr₂O₃ to appear in a compact cassette, when Crolyn (the Du Pont name for Cr₂O₃) appeared under labels such as Advent and Memorex.

It is interesting to note that, due to a quirk of American Patent law which allows applications to be patented as well as processes, Du Pont held the master patent for the use of Cr₂O₃ as a magnetic recording medium in America. Process patents are held in all the other countries of the world where applications are not patentable, and it was this situation that led to Agfa Gevaert trying to circumvent this Patent in 1972. They too had invented a process for the manufacture of Cr₂O₃, which differed from the Du Pont method, but were prevented from selling any tape in America because of the patent held by Du Pont. The ridiculous situation of being barred from one of their important markets eventually forced Agfa Gevaert to take the Patent to court. In 1972, the court decided that Agfa Gevaert could not sell a tape for the Du Pont process and pay for the privilege of selling Cr₂O₃ tape in America.

Since the date of the appearance of Cr₂O₃ as a magnetic medium for cassette tapes, other manufacturers and those who were licencees had been attempting to either equal the performance of this substance, using cheaper materials, or oxide, or to, as it were, busy looking at other substances. One development that appeared to result from a parallel, but separate, development project was the dual layer coated tape. These had been proposed as early as 1953, when Kornej suggested that short wavelength performance could be improved if a multi-layer tape was prepared with a high coercivity surface layer, followed with successively deeper layers using oxide of a lower coercivity. Another quite novel idea proposed by Gabor and Bauer was that a dual layer tape should be made with the top layer oriented in a vertical direction.

However, it was Sony and 3M that eventually produced a practical dual layer tape, with Sony marketing its product as early as 1974 and, later that same year, 3M announcing its own product. Again, it would seem that the master patent is probably held in America by 3M, so it is highly likely that Sony have had to come to some cross-licence agreement with 3M in order to sell their tape in the USA.

Developments in heads

No history of magnetic recording would be complete (and this one is far from being so) without a mention of the development of ferrite heads. With the appearance of Cr₂O₃ tape, which is a much more abrasive material, there was a particular requirement for a hard wearing head to be produced which would withstand many hours of use. This was very important in video recording where the head to tape speed is high and the track width is narrow.

Early heads had been made of laminated Mumetal and although experiments had been made with harder materials such as cobalt, this proved, until recent years, too intractable to be used as a substitute. Permalloy and other harder grades of metal have been used with increasing success, but it was the ferrite head that made the big news when it was first developed. One of the first descriptions of a ferrite head appeared in 1955, to be followed some years later by the classic paper by S. Dukner of Philips, which described the method of using glass as a spacer and as a bonding material in the front gap of the head. This was a significant improvement that reduced the number of rejected heads that had suffered from chipping at the gap edges.

By 1968, Matsushita had developed the hot pressed ferrite head and about this same time Akai appeared with a glass-crystal ferrite head, made from a monocrystalline block of ferrite.

Other important head types used in specialist fields were the flux sensitive heads, either using a saturable limb in the magnetic circuit of the head, or the semiconductor Hall effect. Described by Camras in the mid-1950s, it was C. Pedersen who was kind enough to write and point out an error in spelling P.O. Pedersen's name in earlier parts of this series.

My apologies to Mrs Pedersen who was kind enough to write and point out an error in spelling P.O. Pedersen's name in earlier parts of this series.

References
68. The birth of the computer, pub. 3M Company 1971.
70. U.S. Pat. 2,619,454; P. P. Zapponi, Nov. 25, 1925.
Thick-film amplifier
A thick-film hybrid amplifier suitable for audio applications will deliver around 15W (average) into an 8Ω load. The class B quasi-complementary circuit has a frequency response from 0 to 80kHz, an input sensitivity of 350mV (typical), and a t.h.d. figure of 0.2%. The device is mounted on an integral heat sink measuring 30 x 30mm which allows full rated output at a temperature of 55°C. Tadiran, 193 Regent Street, London W1.
WW312 for further details

De-soldering instrument
Adcola Products have introduced an automatic de-soldering instrument for removing d.i.l. i.c.s from p.c. boards. The unit, which is called the Removic, consists of an operating gun powered from a control box which adjusts the temperature from 350 to 750°F. In operation the gun is placed over an i.c., a handle is pressed which positions extractor claws and heater blocks on the i.c. When the solder melts the component is extracted by applying steady pressure on the handle. Adcola Products Ltd, Adcola House, Gauden Road, London SW4. WW308 for further details.

Mini drill
A small power drill manufactured by Expo operates from a 12V 1A supply and has a chuck speed of around 9000 r.p.m. The drill, which is supplied with a range of accessories which include twist drills, cutting, milling, reaming and grinding tools, is priced at £9.17 plus v.a.t. and is available from Electroprian Ltd, P.O. Box 19, Orchard Road, Royston, Herts SG8 5HI.
WW310 for further details

Alarm unit
The Tellit is an audible alarm unit using a continuous loop of tape which may be recorded with any short message or warning specified by the user. Four basic types are available including an uncased playback mechanism for mounting into customers equipment. Highland Electronics Ltd, 33 Dallington Street, London EC1V 0DB.
WW311 for further details

Counter
A 75-MHz counter/timer, model 5308A, measures frequency, frequency ratio, period, period average, and time inter-

Triacs
ITT Semiconductors have made their first step into the triac market by announcing the TC range of devices. These components, which have been designed and manufactured in the UK, are available with current ratings of 4.6, 8, 10, 12 and 16A at voltages of 200, 400, 500 and 600V. The complete range uses a centre-gate construction for improved dv/dt capability, and has glass passivated chips which improve high-voltage protection. All of the devices are housed in the TO-220AB plastic package and are priced between £0.416 and £1.88 each depending on type and quantity. ITT Semiconductors, Foots Cray, Sidcup, Kent.
WW 335 for further details

Microwave coupler
Walmore are now supplying the Norsal 4834 microwave coupler. This device, which is claimed to be the first to cover the range 2 to 12.4GHz, has a v.s.w.r. of 1.4:1, amplitude imbalance of ±0.5dB, and a phase imbalance of ±7°. The coupler is designed to handle input powers of 20W average with a 2kW
Variable filters
The EF3/03 and EF3/04 are recent additions to the Barr & Stroud range of variable filters. The units are high-pass and low-pass filters respectively which may be switched into several modes including band-pass and band-stop. The cut-off frequency is variable from 0.1Hz to 100kHz with a stop-band attenuation rate of 48dB/octave and a pass-band response from 0 to 700kHz.

Signal generator
The Fluke synthesized signal generator, model 6010A offers a keyboard entry with a manual fine control, and a built-in microprocessor which can store up to nine programmes. Each programme consists of a frequency, a frequency range, an amplitude, and a modulation (c.w., a.m., or f.m.). These programmes can be recalled at any point in a testing sequence. The instrument has a frequency range from 10Hz to 100kHz in 0.1Hz steps and 10Hz to 10MHz in 10Hz steps, an accuracy of ±3 parts in 10^6 in the temperature range 0 to +50°C and a variable output level from 0.25mV to 5V r.m.s. The selected frequency is indicated on a seven-digit l.e.d. display. Fluke (Nederland) B.V., Ledeboerstraat 27, Tilburg, Netherlands.

Testing system
The type 8309 test set connects directly to a computer terminal and will transmit to, or receive messages from the terminal for the purpose of checking and/or fault analysis. The 8309 can be operated at data rates between 110 and 9600 bits/sec in a synchronous or asynchronous mode. In the former mode it responds to either the EBCDIC or ASCII SYN character set, while in the latter case the number of stop bits sent by the transmitter may be selected as required. Data Recognition Ltd, Loverock Road, Battle Farm Estate, Reading, Berks.

Bezel
A recent addition to the Roxburgh range of switches and accessories is a bezel incorporating a l.e.d. mounting facility. The bezel fits the J50, J60 and 800J switches and accepts a 0.2in l.e.d. Roxburgh Electronics Ltd, 22 Winchelsea Road, Rye, Sussex.
Sansui tuner

The TU-7700 stereo tuner is one of a series of amplifiers and tuners which Sansui have added to their range. This model will receive f.m. broadcasts within the frequency range 88-108 MHz and a.m. broadcasts in the range 535-1605 kHz. Containing a high integrated-circuit count, the TU-7700 is to retail at a suggested price of £149.10 plus VAT. Vermitron Ltd, Thornhill, Southampton SO5 9QF.

Four-channel tape deck

Well known on the Continent, but new to the UK, the Dokorder range of reel-to-reel recorders is being distributed by Acoustico Enterprises. The top of the domestic range is the model 7140 which is fully equipped to record and replay up to four tracks on 1 1/2in tape. The head block is interchangeable to offer two track stereo recording and a consequent improvement in signal-to-noise ratio.

A useful facility is offered by the so-called “Multi-Sync” which is a switching arrangement which connects one or more record head tracks to the replay amplifier to permit monitoring of previously recorded tracks, whilst adding in synchronism further recordings to remaining free tracks.

Other facilities include sound-on-sound, echo, bias and record equalisation for two types of tape and A-B monitoring. Acoustico Enterprises Ltd, Unit 7, Space Waye, North Feltham Trading Estate, Feltham, Middx TW14 0TZ.

Eagle range

Eagle International have added considerably to their range of products with the introduction of the 2000 series of equipment. This comprises the A2004 and A2006 amplifiers with a rated output of 20W per channel and having a rather more complex range of tone controls than usual. A matching tuner amplifier, the A2008, provides for reception of the a.m. and f.m. bands and also there is a synchronous motor, belt drive turntable, the D2005. A second turntable, the D2006, is classified by Eagle as a servo-monitor type, though exactly what this means is not quite clear. Finally the 2000 series is completed by a choice of five loudspeakers and two headphones. Eagle International, Heather Park Drive, Wembley, Middx HA0 1SU.

New belt drive turntables

The BDS80 and BDS90 are two turntables added to the series produced by BSR McDonald. Both models are semi-automatic in that when a record has finished playing the arm lifts up, returns to rest, locks and switches off. Both decks are also fitted with a click suppressor and muting switch which prevent an unwanted noise being fed to the amplifier while the automatic set-down and switch off are in operation. A special feature of the BDS90, according to the manufacturer is a new low-resonance tubular aluminium tone-arm, located in a concentric gimbal style mount which carries a calibrated stylus pressure control. The rumble figure for both units is stated to be 55dB. Standard units are available for either 100-125V 60Hz or 200-240V 50Hz. BSR Ltd, Monarch Works, Cradley Heath, Warley, Worcs.

Three new speakers

A new range of loudspeakers from Celestion are the UL6, UL8 and UL10 models. The two smaller models in the range are both two-unit systems, the bass driver being supplemented by a passive auxiliary bass radiator unit. The largest model is the UL10 (see photograph) which houses three in-line drive units. Cabinet dimensions are 673mm height, 317mm width, 380mm depth. Power handling is rated at 50W continuous or 100W peak music power. Sensitivity is sensibly rated using pink noise, an input of 12 volts r.m.s. being required to produce 96dB s.p.l. at 1 metre. Frequency response is stated to be 40Hz to 20kHz ±2dB. Drive units consist of the HF2000 super tweeter, the MD7000 pressure-dome mid range unit and a 254mm bass drive unit. Rola Celestion Ltd, Ditton Works, Foxhall Road, Ipswich, Suffolk IP3 8JP.

New life for Wharfedale

Rank Radio introduced a new XP range of speakers carrying the well-known names of Denton, Linton and Glendale. All have greater power handling capability and are claimed to have improved performance over their predecessors. Taking the smallest in the range, the Denton 2XP, this is a two-way bookshelf enclosure of acoustic suspension.
Amplifier with limiter circuit
The new Cambridge Classic One amplifier has several unusual features including I.E.D. monitoring of pre-amplifier signal level in 5db steps. An I.E.D. indicator is used as a power supply indicator. This also flashes at a regular rate whenever a power amplifier limiter circuit is brought into operation. A circuit in conjunction with the input selector switch is provided to limit the effect of switch-on transients causing 'thumps' through the loudspeakers. Main specifications are: power output at 1kHz, 60W per channel into 8Ω; t.h.d. is less than 0.05% at all audio frequencies, typically 0.005%; signal-to-noise ratio is 65dB ref. 2mV (pickup input). 75dB ref. 250mV (line inputs); input sensitivities are pickup 2mV/47kΩ, tuner 250mV, cassette 100mV and auxiliary 100mV. The pickup input overload capability is rated at 45dB. Facility for the simultaneous use of three tape recorders is also provided. Cambridge Audio Ltd, Lamb House, Church Street, London W3 2PB.

Solid State Devices
Names of suppliers of devices in this section are given in abbreviation after each entry and in full at the end of the section.

Shift register
A 512-bit dynamic shift register with built-in recirculating and command logic has a maximum operating speed of 4MHz.
WW350 for further details  G.E.

Diodes
The 501PD series of diodes has an average current rating of 500A at an 84°C case temperature, and are available over the voltage range 2kV to 4kV with non-repetitive ratings up to 4.4kV.
WW351 for further details  I.R.

R.a.m.
The 1103-1 is a 1k dynamic r.a.m. offering a 340ns full-cycle time and a 150ns read access time. All address inputs are fully decoded and typical power dissipation is 0.4mW/bit.
WW352 for further details  ITT

Schottky register
A 4-bit register having three-state and standard t.t.l. outputs has been built using Schottky technology. The device is available in three packages and is priced from $3.08 100 up.
WW353 for further details  AMD

Programmable op-amp
Parameters of the MC3476 op-amp are programmed by an external resistor to suit power supplies from ±6 to ±15V. The device does not require frequency compensation, has offset null capability and is protected against short circuits.
WW354 for further details  Motorola

Bridge rectifiers
A new range of bridge rectifiers has a 75mH capacitor in parallel with each diode to absorb reverse voltage transients. The range has repetitive peak reverse voltage ratings from 50 to 600V and an average direct current rating of 2A.
WW355 for further details  ME

Memory driver
An n-channel m.o.s. memory driver i.c., type MC3459, contains four identical driver circuits and is capable of maintaining a propagation delay of 20ns when driving a 360pF load.
WW356 for further details  Motorola

A-to-d converter
The MC904 is claimed to be the first monolithic i.c. with all the functions necessary for a digital panel meter or multimeter. The device has a resolution of 100µuV and an input impedance of 1000MΩ and is housed in a 28-pin d.i.l. package.
WW358 for further details  Macro

Image sensor
A new 1728-element charge-coupled linear image sensor, suitable for use in optical page scanning systems, has been introduced by Fairchild. The CCD121 is capable of reading a standard 8½ × 11-inch page in less than one second and up to 200 lines per inch can be resolved.
WW359 for further details  Fairchild

Count/display i.c.
A t.t.l.-compatible universal count/display i.c., type ZN1040E, offers a count rate of over 5MHz. The device requires a 5V supply and is housed in a 28 pin d.i.l. package.
WW360 for further details  Ferranti

Low-voltage sensor
An i.c. intended for use with battery operated equipment, is internally set to trigger when a voltage drops to 2.4V ±2%. This value can be increased by adding a resistor.
WW361 for further details  Bowmar

Sample/hold amp
An adjustment-free sample/hold amplifier called the MN343 provides a droop rate of less than 0.3mV/ms and an acquisition time of better than 10µs.
WW362 for further details  Tranchant

F.e.t. input amplifier
The 3670 is a f.e.t.-input i.c. instrumentation amplifier. Gain adjustment, from 1 to 10000/V, is made with one resistor. Maximum bias current is 10pA, and the input impedance is 100 ohms.
WW363 for further details  Burr Brown

G.E. Electronics (London) Ltd, 182/4 Campden Hill Road, Kensington, London W8 7AS.
International Rectifier, Hurst Green, Oxted, Surrey RH8 9BB.
ITT Semiconductors, Footscray, Sidcup, Kent.
Advanced Micro Devices Inc., 901 Thompson Place, Sunnyvale, California 94086, U.S.A.
Motorola Ltd, Semiconductor Products Division, York House, Empire Way, Wembley, Middx.
Macro Marketing Ltd, 396 Bath Road, Slough, Bucks.
Ferranti Ltd, Electronic Components Division, Gem Mill, Chadderton, Oldham, OL9 8NP.
Bowmar Arizona, Inc., 2355 West William I. Field Road, Chandler, Arizona 85224, U.S.A.
Tranchant Electronics (UK) Ltd, Tranchant House, 100a High Street, Hampton, Middx.
Burr-Brown International, 25a King Street, Watford, WD1 8BT.
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- 12": £8.88
- 14": £9.88
- 16": £7.87
- 18": £6.87

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DM344A Distortion Factor Meter. Designed to make accurate and rapid measurements of total harmonic distortion generated within high quality audio amplifiers, recording and transmission equipment. Selling Price: £172.50 + VAT.

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AM324 AF Millivoltmeter. Designed for voltage measurements in the audio and low frequency ranges and for use in conjunction with distortion measurement equipment in audio systems. Selling Price: £75.00 + VAT.

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### LOW VOLTAGE TRANSFORMERS

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### Safety Marks Issuing Transformers

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<td>20</td>
<td>4</td>
<td>0.75</td>
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### Battery Charger Types

- **Primary 200-250 Volts 12 And 24 Volt Range**
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- **60 Volt Range**

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- **2N2222**
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Pl. 3 Semiconductor set  £4.70

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Pl. 2 Resistors, capacitors, pots  £2.40
Pl. 3 Semiconductor set  £3.35

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Pl. 3 Semiconductor set  £3.10

BAILEY-BURROWS PRE-AMP
Pl. 1 F/Glass PCB  £2.05
Pl. 2 Resistors, capacitors, pre-set, transistors  £4.95

FOR DETAILS OF COMPONENT PACKS FOR THIS DESIGN PLEASE WRITE FOR FREE LIST.

20 WATTS/CHANNEL

STUART TAPE RECORDER
A set of three printed-circuit boards has been prepared for the stereo integrated circuit version of this high-performance Wireless World published design.

TRIP PA. 1 Reply amplifier F/Glass PCB  £0.90
TRIP PA. 1 Record amp., meter drive cct. F/Glass PCB  £1.40
TROS PA. 1 Bias/erase stabilizer cct. F/Glass PCB  £1.00

FORWARD DETAIL CROSSOVER

An essential and critical component in a high-quality speaker system is the crossover unit conventionally comprising of a series of passive networks which unfortunately, though introducing reactive impedances between the amplifier and the speakers, result in the loss of the advantage of high-gain amplifiers damping factor and renders the speakers prone to overshoots and resonances. An elegant solution to this problem, described by D. C. Read in Wireless World, involves the use of a series of active filters splitting the output of the pre-amplifier into three channels, of closely defined bandwidth, each of which is fed to the appropriate speaker by its own power amplifier. A design for a suitable 20-watt amplifier, based on a proven Texas circuit, was also described by Mr. Read. The printed-circuit board for this has been designed such that three amplifiers may be stacked and mounted together on a common heatsink to achieve a conveniently compact module.

ACTIVE FILTER CROSSOVER Package

SEMICONDUCORS AS USED IN OUR RANGE OF QUALITY AMPLIFIERS

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20 WATTS/CHANNEL

ACTIVE FILTER CROSSOVER Package

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POWER SUPPLY FOR 20W/CHANNEL STEREO SYSTEM Package

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<tr>
<td>F/Glass PCB</td>
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<tr>
<td>Reg. diodes, zener diodes, capacitors, fuse holders</td>
<td>£8.00</td>
</tr>
<tr>
<td>Toroidal transformer</td>
<td>£4.95</td>
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</table>

TOROIDAL T20 + 20
Developed by Texas engineers and published in a series of articles in Practical Wireless, The TEXAN was a remarkable breakthrough in delivering true Hi-Fi performance at exceptionally low cost. Now further developed to include a true Toroidal transformer, this slimline integrated circuit design, based upon a single F/Glass PCB, features all the normal facilities found on quality amplifiers, including scratch and rumble filters, adaptable input selector and headphones socket.

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<table>
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<th>Type</th>
<th>Single</th>
<th>Dual</th>
<th>Knob</th>
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<td>35p</td>
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<td>35p</td>
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- **B84**
  - 100 Silicon Diodes 6N7 Mn glass equivalent to 1N418

PLASTIC POWER TRANSISTORS

40 WATT SILICON

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| 90 WATT SILICON

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<td>30p</td>
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</table>

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£7.50

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Complete kit is described — Television £19.50 plus 40p for UK

MAINS TRANSFORMERS

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<tr>
<th>Type</th>
<th>AC</th>
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<td>A 18V 1 amp suitable for SS 103</td>
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<td>B 25V 2 amp suitable for SS 110</td>
<td>£2.00</td>
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<td>C 30V 3 amp suitable for SS 140</td>
<td>£2.50</td>
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Add 3p for P&P per transformer

BRIDGE RECTIFIERS

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<th>Type</th>
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<td>A 45V/1 2A 70% 8 &amp; 100V 2A</td>
<td>35p</td>
<td></td>
</tr>
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</table>

MICRO RELAYS 230 240V AC 3 change over heavy duty 120p For GPO telephone engineers each 55p

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**DESIGNER APPROVED KIT**

In Hi-Fi News there was published by Mr. Linsley-Hood a series of four articles (November 1972–February 1973) and a subsequent follow-up article (April 1974) on a design for an amplifier of exceptional performance which has as its principal feature an ability to supply from a direct coupled fully protected output stage, power in excess of 75 watts whilst maintaining distortion at less than 0.01% even at very low power levels. The power amplifier is complemented by a pre-amplifier based on a discrete component operational amplifier referred to as the Lithium which is employed in the two most critical points of the system, namely the equalization stage and tone control stage, positions where most conventional designs run out of gain at the extremes of the frequency spectrum. Unusual features of the design are the variable transition frequency, the tone controls and the variable slope of the scratch filter. There is a choice of four inputs, two equalized and two linear, each having independently adjustable signal level. The attractive slimline unit pictured has been made practically by highly compact PCBs and a specially designed Toroidal transformer.

**NOVEL STEREO FM TUNER**

In the April and May 1974 issues of Wireless World there was published by J. Skingley and C. P. Thompson a novel design for an F.M. tuner which combines consistent high performance with the elimination of the critical setting procedure required by too many earlier tuners. The front end is ready built pre-aligned module which then feeds an amplifier driven screened three section ceramic filter leading to an integrated circuit five-stage limiting amplifier providing excellent r.m.s. rejection. This is followed by a single coil integrated balanced demodulator from which the audio output may be taken. Temperature compensated varicap tuning allows stations to be selected within a ±500Hz tuning potentiometer or by a choice of six preset push-button controls. Each of the preset controls can be adjusted on the front panel with the settings being indicated by six LED lamps behind an acrylic screen printed facia panel insert. Additional circuitry includes temperature compensated AFC restricted to less than station spacing, inter-station muting, a single-lamp LED tuning indicator and a linear scale frequency meter. The stereo decoder, built on a separate board, is based on a well-proven integrated circuit phase-locked-loop to which has been added active filters to remove sub-carrier harmonics and 'birdies'. The power supply, to ensure station holding stability, uses an integrated circuit voltage regulator which is powered via a low-inductance field-specimen designed TOROIDAL TRANSFORMER.

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**MARQUEE**

**STEREO PRE-AmPLIFIER TYPE PA100**

This stereo pre-amplifier has been conceived from the latest circuit techniques. The PA100 stereo pre-amplifier incorporates on less than eight million point transistors, two of these are especially selected low noise N.P. devices for use in the front stages. These pre-amplifiers are specified to be equal in quality performance of the PA100, which also has a STEREO/BAND switch, volume, balance and continuously variable bass and treble controls.

**SPECIFICATION**

- **Frequency response:** 20Hz-20kHz flat to +0.1dB & -0.3dB
- **Input Impedance:** 40kΩ 50Hz-20kHz
- **Output Impedance:** 40kΩ 50Hz-20kHz
- **Signal to Noise Ratio:** Better than 60dB
- **Frequency Response:** 20Hz-20kHz
- **Input Sensitivity:** 80mV

**PRICE**

£14.45

**STEREO 200**

The Stereo 200 amplifier is included in every unit and is tested on a plug-in tester before leaving the factory. The unit embodies the best methods of design and construction. The stereo 200 stereo pre-amplifier is supplied with the mid-frequency system. This amplifier is suitable for use with any type of loudspeaker, and is designed to provide a clean and powerful signal to the next stage of the amplifier. The unit is simple to install and is ideal for use with any type of loudspeaker.

**PRICE**

£25.00

www.americanradiohistory.com
TRANSMISTORS AND DIODES (25%)  

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INTEGRATED CIRCUITS (8%)  

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CATHODES (8%)  

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Circuits 250V AC each £25  

- AMP TYPE  
- WATT  
- TYPE  
- VOLTAGE  
- CURRENT  
- POWER  

CIRCUIT BREAKERS  

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FILTERS  

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CARES INTEGRATOR Circuit 250V AC each £25  

- AMP TYPE  
- WATT  
- TYPE  
- VOLTAGE  
- CURRENT  
- POWER  

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Sinclair Scientific kit

Britain's most original calculator – now in kit form
The Sinclair Scientific is an amazing calculator. It offers logs, trig, and true scientific notation over a 200-decade range – features normally found on calculators costing around £50 or more.
Yet even ready-built, it costs a mere £21.55 (including VAT).
And as a kit it costs under £15!
Forget slide rules and four-figure tables
On the Scientific, you can handle directly all three trig functions, their inverses, log_{10}, antilog_{10}, giving quick access to x (including square and other roots).
Plus, of course, the four arithmetic functions and any calculation based on them.
In fact, virtually all complex scientific or mathematical calculations can be handled with ease.

Sinclair Cambridge kit
At its new low price, the Sinclair Cambridge kit remains unbeatable value.
The Cambridge is now Britain's most popular pocket calculator. And it's not surprising. Check the features – then ask yourself what other calculator offers such a powerful package at such a reasonable price.

Take advantage of this money-back no-risk offer today
The Sinclair Cambridge and Scientific kits are fully guaranteed. Return either kit within 10 days, and we'll refund your money without question. All parts are tested and checked before despatch – and we guarantee any correctly-assembled calculator for a year. This guarantee also applies to calculators supplied in built form.
Simply fill in the preferential order form below and post it – today!

Sinclair Radionics Ltd,
London Road, St Ives, Huntingdon, Cambs., PE17 4JH. Tel: St Ives (0480) 64646

So is the Scientific difficult to assemble?
No. Powerful though it is, the Sinclair Scientific is a model of tidy engineering.
All parts are supplied – all you need provide is a soldering iron and pair of cutters. Complete step-by-step instructions are provided, and our Service Department will back you throughout if you've any queries or problems.
Of course, we'll happily supply the Scientific or the Cambridge already built, if you prefer – they're still exceptional value. Use the order form.

Features of the Scientific
- 12 functions on a simple keyboard
- Scientific notation
- 200-decade range
- Reverse Polish logic
- 25-hour battery life
- Genuinely pocketable

Features of the Cambridge
- Only 43\,\text{mm} \times 2\,\text{in} \times 1\,\text{in}. Weight 3\,\text{oz}.
- Fully-floating decimal point.
- Algebraic logic.
- Constant on all four functions (+ - x -).
- Constant and algebraic logic combine to act as limited memory.
- Clear, bright 8-digit display.
- Operates for weeks on 4 AAA batteries.

To: Sinclair Radionics Ltd, FREEPOST, St Ives, Huntingdon, Cambs., PE17 4BR.
Please send me
Scientific kit – £14.95 inc. VAT
Scientific built – £21.55 inc. VAT
Cambridge kit – £9.95 inc. VAT
Cambridge built – £13.99 inc. VAT

* I enclose a cheque for \ldots\ldots\ldots made out to Sinclair Radionics Ltd, crossed.
* Please debit my *Access/Barclaycard account number:

Signed
Name
Address

Please print: FREEPOST – no stamp required.
OUTSTANDING VALUE ONLY

Unit containing selector mechanism

COIN Mfg. Airflow Developments Ltd., continuously Precision built to high

3 Price
240V. Post 185

3 c:
(I) Coil ohms;

Price

ce
heavy

MINIATURE switch. 10 for latching
16, 50 amp contacts. 70 ohm coil
or

POWER lb.

Tel.: 01-437 0576

Certainly great to have around the house.

VAT AT THE APPROPRIATE RATE MUST BE
ADDED TO ALL ORDERS FOR THE TOTAL VALUE OF GOODS INCLUDING POSTAGE.

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BARGAINS NOW AVAILABLE . . .

Top class used instruments from such famous names as Tektronix, Hewlett-Packard, Solartron, Philips, STC, Racal, etc., etc.

ALL IN SUPERB CONDITION AND AT REDUCED PRICES!

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<tr>
<th>AUDIO EQUIPMENT</th>
<th>Sale Price Range</th>
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<td>£100</td>
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<td>£88</td>
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<td>SR268 Source &amp; Detector</td>
<td>£109</td>
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<td>B221, B221L1, LCR Bridge</td>
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<td>B224, LCR Bridge</td>
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<td>Hewlett Packard 3734, 5 digits, 3-50MHz</td>
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<td>5271A, 8 digits, 5 Hz-10MHz</td>
<td>£80</td>
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<td>5218, 7 digits, ±12Hz</td>
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<tr>
<td>5245, 8 digits, DC-50MHz, 3 x 10⁻³ per day ageing rate</td>
<td>£90-£550</td>
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<td>5246, 6 Digit Counters DC-50MHz</td>
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<td>5/52A, Pre-Scaler plug-ins, 350MHz</td>
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<td>5253B, Converter plug-ins, 0.12MHz</td>
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<td>5255A, Converter plug-ins 3-12.4GHz</td>
<td>£280-£480</td>
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<tr>
<td>5260A Automatic Divider Unit, DC-12GHz (with 50MHz Counter)</td>
<td>£570-£980</td>
</tr>
</tbody>
</table>

Marconi TF1411/6, Digital 0-15MHz | £130 |

<table>
<thead>
<tr>
<th>DEVIATION METERS AND POWER METERS</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>Marconi TF7910, 4.1024MHz</td>
<td>£132</td>
</tr>
<tr>
<td>TF2502, Power Meter</td>
<td>£130</td>
</tr>
<tr>
<td>TF934, Power Meter</td>
<td>£80-£120</td>
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<tr>
<th>DIGITAL VOLTMETERS</th>
<th>Sale Price Range</th>
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<tbody>
<tr>
<td>Davis 3800A, Digital Multimeter 0-1 kV Max RDG 1999</td>
<td>£98-£140</td>
</tr>
<tr>
<td>5230, DVM 0.02% 10V Max RDG 19999</td>
<td>£246-£450</td>
</tr>
<tr>
<td>5530, DVM 0.02% 1kV Max RDG 19999</td>
<td>£280-£610</td>
</tr>
</tbody>
</table>

Dynamco DM2022, D40 DC, 0.02% 10V-resolution 2kV | £98-£340 |
| DM2140/B1/B1, Mean AC Converters | £98-£240 |
| DM50/B1/B3, RMS AC Converters | £61 |

Hewlett Packard 34880 and Range of plug-ins (Complete) | £540 |
| 3480 Main Frame, 5 Digits | £198-£480 |
| Complete with 3482 DC plug-in | £349 |

Sokkia LM-1420, 2.0 0.05% 2.5 uV resolution to 1kV DC | £98-£190 |
| LM1420B2A, DC and RMS/Mean AC | £319-£400 |
| AC112B, A.C. Converter 30 mV-300V Mean Reading | £40-£190 |
| LM1480/B, Max. RDG 29999 5uV-2kV DC | £280 |
| LM1604, Max. RDG 9999 1µV-1kV DC | £280-£420 |
| LM1887, Max. RDG 101999, 10µV-1kV DC | £259 |

<table>
<thead>
<tr>
<th>OSCILLOSCOPES</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tektronix 546, DC-50MHz, Main Frame with delayed sweep dual time base</td>
<td>£240-£350</td>
</tr>
<tr>
<td>547, DC-50MHz, Main Frame with delayed sweep dual time base</td>
<td>£280-£390</td>
</tr>
<tr>
<td>1A1, Plug-in for 546 or 547. DC-50MHz 5mV-20V/Dual Trace</td>
<td>£90-£290</td>
</tr>
<tr>
<td>1A2, Plug-in for 546 or 547. DC-50MHz 50mV-20V/Dual Trace</td>
<td>£70-£120</td>
</tr>
<tr>
<td>565, DC-80MHz, Main Frame with Delay Time Base</td>
<td>£240-£440</td>
</tr>
<tr>
<td>568, Plug-in for 565. DC-80MHz 100mV-20V/Dual Trace</td>
<td>£60-£90</td>
</tr>
<tr>
<td>564, DC-10MHz/140MHz Storage or Sampling Main Frame</td>
<td>£220-£330</td>
</tr>
</tbody>
</table>

Necessary plug-ins supplied (Real time and Sampling) | £60-£440 |

Solara 5171B, DC-50MHz, Main Frame | £500-£890 |
| CX1741, Plug-in DC-50MHz 50mV-20cm Dual Trace | £60-£190 |
| CX1744, Plug-in Time Base 0.5-5s/cm (Computable) | £285-£390 |

Telequipment 0535 with 2 X 'A' Input Storage Dual Trace | £220 |
| A, DC-15MHz, Single Trace Amplifier plug-ins (Two Plug-ins Per Main Frame) | £10-£220 |

<table>
<thead>
<tr>
<th>POWER SUPPLIES</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>30/10, 0.30V, 10A, Pre-set</td>
<td>£38-£509</td>
</tr>
<tr>
<td>55E, 0.15%, 1A Pre-set</td>
<td>£22-£400</td>
</tr>
</tbody>
</table>

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<thead>
<tr>
<th>RECORDERS</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>5-127 Ultra Violet Light Beam, 12 Channels</td>
<td>£269</td>
</tr>
</tbody>
</table>

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<thead>
<tr>
<th>SIGNAL SOURCES</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>General Radio 1363. Ultra Oscillator (needs power supply)</td>
<td>£80</td>
</tr>
<tr>
<td>Multiband D80A, 0.01Hz-12kHz, Decade 2 Phase</td>
<td>£120-£198</td>
</tr>
<tr>
<td>1kHz-1111kHz, Decade 0-7V</td>
<td>£230</td>
</tr>
<tr>
<td>Marconi TF442L/4, 10kHz-12MHz, Xtal check Int/Ext AM 50 MHz</td>
<td>£198-£420</td>
</tr>
<tr>
<td>TF801D/1, 10MHz-470MHz Int/Ext AM and Pulse modulation</td>
<td>£225-£599</td>
</tr>
<tr>
<td>Wayne Kerr 0.220, 10kHz-10MHz Video Oscillator</td>
<td>£98</td>
</tr>
</tbody>
</table>

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<thead>
<tr>
<th>SPECTRUM ANALYSERS</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hewlett Packard B5518/BS1, 10MHz-12GHz (Extends to 4GHz with extra accessories)</td>
<td>£2,310</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>SWEEP GENERATORS</th>
<th>Sale Price Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>89693, 3-7.8GHz plug-ins (For use with 89690 Main Frame)</td>
<td>£290-£498</td>
</tr>
<tr>
<td>TELEPHONE TV AND MICROWAVE</td>
<td></td>
</tr>
<tr>
<td>Hewlett Packard</td>
<td>£2,420</td>
</tr>
<tr>
<td>Marconi QA 2000A, White Noise Test Set</td>
<td>£712</td>
</tr>
<tr>
<td>(Ifills also available at extra charge)</td>
<td></td>
</tr>
<tr>
<td>TF 2905, 5. Sin 2 P &amp; B Gen, 225 Lines 60Hz</td>
<td>£298</td>
</tr>
<tr>
<td>TF 2909, Gray Scale Gen, 625 Lines</td>
<td>£400-£750</td>
</tr>
<tr>
<td>PM 5508B Pattern Gen 625 Lines PAL UK Systems</td>
<td>£159-£280</td>
</tr>
</tbody>
</table>

| REDUCED PRICES! | |
| 15P TV Studio Precison Signal Generator, Sin 2 P & B Window, Staiacase 525 lines | £197 |
| Requires all drives (reduced to 625 lines as arranged at extra charge) | £197 |
| Siemens REL535, Contact Fault Locators, 1 MHz Test Signal Variable labels, High sensitivity | £80 |
| STC 74186, Multimeter Test Sets | £36-£450 |
| 74184B, Selective Measuring Sets | £80 |
| 74216, Noise Generators, 20kHz-4kHz | £80-£180 |
| 74308, Oscillograph 10kHz-10MHz | £65-£110 |
| 74600, R.F. Attenuators, 10 steps each unit total Att 0.9 | £8-£90 |
| 7482B, Selective LevelMeasuring Set | £80-£100 |
| Wandel & Guettelmann TFM43-1, 16MHz, Selective Meters | £210-£410 |
| VZM1 Differential Phase Meters (TV) | |
| VZM01 Sampling Attachments (Complete) | £310 |

| TRANSFER FUNCTION ANALYSER | |
| JM1606, /XM1639 0.0001Hz-15kHz Phase resolution 10 minutes of arc | £1,800-£4,200 |
| WAVE ANALYSERS | |
| 248A, 5-30MHz, Harmonic Analyzers | £132 |

<table>
<thead>
<tr>
<th>ASSOCIATED EQUIPMENT</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Marconi TD1, Differential Voltmeter, 0-1000V</td>
<td>£200</td>
</tr>
<tr>
<td>RC24, D.C. Electronic Multimeter</td>
<td>£50</td>
</tr>
<tr>
<td>Siemens Multitron, R.F. Voltmeter, 0-1000V</td>
<td>£50-£100</td>
</tr>
<tr>
<td>Marconi M19, 30kHz Recorder, 4 channels, 8khz at Mains</td>
<td>£850</td>
</tr>
<tr>
<td>Marconi CAT 90C, Headset and Mic Amplifier, Complete</td>
<td>£550</td>
</tr>
<tr>
<td>MP Surface Pyrometer 50-600°C (with Probe)</td>
<td>£190</td>
</tr>
</tbody>
</table>

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E.H.T. POWERUNIT.

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**CASED TRANSFORMERS**

Rated 40 watt iron cored split case with 2 core power unit for sale. Suitable for general purpose work. £5.00 each. 240 volt output (Please consult your supplier). Dimensions: 150 x 250 x 200 mm. 75 Amp. 5000 V.A. 250 volt output (Please consult your supplier). Dimensions: 150 x 250 x 200 mm. 75 Amp. 5000 V.A.

**SAFETY INSULATING**

- Nom. 120-240 VAC, 120-240 volt Centre Tap with screen
- 240 volt and 230 volt Centre Tap with screen
- 240 volt and 230 volt Centre Tap with screen
- 240 volt and 230 volt Centre Tap with screen
- 240 volt and 230 volt Centre Tap with screen
- 240 volt and 230 volt Centre Tap with screen

**MINIATURE & EQUIPMENT**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**12 and 24 VOLTS PRIMARY 200-240 Volts**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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- 240 volt Centre Tap with screen

**MINIATURE NEONS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**PHONE DIALS** (New) £1.

**FUSE HOLDERS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**OVERLOAD CUT-OFFS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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- 240 volt Centre Tap with screen

**ADVANCE TRANSFORMERS “VOLSTAT”**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**ADVANCE TRANSFORMERS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**COPPER LAMINATED P.C. BOARD**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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- 240 volt Centre Tap with screen

**COILS**

- 240 volt Centre Tap with screen
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**CABLES**

- 240 volt Centre Tap with screen
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**CIRCUIT BREAKERS**

- 240 volt Centre Tap with screen
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**CUT-OUTS**

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**DEVICE REED RELAYS**

- 240 volt Centre Tap with screen
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**FUSE LINKS**

- 240 volt Centre Tap with screen
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**FUSES**

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**HOLES**

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**INDUCTORS**

- 240 volt Centre Tap with screen
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**LATCH RELAYS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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**LUMINOUS CIRCUIT BREAKERS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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**LTD TRANSFORMER**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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**MINIATURE NEONS**

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**MUTTER RELAYS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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**NEW! 2" AND 4" PANEL METERS**

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**WE REGRET THAT ALL ORDERS VALUE UNDER £5 MUST BE ACCOMPANIED BY THE REMITTANCE**

**HIGH-SPEED MAGNETIC COUNTERS**

- 240 volt Centre Tap with screen
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- 240 volt Centre Tap with screen

**RELAYS, SIEMENS-VARLEY PLUG-IN.**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**WATT CARBON FILM RESISTORS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
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**MINIATURE NEONS**

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**COMPLETELY ASSEMBLED, GIANT 100MM**

- 240 volt Centre Tap with screen
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**EATON CONNECTORS**

- 240 volt Centre Tap with screen
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**FUSE BASES**

- 240 volt Centre Tap with screen
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**FUSE HOLDERS**

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**MINIATURE “ELAPSED TIME” INDICATORS**

- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen
- 240 volt Centre Tap with screen

**TRANSFORMERS**

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STEREO IC DECODER HIGH PERFORMANCE PHASE LOCKED LOOP (as in VW - July '72)  

MOTOROLA MC1310P DELIVERY SPECIFICATION  
Separation 40dB 50kHz: ±1kHz  
I/F level 5650mV rms  
Input impedance 2 kohm  
Will drive up to 75mA stereo on lamp or LED.

KIT COMPRISSE FIBREGLASS PCB  
Rubber feet, screws, perspex. Panel front. PMT Components: Comprehensive Instructions  
LIGHT ELECTRA TUBES  
Suitable tubes on indicator for above  
MC1310P only £2.15 plus p.p. 10p.

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MINIATURE UNSELECTORIZED P.O. Type 2230 1 2 1.2 watts bridging. 1 mm bridging wires. This condenser takes up 25% less space and occupies no more than the standard 3000 type Relay. £6.80 ex.

MINIATURE DIGITAL INDICATOR size of sticky 'n' stick illuminated by red or night light filament lamps weight 35g only 0.9 x 0.9 x 1.1 240V. 50Hz. 100cf 1.2 x 1.2 x 1.2

SUGGESTED SMALL ORDER Q 100 units 100 ¥ 0.95 ex.  
MINIATURE DIGITAL INDICATOR size of sticky 'n' stick illuminated by red or night light filament lamps weight 35g only 0.9 x 0.9 x 1.1 240V. 50Hz. 100cf 1.2 x 1.2 x 1.2

HIGH SPEED COUNTERS £2.75 each + £1.25 x 10 counter per second with 4 figures. The counter is capable of counting up to 10000 volts. Suitable for industrial use. This 10000 counter is normally used with an auxiliary indication within ±0.1% accuracy.

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- Unit Oscillator, General Radio Type 1200C, 600pH, 50/25MHz, Accuracy 1%, 0-100%, Output into 50 ohms — 150 MW with 951 1201 — 2010 as illustrated.
- ADVANCE Audio Signal Generator H18 15H-50 kHz, 200V-20mvmt, 208Site Wave & Square Wave.
- Audio Signal Generator J18 15H-50 kHz, 0-2V0V.
- VHF Square Wave Generator 5011 S. kHz-10MHz, Max 0-1 V p-p.
- Signal Generator D1 ARM/E.
- S.S.D. Generator 702 30H-30MHz, 3 bands.
- Signal Generator 701.
- H.F. Signal Generator 201 200MHz in 1 MHz bands. Input: 0.1 input to 0.1 output level variable.
- 1/11 1V 75 ohms impedance -50 dB at 1kHz, Ext. Mod. 30MHz 505 - 1655.

**Audio Oscillator**

- GENERAL RADIO
  - Audio Oscillator 1214 A. P.O.
- HEWLETT PACKARD
  - U.H.F. Signal Generator 614A 900-2100MHz, Accuracy Output. 0-1 U.V. to 50 MHz, 0.5% level variable.
  - Audio Oscillator 201 C, 20 Hz-70 kHz, 0 dB in 1 dB steps. Distortion less than 0.5%. 5kHz, 100kHz, 200kHz, 1MHz.
  - F.M./A.M. Signal Generator 2012A, 50KHz.

**SIEMENS**

- MARCONI INTS.
  - Oscillator 6214 20-230MHz Suitable for use with Marconni TF 1215 & 1225A 0.05 sec.
  - FM/AM Signal Generator TF 954A, 5-1 MHz 770 kHz, 2V-2V0Vp-p. Superheterodyne.
  - Marconni 6251, 500kHz.
  - Signal Generator TF 1205, 70Hz-70kHz, 0.005% level variable.
  - MMG 1, 300MHz.

**Audio Signal Generator**

- Siemens Level Meter 3D 235 10kHz 1.1MHz Complete system by Siemens. Compares 3D 5V Full Scale Oscillator 3D 335 Synchronous Meters 3W 833 Sweep Attachment 3D 346 Sweep Transmission 3D 348 Scope Level Testing Receiver 3D 335 Level Measuring Meter 3D 323 0.125-1kHz Output Signal Transformer 39 0.31-1.2kHz.
  - ETC.
  - Decade Filter 7414A 30kHz 12.8MHz For analysing gain and lowering an interference signal system, particularly useful with 7412 precision decade品味.
  - Selective Level Measuring Set 7418 60 16kHz.

**TELEPHONE TEST EQUIPMENT**

- NUMERICAL TELEPHONE CLUSTER.
  - Tube B 2159 C 0.1mg. 10kHz Bandwidth.
  - NUMERICAL TELEPHONE CLUSTER.
  - Tube B 2159 C 0.1mg. 10kHz Bandwidth.

**Components**

- TELEPHONE TEST EQUIPMENT.
  - NUMERICAL TELEPHONE CLUSTER.
  - Tube B 2159 C 0.1mg. 10kHz Bandwidth.

**Pulse Generators**

- ADVANCE Double Pulse Generator PG 56
  - Pulse Amplitude 0.5 10V 500 kHz, 10V at 1000 p-p.
  - Pulse Generator PG 55.
  - P.O. Modular Pulse Generator Advance Type PG 52 System at 5 Signal generating & Processing units. Registration times up to 20MHz & Output pulses to 50 (500) ohm rise & fall times 0.5ns, its versatility enables the production of complex pulse & ramp waveforms, not obtainable from conventional test generators.

**Resistance & Decade Boxes**

- MARCONI Digital Pulse Generator TF 1400 S 10MHz 5V 1 kHz, 5V 1kHz 10V 1MHz.
  - NEGRETTI & ZAMBRA Quick Reading Potentiometer.
  - TRINITY Variable Potentiometer Type 32/18 A precision equipment as used in laboratories and electrical measuring laboratories.

**T.V. Test Equipment**

- MARCONI Signal Squared Pulse & Bari Generator TF 2920, 300MHz.
  - GRESHAM LION Waveform Generator 625 lines Square-squared Pulse.
  - Pre-amplifier Generator 625 lines.

**Miscellaneous**

- Telecommunication Type 7500V 6kV, Output 7500V 50Hz without counter. Sensitivity 1000V 10psec.
  - MARCONI Digital Factor Tester Type 1422, Fundamentals Frequency Range 100Hz-10kHz. Basic frequency range 10 to 10kHz, 100kHz, check points up to 500kHz. 2 & 10MHz.
  - Portable Recorder Type TF 880 5kHz-10kHz 10kHz-100kHz.
  - Precollator Transformer TF 374, 10kHz checked, 5kHz, 1kHz, 10kHz, 100kHz, 1MHz.
  - Whim-Philips TF 4536, 10kHz.
  - Whim-Philips TF 1301 2000 100kHz.
  - Advance Calibrator Type 9035.
  - R.P.L.
  - Megger Megger Type 154 10kHz-100kHz.
  - Megger Megger Type 154 10kHz-100kHz.
  - Megger Megger Type 154 10kHz-100kHz.
  - Megger Megger Type 154 10kHz-100kHz.
  - Megger Megger Type 154 10kHz-100kHz.
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101 - DC 50 Mag 5mV/cm Dual Trace £15.00

102 - DC 50 Mag 50mV/cm Dual Trace £60.00

CA - DC 24 Mag 10W/cm 5A58 DC to 33 Mag calibrated sweep delay £325.00

546 Dual time base/Delayed sweep DC 50 Mag £57

547 Dual time base/Delayed sweep alternate display DC 50 Mag £325.00

Newlett Packard

Model 130C 200-240V £160.00

This scope is a versatile all purpose instrument for laboratory: educational: or industrial process measurement and medical applications. The output of rf detectors: strain gauges: transducers: and other low level devices may be viewed directly without preamplification. The Model 130C is easy to operate even by inexperienced personnel. Specifications: Tube: Range - 1.5 cm on 6 cm. 21 ranges in a 1.25 logarithmic accuracy ± 3%: viewing provides continuous adjustment between steps and extends the 5 cm step to at least 12.5 cm (automatic triggering: horizontal: and vertical: input: bandwidth: Bandwidth: d.c. coupling: dc to 500 KHz: a coupled input: 10: he to 500 KHz: an: un-coupled: 25: he to 500 KHz: at 0.2 mV/cm deflection factors)

135 Scope 17BA DC - 50 MHz Main trace £100

Various plug-ins available: for £25 each

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(a) A degree in a scientific or engineering subject.
(b) Degree-standard membership of a Professional Institution.
(c) A Higher National Certificate or Higher National Diploma in a scientific or engineering subject.
(d) A qualification equivalent to (c) above.

In addition the following relevant experience is required:

(a) Applicants with 1st or 2nd class honours degrees — at least two years' postgraduate experience.
(b) Applicants with other qualifications — at least five years' post-qualifications experience.

Salary Scale: £3254 - £4454 with entry point dependent upon experience beyond the minimum required.

**Senior Scientific Officer**

Applicants should be at least 25 and under 32 years of age, although the upper age limit may be waived if experience of special value can be offered.

Applicants should have obtained a 1st or 2nd class honours degree and have had a minimum of four years' appropriate post-graduate experience.

Salary Scale: £4185-£5778. Entry will normally be at the minimum of the scale but applicants with experience of special value may be entered above the minimum.

Applications stating the field of work and the grade required should be made to:

**Administration Officer**

HM GOVERNMENT COMMUNICATIONS CENTRE
Hanslope Park
Hanslope
MILTON KEYNES MK19 7BH

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**Looking for a new job?**

We have regular contact with hundreds of Electronics and Electrical companies needing qualified technicians and engineers and can therefore help you find an interesting and well paid job. All you need do is to return the coupon below or give us a ring.

Our service is confidential and costs you nothing.

**TJB Technical Services Bureau,**

3A South Bar,
Banbury, Oxfordshire.
Banbury (0295) 53529

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Please send me an “Application for Employment” form
NAME ..................................................
ADDRESS ...............................................
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Find your place in British Gas

Training in Engineering

CONTROL AND INSTRUMENTATION

British Gas has vacancies from time to time for control and instrumentation engineers at graduate level, technician engineer level and technician level. The posts concern demand good standards of ability and a high sense of responsibility. To help fill them, we are prepared to train small numbers of selected men of suitable educational and technical ability who have previously worked in a related technological field, such as light current electrical engineering, or electronic control engineering.

The courses are designed to train suitable men quickly to a standard which will enable them to contribute effectively to the Industry's work at the earliest possible stage, following which expertise will be developed through more advanced training.

If you are thinking of a change of career and would like to be involved in an undertaking which gives scope for personal development, and at the same time contributes to the wellbeing of the country, write in confidence, giving full details of your age, educational and technical qualifications and experience, and indicating the area of the country in which you would prefer to work, to:

Mr. T. A. Lucas,
Communications and Instrumentation Dept. (WW),
British Gas, National Westminster House,
326 High Holborn, London WCIV 7 PT.

BRITISH GAS

C.C.T.V. SYSTEMS

Teletape Video, U.K.'s most progressive video systems company, are seeking a top man to organise and run a new division formed to actively develop the C.C.T.V. security surveillance activities of the company. Existing international contracts run into six figures, and we hold distribution rights on all good agencies so the opportunities are unlimited.

We are looking for a man fully experienced in all aspects of this business with good product knowledge and capable of dealing at Government level on large contracts. He should be capable of working entirely on his own initiative and will report directly to the Managing Director.

We envisage a good basic salary plus a sensible incentive participation on sales, with a company vehicle, etc.

This is a unique opportunity.

Please apply in confidence to:
Iain Crummond
Managing Director
TELETAPE VIDEO
76 Brewer Street, W1R 3PH
Telephone 734 1319 or 434 1287

Opportunities in the ELECTRONICS FIELD

Men with analogue or digital qualifications / experience seeking higher paid posts in: TEST - SERVICE - DESIGN SALES.

Phone: Mike Gernat, Ref. W.W.

NEWMAN APPOINTMENTS
380 Oxford Street, W.1,
01-629 0501

Applied Physicist or Electronics Engineer

required by the Institute of Cancer Research at Sutton, Surrey, to join an active R and D programme on ultrasonic diagnosis of cancer. The work will be varied and responsible in a joint hospital/research institute environment and will call for a range of practical abilities in the development and application of novel instrumentation. Previous experience in ultrasonics would be useful but is not essential.

Appointment will be on the MRC Technical Officer (JTO) scale at a salary in the range £2,688 - £3,792 (JTO) or £1,484 - £2,592 (JTO) plus Threshold award of £229 plus London allowance of £312 Scales under review

Applicants should normally hold an appropriate degree or equivalent qualification and preferably have some R & D or industrial experience. Exceptionally possession of two appropriate "A" levels can be accepted for appointment to a junior post.

Applications with curriculum vitae in duplicate and naming two referees to: The Secretary, Institute of Cancer Research, 34 Summer Place, London SW7 3NU, quoting ref. 300/0/85.
How do you value your electronics testing ability?

All the benefits of the world's largest radio-telephone exporters could soon be yours. If you value your expertise highly, this is where to get most mileage from it.

Career progression paths are long and wide—the company's expansion rate has been unaffected by the present economic situation. Equally significant is the importance Pye Telecom rightly attaches to quality and testing to exacting specifications. Their VHF/ UHF advanced design communications equipment, not only is reliability crucial in furthering the company's progress, but frequently lives depend on the performance of the equipment, because fire, police and ambulance services use it extensively.

If you have practical experience of this work, maybe in the armed forces, it will pay you handsomely to get full information about the conditions, the relocation assistance and other attractions. Work and live in the attractive university city of Cambridge—alternatively in the nearby expanding town of Haverhill where there are excellent possibilities for private and rented housing.

Phone or write today to:

Mrs Audrey Darkin
Pye Telecommunications Ltd
Cambridge Works
Elizabeth Way, Cambridge CB4 1DW
Tel: Cambridge 8985

Mrs Cath Dawe
Pye Telecommunications Ltd
Colne Valley Road
Haverhill, Suffolk CB9 8DU
Tel: Haverhill 442

RADIO ENGINEER

Telerenters (London & Provincial) Limited have vacancy for Senior Engineer at Watford.

The Engineer should have detailed knowledge of RF and audio measurements and be familiar with European specifications to enable him to implement and manage a test laboratory for the type approval of domestic radio, audio and recording equipment. Factory experience essential.

Salary: £4,000-£5,000.

Write giving brief details to: The Director, Telerenters (London & Provincial) Limited, 155/159 Queen's Road, Watford, Herts WD1 2QH.
RADIO TECHNICIANS

Are you a Radio Technician with a City and Guilds Intermediate Telecommunications Certificate or equivalent, plus 1 year's practical work-shop experience? If so, then why not join the Home Office. We have a vacancy at Baldock, Hertfordshire to carry out installation, maintenance, modification and construction of complex specialised radio communications equipment and systems.

Pay
£2010 at 17, £2230 at 19, rising to £3385 a year.

A Secure Future
with a non-contributory pension scheme, good prospects of promotion and a generous leave allowance. 5-day week of 42 hours.

Interested?
Then telephone or write for an application form to:
Mr J J Willis, Directorate of Radio Technology, Room 514, Waterloo Bridge House, Waterloo Road, London, SE1 8UA. Telephone 01-275 3006.

SERVICE ENGINEER

Required to service our range of scientific and laboratory instruments, which include Fraction Collectors, U.V. Monitoring, Ultramicrotomy equipment. The applicant should be resident in an area North of the Thames to Luton or prepared to move. A good working knowledge of modern electronics and a scientific background is desirable.

The Company offers excellent working conditions including pension scheme, profit sharing bonus scheme, BUPA membership and Company car.

Write or telephone for an application form.

The Service Manager
LKB Instruments Limited
232 Addington Road
Selsdon, South Croydon, Surrey
01-657 8822

LKB INSTRUMENTS LTD.

CATV - MATV ENGINEERS AND PRODUCTION MANAGER

Canada's Leader in Cable Television requires personnel for research and production departments:
Openings for research in amplifier, passives, converter, aerial designing - your choice.
Opening for CHIEF ENGINEER - should have CATV or related experience. Electronics person strong in leadership. Methods and Mechanical Acumen required for capacity of PRODUCTION MANAGER.
Good salaries, generous benefits. Please Airmail complete Personal History and references to:
Mr. J. E. Thomas
Lindsay Speciality Products Ltd.
50 Mary Street
Lindsey, Ontario
Canada

CHELSEA COLLEGE UNIVERSITY OF LONDON

ELECTRONICS ENGINEER

Required for interesting design and development work in an Electronics Workshop, catering for the prototype equipment requirements for teaching and research in the Departments of Electronics and Physics.
Salary £2,849-£3,305 per annum, including London Allowance.
Five-day, 37½-hour week.
Generous holidays.
Application forms and further details from Mr. M. E. Cane (5EW), Chelsea College, Pulton Place, London SW6 5PR.

CHELSEA COLLEGE UNIVERSITY OF LONDON

ELECTRONICS EXPERIENCE

required to take charge of Electronics Workshop for the design and production of prototype electronic equipment for electronics and physics research and teaching, and also for the servicing and maintenance of a wide range of commercial electronic equipment.
A wide practical experience and a sound theoretical knowledge of electronics is essential. Experience in microwave instrumentation would be an advantage. Five-day, 37½-hour week. Salary (Technical Staff Grade 6) £3,254-£3,860 per annum, including London Allowance.
Further details and application form from Mr. M. E. Cane (5W7), Chelsea College (University of London), Departments of Electronics and Physics, Pulton Place, Fulham, London SW6 5PR.

www.americanradiohistory.com
ELECTROSONIC LTD SE LONDON

SENIOR DEVELOPMENT ENGINEER
AUDIO PRODUCTS c£4200 pa

Electrosonic Ltd are seeking a professional engineer with at least five years experience of developing audio equipment. He will primarily be required to make significant technical contributions to the company’s new products which will range from AV replay equipment to studio mixers.

It is also expected that through his commercial awareness of the audio field he will help to define the product range and expansion in this important area of the company’s business.

MANAGER —
ELECTRONIC TEST DEPARTMENT c£3,500 pa

A candidate is required having wide experience in a production test shop. Technical ability in analogue and digital circuitry is essential together with experience of supervising the work of others.

Duties will include the organisation and day-to-day running of the test shop, providing technical oversight, training of junior engineers, the introduction and programming of automatic test equipment and supervision of quality control.

TEST AND SERVICE ENGINEERS
£2,400-£2,800 pa

Vacancies exist in both these departments for electronic engineers having a minimum of two years experience of control and/or audio systems. On the job training will be given and opportunities for advancement are available. Service engineers will be required to work both in the factory and on site and the holding of a current driving licence is desirable.

The company is leader in the rapidly expanding fields of lighting control, audio and audio visual systems and offers a wide range of interesting work in an attractive environment and excellent conditions of employment.

Apply: Personnel Director, Electrosonic Ltd, 815 Woolwich Road, Charlton SE7 8LT. Tel: 01-855 1101
APPOINTMENTS

Glasgow, G4
The Holders of the Director, BMA Applications, with Starting Salary: £2046–£2562 + £312 London Weighting + current Threshold payments. Applications giving details and names of two referees to Senior Administrative Assistant.

Colour CCTV Engineer

We are looking for a first-class electro-mechanical service engineer to maintain television, film, electronic and other equipment including Shibaden 1212 colour cameras, IVB colour VTRs, Sony U-matic recorders, etc. This equipment is stored in our theatre, studio, cinema and boardrooms in Shell Centre and at our Conference Centre at Teddington. The job also involves production of programmes, organisation of staff, design and building of prototypes of mechanical, optical and electronic units in our busy AV Centre. Starting salary would be negotiable, dependent upon qualification, experience: 5 day week, contributory Pension Fund, free 3-course lunches, 4 weeks annual holiday. Sports and social facilities in the building, including squash, badminton, swimming.

Telephone or write for an application form to: Shell International Petroleum Company Limited, I.P. 112, Shell Centre, London SW 7 NA. 01-034 2828.

GLASGOW COLLEGE OF TECHNOLOGY

PART-TIME B.Sc. DEGREE (C.N.A.A.)
IN ELECTRICAL ENGINEERING
with choices in Power or Electronic subjects

Holders of a good HNC or HND in Electrical/Electronic Engineering or related disciplines may be eligible to enter the above course commencing 18th August, 1975.

If you are interested in the above course, write or telephone now for application form and further details to:
The Academic Registrar, Glasgow College of Technology, North Hanover Place, Glasgow, G4 0BA. (Telephone: 041-332 7090).

BRITISH MEDICAL ASSOCIATION
DEPARTMENT OF AUDIO/VISUAL COMMUNICATION

Electronics Officer

Up to £3008 (increase is currently being negotiated) + £410 London Weighting

With lively interest in Audio/Visual aids for education.
Duties will include liaison with medical teachers and providing information and advice on a wide range of equipment, supervising a workshop for repair and maintenance of closed circuit television and audio equipment for research purposes; supervising the design and construction of prototype equipment and an audio cassette duplication service; undertaking sound and television recordings in a small studio.

Applicants should have at least 5 years’ experience in the educational uses of audio/visual aids.

Starting salary according to qualifications and experience.
Applications, with a full curriculum vitae and the names of two referees, to the Director, BMA House, Tavistock Square, London WC1H 9JP, not later than 11th July.
TEST ENGINEERS
S. LONDON
UP TO £2,800 p.a.

Dolby Laboratories is a young, go-ahead company with a world wide reputation for their audio noise reduction system. Test Engineers with a good understanding of basic circuits are required to test and trouble shoot professional audio P.C.B.s and equipment. This is interesting and well paid work. We give over four weeks’ holiday per annum.

Write or phone:
Mr. C. Keys
Dolby Laboratories Inc.,
346 Clapham Road,
London, S.W.9
Tel. 01-720 1111

Opportunities for Electronics Engineers
To change to wider fields of electronics – join the EMI Service Team at Hayes.

Vacancies exist on repair and calibration of a wide range of electronic test gear including oscilloscopes, DVMs, pulse generators, power supplies etc.

Also
Servicing and commissioning closed circuit television equipment including cameras, VTRs, Monitors etc.

Applicants should have at least 5 years practical experience. These positions offer varied and interesting work. Attractive starting salaries, subsidised lunches, 4 weeks holiday and excellent sick pay and pension schemes.

For further details telephone or write to - M. Ford, 01-573 3888, Ext. 2167, EMI Service, 254 Blyth Road, Hayes, Middlesex.

VIDEO ENGINEER
Good all-rounder required by London’s liveliest video dealers.

Due to continuing expansion in all departments we need another good engineer. Our team is small but very good so we are seeking someone with sound practical knowledge and wide product experience.

In return we are prepared to pay a more than generous wage with fringe benefits.

Please contact in confidence:
Peter Ellis, Technical Manager
TELETEAPE VIDEO
76 Brewer Street
London, W1R 3PH
Telephone 743 1319 or 434 1267

The international music, electronics and leisure Group.

UNIVERSITY OF SOUTHAMPTON

Research Fellow in Optical Communications

Applications are invited for a Research Fellow or Assistant to join an active group working on Optical Fibre Communications in the Department of Electronics. The work involves the use of fibres in multi-access communication systems including the development of couplers and junctions. The normal qualifications required are a Ph.D. or equivalent research experience although applicants with other qualifications will be considered. Some knowledge of electronics, communications or opto-electronics is desirable. Salary will depend on experience and qualifications and will be within the range £1,800 to £2,400 per annum (under review) plus free paying.

Applications, including curriculum vitae and the names of two referees, should be sent to Mr. D. A. Copland, The University, Southampton SO9 5NH, quoting reference 383/R/WW

HOME OFFICE, London

Directorate of Radio Technology
Telecommunications Officers (£4,190-£4,620)

to be responsible for the study of radio propagation matters over the whole of the radio frequency spectrum (10kHz-275 GHz) and for the forward planning, management and regulation of frequency bands allocated to broadcasting, fixed, maritime and land mobile, and space services.

Duties also include preparing specifications and type-approval of equipment for fixed and mobile services, application of computer techniques to frequency assignment problems, development of equipment for the location and suppression of radio interference, technical advice on all aspects of licensing of radio services and advice in connection with the international radio monitoring service.

Candidates (aged at least 25) must have ONC in Engineering (with a pass in Electrical Engineering 'A') or in Applied Physics or an equivalent qualification. In addition they must have had at least 7 years' experience of skilled work on radio, radar or other electronic work.

Salary: starting at £4,190, rises to £4,625. Good promotion prospects. Non-contributory pension scheme.

For further details and an application form (to be returned by 10 July, 1975) write to Civil Service Commission, Alencon Link, Basingstoke, Hants. RG21 1JB, or telephone Basingstoke (0256) 68551 (answering service operates outside office hours) or London 01-839 1992 (24 hour answering service).

Please quote ref. T/9017
APPOINTMENTS

UNIVERSITY OF SURREY

ELECTRONIC ENGINEER

Applications are invited for the above position in the Electronic Workshop of the Psychics Department. The person appointed will work, together with two other members of the technical staff, under the general direction of a Chief Technician.

Applicants should have a good electronics background, a sound theoretical knowledge and should have experience in the development and construction of computer interfacing and be familiar with nucleonic instrumentation.

Qualification: HNC or equivalent. Salary scale: £2,844 - £3,450.

For further details and application forms please apply to the Staff Officer, University of Surrey, Guildford, Surrey GU2 5XH or Tel: Guildford 71281, Ext. 452.

4654

NORTHAMPTON COLLEGE OF TECHNOLOGY

DEPARTMENT OF ENGINEERING

LECTURER GRADE I IN ELECTRICAL ENGINEERING

Applicants should have had previous experience of light current/electronic work and hold an H.N.C. or a final C.G.I.T. certificate with electronic subjects. Previous teaching is not essential, although desirable.

Duties will commence on 1st September 1975. Salary scale (under review) £1869 - £1363 per annum plus threshold payment.

On 1st September 1975 the Northampton Colleges of Technology, Art and Education will amalgamate to form a new College of Higher Education (Nene College).

Further particulars and application forms can be obtained from the Chief Administrative Officer, Northampton College of Technology, St. George's Avenue, Northampton NN2 6JB. Telephone (0604) 713505.

4769

PRODUCTION MANAGER

for small quartz crystal manufacturing plant

in NEW ZEALAND

An opportunity exists for a Production Manager familiar with all aspects of quartz crystal manufacturing for the communications market. Past experience should encompass grinding, vacuum plating and finishing to frequency.

The company, Hatfield Crystals Ltd., has recently entered the field of quartz crystal filter manufacture thus, although not an essential, it would be useful if the applicant has knowledge of quartz crystal design, particularly monolithic crystal filters in the 10.7 MHz band.

The successful applicant must be prepared to reside permanently in New Zealand and will be sponsored through the Migration Department of the New Zealand High Commission. The company is located at Napier, North Island, in a temperate climate not unlike the South of France. An attractive salary together with the usual fringe benefits will be offered.

Applicants to write in the first instance to:

The Managing Director
HATFIELD INSTRUMENTS LTD.
Burrington Way
Plymouth, PL5 3LZ
Devon

4646

TEST ENGINEERS

We have vacancies for Test Engineers to fault find and test a wide variety of quality control equipment, with experience of working on chemical, gas and oil analysis essential.

These positions would be ideal for ex-service personnel with relevant experience.

Good rates of pay, 4 weeks holiday, pension and sick pay schemes.

Ring Sylvia Borra 01-692 1271 Ext 393
or write to her at
The Personnel Department

CENTURY WORKS, CONNINGTON ROAD
Lewisham, London SE13 7LN

NEW GRAM AND SOUND EQUIPMENT

GLASGOW. Hi Fi, Cassette Decks, Tape Recorders, Video Equipment, always available we buy, sell and exchange for HI-FI sets and photographic equipment. VICTOR MUIRIS Audio Visual Ltd, 36 Argyle Street, Glasgow, G1, 8/10 Glassford Street, Glasgow, G2, 31 Sauchiehall Street, Tel: 041-221 8508.

411
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**SITUATIONS VACANT**

**HI-FI AUDIO ENGINEERS.** We require experienced Junior and Seniors and will pay top rates to get them. Tell us about your abilities. 01-427 4657.

**ELECTRONICS TECHNICIAN** required in Department of Psychology, University of Reading. Should have or be completing final C & G in electronic servicing or equivalent. Salary circa £2440-£2650 p.a. (grade 5). Apply with names of 2 referees and full details, quoting Ref. 722, to Assistant Director (Personnel), University of Reading, Whiteknights, Reading RG6 2AH.

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**TAPE RECORDING ETC.**

**RECORDS MADE TO ORDER**

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PO Box 3, Hawk Street, Carnforth, Lancs. Tel. 2273.

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**ELECTRONIC CRAFTSMEN**

*Is your present job routine and uninteresting?*

We are a research establishment and our craftsmen are engaged on a wide variety of work in the fields of prototype and small batch wiring and assembly, test and inspection, maintenance fault finding and repair. Why not join us and enjoy working in first-class conditions in the country?

Earnings are good and our rates of pay are currently under review. We can offer good housing at low rent (for applicants who live outside the radius of our Assisted Travel Area) together with 3 weeks’ paid holiday with holiday bonus, free pension and excellent sick benefit scheme.

Applicants who should have served a recognised apprenticeship or have had equivalent training together with experience in one of the fields detailed should phone Tadley 4111 (STD 073-56 4111) Ext. 5230, or write to:

**INDUSTRIAL RECRUITMENT OFFICER**

(PA/87/WW) PROCUREMENT EXECUTIVE

MINISTRY OF DEFENCE

AWRE ALDERMaston

READING, BERKS.

RG7 4PR

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**SONY**

**Trainee V.T.R. Service Engineer**

Due to the continued growth in the sales of our products we have a vacancy for a qualified V.T.R. Engineer in our Central Service Division, Ascot Road, Bedfont, Middlesex. Ideally we require an engineer with previous experience of servicing Video products, but we are prepared to train engineers who have experience of televisions and tape-recorders.

The duties will involve the servicing of our wide range of Video products, including recorders, cameras and microphones.

We are offering an attractive salary which will be based on qualifications and experience and generous staff discounts on all of our products.

Please apply to: The Personnel Officer, Sony (UK) Ltd., Pyrene House, Sunbury-on-Thames, Middlesex. Tel: Sunbury 87644.

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**MINISTRY OF DEFENCE (ARMY DEPARTMENT)**

A vacancy exists for a **CIVILIAN INSTRUCTIONAL OFFICER**

Grade III (Telecommunications), TV Servicing at 2 Resettlement Centre, Aldershot.

The post is open to men and women fully skilled and experienced in Televisions Servicing. Appropriate ONC/C Certificate or equivalent qualification desirable. Selection is by test and interview. Starting salary £2825 to £3209 (at age 28 or over), rising to £3925. Non-contributory pensionable employment. Write for application forms to:

Commandant, 2 Resettlement Centre, Gallway Road, Aldershot, Hants. GU11 2DG

Closing date: 16th July, 1975.
Radio Operators. How to see more of your wife without losing sight of the sea.

APPRENTICES

Radio Operators.

How to see more of your wife without losing sight of the sea.

Join the Post Office Maritime Service. We have openings for Radio Operators at several of our coastal stations. The work is just as interesting, just as rewarding as aboard ship, but you get home to see your wife and family more often. You need a United Kingdom General or First Class Certificate in Radiocommunications, or an equivalent certificate issued by a Commonwealth Administration or the Irish Republic.

Starting pay for a man of 25 or over is £2,270, plus cost of living allowance with further annual increases after that. Though we're happy to take people from 19 up.

In addition to your basic salary, you'll get an average allowance of £450 a year for shift duties and there are opportunities for overtime. Other benefits include a good pension scheme, sick pay and prospects of promotion to Senior Management.

For more information, write to: ETE Maritime Radio Services Division (R/8/A), ET 17.1.1.2., Room 643, Union House, St. Martins-le-Grand, London, EC1A 1AS.

Radio Operators. How to see more of your wife without losing sight of the sea.

Join the Post Office Maritime Service. We have openings for Radio Operators at several of our coastal stations. The work is just as interesting, just as rewarding as aboard ship, but you get home to see your wife and family more often. You need a United Kingdom General or First Class Certificate in Radiocommunications, or an equivalent certificate issued by a Commonwealth Administration or the Irish Republic.

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For more information, write to: ETE Maritime Radio Services Division (R/8/A), ET 17.1.1.2., Room 643, Union House, St. Martins-le-Grand, London, EC1A 1AS.
**HANDI-PACKS**
T.V. SOCKETS (metal-type). 5 for 50p.
TO3 TRANSISTOR INSULATOR SETS. 10 for 50p.
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PC BOARD POURER HANDLES, small only. £2.00.
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Mini 4 pole way. 2 for 50p.
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