Wireless World

Walter Cocking retires

For over 40 years the name of W. T. Cocking has appeared in the journal, for it was in September 1929 that he contributed his first article — on the design of a.c. mains receivers. He was then a free-lance radio journalist and consulting engineer and it was not until 1936 that he joined the editorial staff. His great contribution has undoubtedly been his constructional articles. The lucidity of his explanations was born of his practical experience, for in the majority of cases he did a major part of the constructional work himself. Readers may recall his peroration to his article 'Milestones in Receiver Evolution' in our 60th birthday issue (April 1971). He wrote 'The old saying, "an ounce of practice is worth a ton of theory" is still true. It is not that theory is unnecessary. It is more necessary than ever. It needs the practice, however, to drive it home and make one realize to the full what it means'. This epitomizes Cocking's whole approach to his journalistic work — theory and practice must go hand in hand. One recalls, for instance, the number of jigs he produced and on which he wound the deflector and focusing coils for the post-war monochrome television receiver (January-December 1947) before finally deciding on the one to describe for readers to wind their own coils. One correspondent writing congratulating us on our 60th birthday issue wrote 'Cocking's article underlines the debt we owe him on the home-construction side of the journal's activities'.

It would be invidious to try and pick out his most outstanding designs — each was outstanding in some respect when produced — but mention must be made of the Monodial receiver (1932) which was one of the first examples of a superhet using a single tuning control. In the audio field there was, of course, the 'Push-pull Quality Amplifier' (1934) which older readers will recall used two PX4 triodes in push-pull driven with RC coupling from push-pull MHL4 triodes. His more recent constructional projects include the W.W. Colour Television Receiver (1968/1969), a television wobbulator and the recently completed dual-trace oscilloscope unit. In applauding his practical contributions we must not, of course, lose sight of his theoretical articles not all of which bore his own name, he sometimes used the pen name W. Tüsting. Incidentally, much of the theory on which he has written over the years has resulted from the mathematical investigations he carried out preparatory to a constructional project. He is also the author of several books such as 'Wireless servicing manual' and 'Television receiving equipment'.

During the second world war Walter Cocking served in the R.A.O.C. and later R.E.M.E. and attained the rank of major. From 1942 to 1945 he was attached to the Ministry of Supply, as deputy assistant director, scientific research.

On demobilization he was appointed editor of our sister journal Wireless Engineer, which in 1962 became Industrial Electronics. In 1965 he was appointed editor-in-chief of both I.E., which ceased publication in 1969, and W.W.

We wish Walter Cocking, who was 65 on January 7th, a long and happy retirement and hope that we shall occasionally have the opportunity of including contributions from him.
Four-channel Stereo

2—Some commercial quadraphonic matrix systems

by Geoffrey Shorter*

This article does not discuss the virtues of multichannel reproduction for two- or three-dimensional directionality, nor the appropriateness of choosing four channels as opposed to any other number. Given that the record industry has decided that the next development in audio is four-channel reproduction, what is discussed is how this can be achieved using existing two-channel media by matrixing or coding and without recourse to multiplexing or modulated carriers. In the long term, of course, it could be that a carrier technique will replace the matrix techniques, but it is too early to say.

First, it’s important to distinguish between the ‘surround-sound’ systems and the true coding systems. As mentioned in last month’s article, the first type takes the signals from an ordinary stereo record or broadcast and derives signals for two additional speakers, usually placed behind the listener. In manufacturers’ literature this is frequently referred to as composing, quadrailing or synthesizing a sound field.

The simplest way is to put difference signals in the rear speakers with a 180° phase difference between them, but there are many variants, as illustrated in the table* published last month.

Now, using such methods of getting four speaker signals from the two stereo signals as a basis, it is possible to achieve directionality of sound in addition to the normal stereo stage width by ‘coding’ the two stereo signals, in the way suggested in last month’s article. Of those ‘surround-sound’ systems listed only Electro-Voice, Dynaco and Sansui promote a coding system. CBS and Nippon Columbia also have coding techniques. (Those wishing to experiment with surround sound could use the circuits of Fig. 5 in last issue with resistor values chosen to give desired amounts of crosstalk.)

One of the first commercial systems for achieving added ambience and directionality was proposed by Dynaco. This is perhaps the simplest of all systems because it needs only two (existing) amplifiers. Designating signals intended for localization in the left, right, front and back speakers by L, R, F and B the signals L+F+B and R+F−B are used as the two transmitted or recorded signals. In playback through the ‘speaker matrix’ they are fed to left and right speakers, and the front and rear speakers receive 2F+L+R and 2B−L−R respectively (Fig. 1(a)).

Table 1 summarizes data on the system. As with the matrix discussed last month, localization is poor. Mono rendition is also poor (back signals are not reproduced and front signals emphasized by 6dB).

Apart from the disadvantages of mislocalization of sounds and the poor mono compatibility in the Dynaco system, the diamond speaker pattern also requires a blend between the two channels to counteract crosstalk in the speaker network, reducing its simplicity. Also, there is an upset to the amplifier load, which can vary over a 9:1 ratio depending on sound position.

Some improvement of the original diamond set-up can be had by a modification suggested by Tappan, where the code and decode coefficients are as shown in the Table. This results in better performance but it’s still far from perfect. Mono compatibility is better, with the front source enhanced by only 3dB.

Latest technique by Dynaco involves a square speaker set-up (Fig. 1(b)) with no blend added between the two inputs, but with reduced signal levels in the rear speakers (counteracted by listening closer to the rear speakers). Mislocalization still occurs in the rear sectors. In the two-channel mode there is crosstalk between ‘front’ signals of 12dB and in the mono mode the rear signals are 8dB down on the front signals. Data are summarized in Table 2.

In this table mathematical matrices are used to avoid writing lots of similar equations. (Use of the word matrix in quadraphony does not arise from this mathematical sense but from the electrical sense, as indicated last issue.)

For those not familiar with matrix algebra, the matrices in this article can be regarded merely as a convenience. Take the coding equations derived last month

\[
L = aL_F + bR_F + aL_B - bR_B \\
R = bL_F + aR_F - bL_B + aR_B
\]

where \(a=0.924\) and \(b=0.382\). Using matrix notation these can be written

\[
\begin{bmatrix}
L \\
R
\end{bmatrix} =
\begin{bmatrix}
0.92 & 0.38 & 0.92 - 0.38 \\
0.38 & 0.92 - 0.38 & 0.92
\end{bmatrix}
\begin{bmatrix}
L_F \\
R_F \\
L_B \\
R_B
\end{bmatrix}
\]

(Figures rounded off.) Similarly the decoding equations

\[
L_F = aL + bR \\
R_F = bL + aR \\
L_B = aL - bR \\
R_B = -bL + aR
\]

can be written

\[
\begin{bmatrix}
L_F \\
R_F \\
L_B \\
R_B
\end{bmatrix} =
\begin{bmatrix}
0.92 & 0.38 \\
0.38 & 0.92 \\
0.92 - 0.38 \\
-0.38 & 0.92
\end{bmatrix}
\begin{bmatrix}
L \\
R
\end{bmatrix}
\]

where the same coefficients are used in decoding.

*Assistant editor, Wireless World

*To which we can now add data for the Toshiba and Onkyo methods. For Toshiba, front and back crosstalk, respectively, is none and 0dB; 15dB and 0dB; or none and 13dB (switchable). For Onkyo, the crosstalk is none and about 0dB. (Makers claim a 90° phase difference between the two rear speakers, but it is not clear that this is achieved from the circuit diagram.) Scott, Fisher and Lafayette also have ‘exclusive’ surround-sound circuitry in their equipment.
Combining these by the rules of matrix algebra gives

\[
\begin{bmatrix}
    L_F & R_F \\
    L_B & R_B \\
\end{bmatrix}
\begin{bmatrix}
    1 & 0.71 & 0.71 & 0 \\
    0.71 & 1 & 0 & 0.71 \\
    0 & 1 & 0 & 0.71 \\
    0 & 0 & 0.71 & 1 \\
\end{bmatrix}
\]

equivalent to the equations for final outputs on p.4 of last issue.

Now in the discrete case the equivalent matrix would be

\[
\begin{bmatrix}
    1 & 0 & 0 & 0 \\
    0 & 1 & 0 & 0 \\
    0 & 0 & 1 & 0 \\
    0 & 0 & 0 & 1 \\
\end{bmatrix}
\]

and any terms in a matrix in addition to these in this unit matrix represent unwanted terms due to crosstalk.

Thus, without knowing anything about matrix algebra, it is easy to deduce appropriate output equations. In the 2 by 4 matrices under the heading 'code' the two rows represent the two encoded signals, say L and R, the numbers being coefficients of \( L_F, R_F, L_B \) and \( R_B \) in that order. In the 4 by 2 matrices under the decoder heading, the two numbers in each row are coefficients applied to \( L \) and \( R \), each row representing equations for \( L_F', R_F', L_B' \) and \( R_B' \). (Primes in these terms distinguish output from input signals. In the 4 by 4 matrices numbers in the rows are coefficients of \( L_F, R_F, L_B \) and \( R_B \) in the equations for \( L_F' \) etc. representing signals fed to each speaker. (A column gives the speaker signals resulting from a single input to the matrix.)

As well as being able to deduce directional accuracy from the equations for the four outputs in Table 2 (Fig. 7 last month illustrates the case of the Scheiber matrix), one can also deduce what the outputs at each speaker will be if an ordinary stereo source is played through the matrix ('surround-sound' or 'synthesizer' mode).

But it is of more interest to know of the compatibility of coded sources with existing two-channel and mono reproducers, so the outputs in these modes are indicated in the Table.

Other, related, systems involving phase shifts of only 180° — the symmetrical (Scheiber) matrix derived last issue, and the Electro-Voice system — also have the drawback of mislocalization and clear back images are not possible because of the antiphase relation in the back speakers.

In the two-channel mode with the Scheiber matrix, crosstalk between left and right front signals is 7.7 dB, and in the mono mode the back signals are 7.7 dB down. With the E-V matrix, left and right front crosstalk is 10 dB in the two-channel mode, and in mono back signals are 8.3 dB down. For decoding E-V coded material, Zenith have a decoder with coefficients different from the usual E-V coefficients. This gives an improved left-right front crosstalk in the four-channel mode — 10 dB as opposed to 6.5 dB.

A matrix proposed by Tappan, but not taken up commercially, improves the two-channel performance of the Scheiber array by using coefficients shown in the table to give zero left-right crosstalk.

but with back signals emphasized by 3 dB, and in mono back signals are 7.8 dB down. One way of overcoming the severe mono incompatibility, proposed by Scheiber, is to introduce a differential phase shift between the two coded signals of 90° (say \( +45° \) on one channel and \( -45° \) on the other). On decoding, the two signals are treated with a differential phase shift in the opposite sense, thus cancelling the phase shifts. It is claimed that the 90° phase difference doesn't matter in two-channel playback. In mono playback the corresponding signals in the two channels add to give resultants that are all 0.707 dB down.

The Scheiber system does not seem to have been produced commercially and we understand the rights to the system were acquired by Electro-Voice. Encoders and decoders for the Electro-Voice system are in existence, but it seems that Electro-Voice are advancing a new system — as from October last — probably the later matrix developed by Scheiber and announced at that time. We hope to give more details later. (It is for this latest system that Electro-Voice are offering a new i.e. chip which they claim will decode any existing matrix.)

A way of avoiding the mislocalization — illustrated last issue and which can be deduced from the data in the Table — in the simple matrix technique has been adopted by Sansui in their QS matrix system. The basic technique used by Sansui is the same as the matrix derived in the previous article. But before the four inputs are combined in the coding matrix (Fig. 4 in the previous article; the circuit could be an encoding circuit performing the equivalent of the decoding matrix of Fig. 5) the two rear signals are treated to phase shifts of 90° and —90° relative to the front channels. Thus the rear two signals are still 180° out of phase, but no cancellation occurs within each equation for the coded signals with coherent signals overcoming one source of mislocalization.

As an antiphase relation between the two rear signals prevents localization in the back, 90° phase difference circuits are used in the decoder after the matrix (Fig. 5 last month) in the converse way to encoding i.e. —90° shift in the \( L_B \) channel relative to the front, and +90° shift in \( R_B \) channel relative to the front. Decoded outputs are shown in the table and of course some signals are in quadrature. But Sansui claim that localization is unaffected by the phase shifted components.

Crosstalk from any speaker to its neighbours is —3 dB, but to the opposite speaker it is zero, and centre front-to-back side-to-side crosstalk is 7.7 dB as for the basic matrix without phase shifters. Also as with the basic matrix, 'front' left-right crosstalk in the two-channel mode is 7.7 dB (with of course the phase-shifted terms still providing an antiphase output thus giving 'back' images apparently from outside the speakers). In the mono mode the back signals will emerge shifted by +90° and —90° and —7.7 dB relative to the front signals.

One point about the phase-shift circuits used by Sansui is that ideally they should give the required phase shift over a wide band of frequencies, and while it's possible to get a 90° phase shift at any spot frequency it's impossible to get such an absolute shift over a range of frequencies. What is done is to provide a 90° phase difference between two lines using an all-pass filter. By cascading networks of the kind shown in Fig. 2(a) in each of the two lines and adjusting corresponding filters to have a 90° phase difference, a wide phase bandwidth can be approximated. What effectively is being done is to give the front channels a phase shift of \( \phi \), at a certain frequency, and the back channels a phase shift of \( -\phi \) at that frequency. At another frequency the front phase might be \( \phi \), but so long as the rear phase shifts are \( -\phi \) and \( +\phi \), the 90° difference is maintained (Fig. 3).

In the decoder circuit however such a cascade circuit in each channel is costly.
Table 2  Quadraphonic matrix systems

<table>
<thead>
<tr>
<th>Code</th>
<th>Decode</th>
<th>Outputs with coded material</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Four-channel</td>
</tr>
<tr>
<td>DYNACO 'square' array</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[1.0 0.25 1.0 -0.5]</td>
<td>[1.0 0]</td>
<td>[1.0 0.25 1.0 -0.5]</td>
</tr>
<tr>
<td>[0.25 1.0 -0.5 1.0]</td>
<td>[0.25 1.0 -0.5 1.0]</td>
<td>[1.0 0.55 -0.3 0.82 -0.88]</td>
</tr>
<tr>
<td>ELECTRO-VOICE original system</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[1.0 0.3 1.0 -0.5]</td>
<td>[1.0 0.2]</td>
<td>[1.0 0.05 0.9 -0.3]</td>
</tr>
<tr>
<td>[0.3 1.0 -0.5 1.0]</td>
<td>[0.3 1.0 -0.5 1.0]</td>
<td>[1.0 0.59 -0.38 1.06 -0.66]</td>
</tr>
<tr>
<td>ZENITH decode</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[1.0 0.25 1.0 -0.5]</td>
<td>[1.0 0.25 1.0 -0.5]</td>
<td>[1.0 0.55 -0.3 0.82 -0.88]</td>
</tr>
<tr>
<td>SCHEIBER original system</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[0.92 0.38 -0.38 -0.38]</td>
<td>[0.92 0.38 -0.38 -0.38]</td>
<td>[1.0 0.71 0.0 0.71]</td>
</tr>
<tr>
<td>[0.38 0.92 -0.38 0.92]</td>
<td>[0.38 0.92 -0.38 0.92]</td>
<td>[1.0 0.71 0.0 0.71]</td>
</tr>
<tr>
<td>TAPPAN</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[0.71 0.71 -0.71 1.0]</td>
<td>[1.41 1.0]</td>
<td>[1.41 1.0]</td>
</tr>
<tr>
<td>[0.0 0.71 -0.71 1.0]</td>
<td>[1.0 1.0]</td>
<td>[1.0 1.0]</td>
</tr>
<tr>
<td>SANSUI</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[0.92 0.38 -j0.92 j0.38]</td>
<td>[0.92 0.38 -j0.92 j0.38]</td>
<td>[1.0 0.71 j0.71 0]</td>
</tr>
<tr>
<td>[0.38 0.92 -j0.38 -j0.92]</td>
<td>[0.38 0.92 -j0.38 -j0.92]</td>
<td>[j0.71 0 1.0 j0.71]</td>
</tr>
<tr>
<td>CBS</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[1.0 0 -j0.71 0.71]</td>
<td>[1.0 0]</td>
<td>[1.0 j0.71 -j0.71 0]</td>
</tr>
<tr>
<td>[1.0 j0.71 -j0.71 0]</td>
<td>[1.0 j0.71 -j0.71 0]</td>
<td>[j0.71 0 1.0 j0.71]</td>
</tr>
<tr>
<td>Alternative forms of decode</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[1.0 0.25 -0.25 -0.25]</td>
<td>[1.0 j0.71 0.71 -j0.71]</td>
<td>[j0.71(1 + 0.25) j0.71(1 + 0.25)]</td>
</tr>
<tr>
<td>[0.71(1 + 0.25)]</td>
<td>[0.71(1 + 0.25)]</td>
<td>[j0.71(1 + 0.25) j0.71(1 + 0.25)]</td>
</tr>
</tbody>
</table>

Notes:
- Most numbers rounded off to two decimal places
- Equations as under 'code'
- * Equations as under 'code'

Diagram: Wireless World, February 1972

Wireless World, February 1972
and Sansui use two simple phase-shift circuits — Fig. 2(a) — whose 90° frequencies are about an octave apart, which means that the shift is frequency-dependent, reducing to nearly zero at low audio frequencies and increasing to 180° at high audio frequencies.

Listening tests by Sansui suggested the frequency of 90° phase shift should be around a few hundred Hz. Actually this 90° phase relation is constantly varied so that sounds coming from behind have an unpredictable phase — this is done to simulate more closely reverberation on reflected sounds which in a live sound field would have varying phase relations. Sansui claim an increased 'presence' as a result of this and also a better dynamic range of pulsed tones. Listening tests decided the rate and depth of phase modulation so that any direct sound from the back is unaffected. The phase modulator circuit is shown in Fig. 4, where the feed to the lamp is taken from 8 and 9Hz oscillators (additionally producing a 1Hz beat) and from the audio signal, allowing left and right channels to be modulated differently, and giving a 'random' element to the modulation.

For completeness in our table, data for the CBS SQ system® is included. As a matter of interest the SQ system uses similar cascaded networks for its professional coders and decoders. With two such chains, a phase difference of 90° can be maintained to within ±2° from 20Hz to 20kHz. But in the simplest decoders four single-stage second-order networks are used — Fig. 2(b). By proper staggering, the 90° phase differential is maintained to ±10° from 100Hz to 10kHz (giving a 20dB minimum isolation between back channels).

In reference 6 we showed a simplified decoder diagram for the SQ system. Actually the terms formed with j are, of course, phase difference circuits of the kind described, and not absolute 90° phase shifters as some readers may have been led to believe. (The j notation was merely a convenience). An alternative SQ decoder design has a degree of blending with 25% of both front channels put into the other, and similarly for the back channels. Another variant uses a 10% blend between front channels and a 40% blend between back channels. (These are intended for use without the 'logic' circuits which can be used to improve 'discreteness' for simple input signals.)

With the gain of the back channels reduced by 1dB the last mentioned matrix gives a 20dB front crosstalk; an 8dB-rear crosstalk and a centre front-to-back differential of 3dB. This matrix is being used by CBS Masterworks (U.S.A.) and by other manufacturers.

Sony, who are making decoders for the SQ system outside the U.S.A. (non-exclusively), use a front-back 'logic' circuit in their SQ1000 decoder. In the SQ system, a centre-front signal has the undesired (back) outputs in antiphase, and also a centre back signal produces the undesired (front) signals in antiphase. This antiphase condition is detected and a variable gain network is used to enhance front-to-back centre from 3 to 8dB. This decoder has phase-shift networks giving 90° over 100Hz to 10kHz with ±5° accuracy. Crosstalk between back channels is 26dB, but diagonal front-to-back crosstalk remains at 3dB.

Sony are expected to bring out a decoder with a logic which detects and reduces gain for undesired (quadrature) signals accompanying corner signals.

A big claim for the CBS system is its compatibility in the two-channel mode, in which crosstalk between left and right front signals is zero. In mono, all four inputs are reproduced at the same level, but a centre back signal would be suppressed. CBS therefore advise recording engineers not to locate soloists at this position.

Another difference is that in SQ system with two-channel listening the rear signal images are placed in between the two speakers, whereas in the other systems the antiphase rear signals give rise to images outside the two speakers.

As well as the commercial systems discussed, Nippon Columbia in Japan will be using a new kind of matrix from spring onwards. Also the new form of the Electro-Voice matrix will be around, and quite possibly other systems may be proposed. We hope to discuss these in a future article.

References and notes

1. A recent study relates overall subjective effectiveness to the number of discrete channels and suggests it might be possible to design a four-channel system with almost the same effect as eight channels. It shows additional effectiveness decreases as the number of channels increases. However, this study did not relate to the matrixed case. T. Nakayama, et al "Subjective assessment of multichannel reproduction" J. Audio Eng. Soc. vol. 19 1971 pp.744-51.


News of the Month

Electronically generated colour test card
Periodically during January and February, television transmitters in the London area will radiate a different form of test signal during trade test periods. Instead of the normal Test Card F, an electronically generated colour picture signal will be used. This incorporates most of the facilities provided by Test Card F, except for the picture contained in the central circle. The experimental periods are:
- Jan. 24 to 28 BBC-1 transmitters
- Jan. 31 to Feb. 4 BBC-2 transmitters
- Feb. 7 to 11 I.T.A. transmitters
In the case of BBC transmitters, the present Test Cards are generated centrally and distributed to the transmitters by the normal distribution networks. The use of the electronically generated signals will make it possible to maintain a colour signal at those times when the distribution circuits are not available for this purpose.

European conference on high fidelity
The President of the Federation of British Audio, Mr. K. H. Williman, has just returned from a European Conference organized by the German High Fidelity Institute. The objects of the conference were twofold. First, to discuss the formation of a European High Fidelity Institute, intended to cover technical and legal matters and European standards for high-fidelity equipment; and secondly, to discuss the possibility of holding a high-fidelity congress at the new Dusseldorf exhibition site during October 1972. Taking part in the Conference, as well as Mr. Williman, were Monsieur Boissinot (director general of the Society for Information for Sciences and Arts (France)), Ing. Hans K. Friedl (Austrian Institute for High Fidelity), Dr. Boggaerts (Belgian High Fidelity Society), and Mr. Bö Rydin (Swedish High Fidelity Institute). A further meeting is planned.

Festival Hall acoustics exhibit
The Building Research Station has presented to the Science Museum, London, an exhibit which symbolizes the work that it carries out in the field of architectural acoustics and in particular the acoustics of concert halls. The main feature of the exhibit is a scale model of the Royal Festival Hall.

The model shows the Hall sectioned along the centre line, with the seats occupied by a full audience and the stage by an orchestra and choir. Pressing a button illuminates the space above the ceiling of the hall and draws attention to the Helmholtz resonators, microphones and loudspeakers of the ‘assisted resonance’ system, invented and developed at the Building Research Station, which in the Festival Hall has increased the reverberation at the lower frequencies.

The exhibit also includes a full-sized example of the resonators and microphones used in the assisted resonance system. In addition there are illustrations of other features of the construction of the Hall that were incorporated in the original design for the specific purpose of giving the desired acoustic properties.

The exhibit can be seen on the top floor of the Museum at the junction of galleries 61 and 64.

Investigating p.a.p.m.
Marconi Communication Systems have been awarded a study contract by the Comsat Corporation on behalf of Intelsat. Under the contract Marconi will investigate the properties of the pulse amplitude and phase modulation process (p.a.p.m.) using a mathematical model within a large computer.

A large saving in bandwidth for a given amount of information is promised as any radio link could be used to carry up to four-times the amount of information without any need to increase the bandwidth if p.a.p.m. is used.

An identical study is being carried out in America so that a comparison of results can be made and accuracy checked.

The modulation process impresses pulses onto a carrier by altering the carrier both in amplitude and phase.

Attitude sensor equipment
The Inertial System Department of Ferranti is to design, develop and manufacture attitude sensor equipment for the new all-British X4 earth satellite under a contract awarded by Hawker Siddeley Dynamics who are the prime contractors. It is planned that X4 will be launched early in 1974 by an American Scout rocket at the Western Test Range in America.

The X4 is the largest and most complex civil satellite to be undertaken by the U.K. so far. Its purpose is to prove in orbit systems and techniques which will be used on future applications satellites.

The attitude sensor will consist of three orthogonally mounted rate-integrating gyroscopes which will measure very small changes of the satellite’s angular movement in all three dimensions, thus providing the electrical signals necessary to stabilize the satellite in a chosen attitude in space. A fourth standby gyroscope will also be carried; this will be positioned at equal angles about the satellite’s three axes of movement so that it can ‘stand in’ for

This push-button data telephone with visual display has been developed by Siemens, Munich, as a result of a study of the ‘human factor’.

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any one of the other three gyroscopes in the event of a malfunction. Each gyroscope will have its own electronic module, providing electrical power for gyro motors and amplifiers, temperature controllers and switching circuits. The modules use a combination of discrete components and integrated circuits already proven in previous space programmes.

ESRO to build applications satellites

At a highly successful meeting in Paris in early December of the Council of the European Space Research Organization future plans for the organization were discussed and unanimously approved by the representatives of the member nations. A resolution allows ESRO to undertake application satellite programmes at an expenditure of not less than $70M* during the period 1974 to 1980. The programme could include aeronautical, meteorological and communications satellite projects.

*In addition not less than $27M* per year will be spent on scientific satellite projects together with sufficient funds to meet the requirements of the current scientific satellite projects TD-1A (due for launch February '72), HEOS-A2 (due for launch this month, January) and ESRO-4 (due for launch August '72). The budget ensures the future of the proposed satellites COS-B and GEOS and will allow ESRO to participate with America’s National Aeronautics and Space Administration and the U.K. in the small astronomical satellite project SAS-D.

The outcome of the meeting is good news for ESRO who will welcome trying their hand at applications satellites as their activities had previously been limited to scientific satellites; even the future of some of these was very much in doubt until the meeting.

High altitude temperature measurements

The Science Research Council has granted £141,710 to Oxford University for a new experiment to be flown in the NASA satellite Nimbus-F which is due to be launched in 1974. The grant will enable Dr. J. T. Houghton to produce a two-channel pressure modulator radiometer (so called because spectral selection in the instrument will be achieved by the use of a cell of carbon dioxide the pressure of which is modulated). The radiometer will make remote measurements of atmospheric temperature at 40 to 90 km above the earth. No measurements of temperature at these heights have ever been made before on such a large scale. It is known from rocket experiments that marked variations of temperature produced by atmospheric tides and gravity waves occur at these heights, but little is known of their influence on the lower atmosphere. It is also known that there are strong links between the motion of the lower atmosphere and events in the ionosphere as evidenced by radio propagation studies.

Largest coaxial cable

The largest short-wave transmitting station in Europe is currently being built for the German broadcasting organization ‘Deutsche Welle’ near Ettlingen in the Wurtach valley (31 miles south west of Ausburg). The station will have 12 500kW transmitters that can be connected to 74 aerials by means of a computer-controlled switching matrix. The aerials are to be erected in three lanes which extend as arms from a 300m circle. The main aerials will consist of webs of wire suspended from 53 towers which will vary between 35 and 125 metres in height, the mesh forming a radiating curtain on each side with a central reflector. The whole site will occupy an area of nine square kilometres.

About 100 tons of copper, 370 tons of aluminium and 30 tons of Teflon fluorocarbon resin will be used in the manufacture of the coaxial cables to connect the switching matrix to the aerials. The cable has the largest diameter of its type in the world, and is being manufactured on specially constructed machines by Felton & Guillaume Kabelwerke in Cologne and Kabel und Metallwerke in Hanover.

The cable consists of an inner conductor of radially corrugated copper tubing 99mm in diameter and 0.6mm thick, and an outer conductor of spirally corrugated aluminium with a diameter of 246mm. The corrugations give the cable enough flexibility to be wound on to drums. Every 250mm along the cable there are three spacers mounted at 120° to each other, manufactured with Du Pont Teflon, to ensure that the inner conductor stays concentric with the outer conductor. The spacers are held in place by a copper plated C-shaped steel ring.

Ceramic substrate test method

Ceramic sheet is being used in large quantities in the electronics industry and it would appear that methods of testing the strength of the material leave quite a lot to be desired as results are often none too accurate. The most popular test method in current use, a beam bending method, requires lath shaped specimens which must be either ground to certain dimensions and given a ‘standard’ surface finish or cut from larger specimens forming raw edges. Either procedure can mask the desired results.

The biaxial flexure test method developed at the American National Bureau of Standards, the ‘Piston-on-3-balls’ technique, involved supporting a plate of ceramic on three balls equidistant from its centre and uniformly loading a small, circular central area. The area of maximum tensile stress thus falls at the centre of the lower face of the plate and the strength should be independent of the condition of the edges of the plate. This method has been accurately analysed for small deflections (less than one-half the specimen thickness). The technique of supporting the specimen on three balls allows the use of a slightly warped specimen; thus no surface grinding or polishing is required. This makes it possible to test ceramic plates with an as-fired surface finish.

Eleven laboratories constructed their own test jigs according to the design developed by N.B.S. A total of eight types of polycrystalline alumina substrates, chosen to represent different strength levels, were supplied by five manufactur-
The coefficients determined for deviations within laboratories. Included, the variation between laboratories. The resulting straight line through the points is the regression line. The laboratory recorded the diameter of each specimen, its thickness adjacent to the origin of fracture, and the breaking load. The breaking strength values were calculated using a Poisson's ratio value of 0.23.

The results from eleven laboratories were examined by fitting least squares straight lines to their data and examining the resulting lines for indications of systematic differences between the laboratories. Small systematic differences are detectable but are so much less than the scatter within each laboratory that the most significant factor contributing to scatter in the data appears to be variations in the material. Examination of the data for all materials shows that, although there is always an increase in the standard deviation when the variation between laboratories is included, it is always small compared to the variation within laboratories.

The data was also analysed to determine the number of measurements required to give a reasonable measure of average strength. The coefficients of variation (the ratio of the standard deviation to the average strength) were determined for single and for averages of five and ten replicate measurements per laboratory. Results using a single measurement vary from 10.3 to 14.5%. The coefficients of variation for averages of five measurements range from 6.5 to 7.6%. For averages of ten measurements the range is 5.2 to 6.8%.

Laser beam displays
Commercial devices for modulating and deflecting laser beams, intended mainly for projected-light character displays, are now available in the U.S.A. Zenith Radio Corporation have produced in the past few months the M-40R modulator, the D-70R beam deflector and the Cascade 800 scanner — a device that receives the scanning beam from the D-70R and doubles its deflection angle. Demonstrations of the devices have been given in systems for directly displaying news items from news agency wires. The code signals that would normally operate a teleprinter would feed into a converter which produced signals for operating the modulating and deflecting units.

The principle of operation of all the devices is that the laser beam passes through a glass cell in which an optical diffraction grating, of compressions and rarefactions in the glass, is set up by ultrasonic waves travelling through the material from a transducer (70MHz frequency in the beam deflector). A detailed description of the physical mechanism, which is based on Bragg reflection as utilized in crystallography, was given in Wireless World, Nov. 1966, p.546, in a report on the early Zenith work in this field. Sufficient to say here that light modulation is achieved by varying the amplitude of the ultrasonic waves and beam deflection by varying their frequency.

In one demonstration, in a Chicago museum, the incoming signal from the news agency wire was coded and passed into a unit where seven frequencies were generated. These seven frequencies were sent to the M-40R modulator where, as ultra-sound, they split a laser beam into a fan of seven diverging beams. Each of these seven beams was turned on or off in accordance with the incoming signal. The beams were projected onto a slowly revolving phosphor-coated drum and each spot struck by a beam glowed for several seconds. The slow rotation of the drum allowed the words to be read as they came off the agency wire.

In another demonstration, in New York, the revolving phosphor drum was replaced by light-sensitive paper, and a deflector swept the seven-beam fan sideways, thus eliminating the need for any mechanical motion. The paper was dry-processed in a few seconds and reeled off the machine with the news ready to read.

Zenith say that a character generator of this type could also be used to print on microfilm. One practical application would be a readout device which would translate the output of a computer into readable copy and print it on microfilm. The microfilm could then be read in any conventional microfilm projection reader. It is claimed that such a character generator could print computer output information at speeds of about 100,000 characters per second — considerably faster than the present paper readout systems.

Sound field plotting by colour lights
Our cover shows a picture of sound field contours produced by a bass reflex enclosure, using a method described at the 1971 International Congress on Acoustics and briefly discussed in our report in the November issue. It involves moving a microphone, to which are attached coloured lamps, around the field. Each lamp is arranged to light when the sound pressure level is within a certain range. The technique is said to make sound field investigation in auditoria a relatively simple matter. Earlier related methods required the microphone to be moved at constant speed and relied on just one lamp with varying brightness. In the method used for our cover photograph, developed by Z. Barat and M. Vizcian at the Technical University of Budapest, the amplified microphone output is fed to a pen recorder with a sliding contact fitted to the pen. It switches on one of the lamps when the output is within a certain range, as determined by fixed contacts wired to the lamps. By watching the lights, contours of equal sound pressure can be followed by an operator holding a bar with the microphone and lights attached. A colour camera* operated in darkness with an open shutter will record the isobar. The five lamps used represented a 25-dB range in pressure level, each colour corresponding to different 5-dB ranges. Because the width of the fixed contacts was 2% of the 25-dB recorder scale, each contour line represents a minimum uncertainty of 0.5dB — more in practice because of time delays in lighting the filament lamps (48-V, 50mA lamps gave 40ms delay and 6-V, 40mA lamps gave 15ms delay. These delays mean rapid movement should be avoided).

For a laboratory instrument, the sliding contact idea certainly seems to be the cheapest if a recorder is to hand. But if a recorder is not available, electronic level switches are an obvious alternative, especially for a robust commercial instrument. Non-filament lamps would reduce the time delay problem.

Push-button telephones using m.o.s.t.s
An order worth over £100,000 has been placed by the Post Office with the Telephone Division of GEC-AEI Telecommunications for several thousand self-contained push-button telephones using metal-oxide semiconductor electronics. The telephones will be used for large-scale trials under public-service conditions. The telephone, known as the Keyphone 81, has been developed by GEC from their earlier Keyphone 75 to meet Post Office requirements. The Keyphone 81 telephone incorporated the DIA 51 keyset mechanism which is used to replace the rotary dial in all types of GEC push-button telephones. Over 30,000 of these mechanisms have already been delivered. The telephones do not require any conversion of the exchange switching equipment and can, therefore, be supplied as a direct replacement for most dial telephones, unlike the multifrequency telephone (Touch-tone) now being used in North America which does require the installation of interface equipment at the telephone exchange.

*Fitted with a wide-angle lens, an aperture of 5.6 and 14DIN/20ASA film for the cover photograph.
Universal Meter Amplifier

Design for a.c. and d.c. measurements, in two versions with different input impedances

by A. J. Ewins

The basic sensitivity of moving-coil meters, in terms of the minimum input voltage for full scale deflection (f.s.d.) and input impedance in ohms/volt, can be greatly increased by amplifying the input signal before applying it to the meter. In various v.v.m.s and transistorized multimeter designs, where a.c. and d.c. voltage and current signals are measured, this is normally achieved by separate amplifiers for the a.c. and d.c. signals. The most obvious reason for using separate amplifiers is probably the fact that a.c. signals need to be rectified before they can be measured by a d.c. meter. Also, there is a certain degree of incompatibility between the design requirements of the a.c. and d.c. amplifier. In the case of the d.c. amplifier it is not usually necessary to have a voltage gain greater than unity, provided the basic voltage sensitivity of the meter is adequate, it being only necessary to increase the input impedance. This can be readily achieved by the sort of circuit shown in Fig. 1, in which zero drift is not a problem provided the two transistors are suitably matched. The a.c. amplifier, however, needs to have high open-loop gain in order that a large amount of negative feedback may be applied to overcome the rectifying diode’s pedestal voltage and extreme non-linearity. Thus, an a.c./d.c. amplifier needs to have high open-loop gain and low zero drift, which are not always compatible.

However, the writer felt that if an amplifier could be constructed to drive a moving-coil meter, and have a frequency response from d.c. to some tens of kilohertz, the construction of a multimeter could be somewhat simplified and possibly some additional, advantageous, features could be introduced.

The amplifier designs presented in this article achieve this objective without degrading the performance of either the a.c. or d.c. sections. They do so with good d.c. stability, excellent scale linearity and a frequency response extending from d.c. to 100kHz.

The amplifiers have been designed around the Avo Model 9 meter movement, which has a basic current sensitivity, when shunted with a 10kΩ resistor to produce the required damping, of 50µA and a voltage sensitivity of 125mV.

The circuit design

Fig. 1. Type of amplifier, for d.c. only, suitable for increasing input impedance.

When this meter movement is used in conjunction with the two amplifiers described, its current sensitivity is increased to 500nA, or 5nA, with a practical voltage sensitivity of 100mV, making possible input impedances of 2MΩ/volt and 200 MΩ/volt. An additional advantage obtained by using this type of amplifier, when measuring d.c. signals, is that the meter always gives a ‘positive’ indication regardless of the polarity of the signal applied to the input terminals.

Fig. 2. Basic circuit from which amplifier designs were evolved. The two pairs of transistors are provided by the CA3018 integrated circuit.
volt line. The meter circuit is connected between $T_r$ collector and the zero volt line. Negative feedback is achieved by connecting the base of $T_r$, via $R_s$, to the junction of $R_3$, $R_{m_1}$, and $R_{m_2}$.

With no input signal to the amplifier the base of $T_r$ will be at some voltage slightly negative of the zero volt line due to the voltage dropped across $R_s$ by $T_r$'s input bias current. Provided that the differential amplifier is balanced, under open-circuit input (by $R_{C_D}$) and short-circuit input (by $R_{C_U}$) conditions, for minimum deflection of the meter, the base of $T_r$ must be at the same potential as the base of $T_r$ and the voltage dropped across $R_s$ must be the same as the voltage dropped across $R_3$. Thus the potential at the junction of $R_s$, $R_{m_1}$, and $R_{m_2}$ must be equal to zero volts. A portion of the input bias current to $T_r$ must therefore flow directly through the meter. However, as will be seen later, the value of this current may be readily designed to be about 0.3% of the meter's f.s.d.

On application of a voltage to the input terminals, the balanced states of $T_r$ and $T_r$ are initially upset, causing the voltage at the collector of $T_r$ to shift according to the polarity of the input signal. The voltage shift at the collector of $T_r$ is amplified by the Darlington pair, producing an even greater and opposite shift in the collector voltage of $T_r$. A voltage is thus developed across $R_s$, in opposite phase to the input voltage, which returns the differential pair to a balanced condition. Owing to the presence of the meter diodes, the current flowing through the moving-coil meter, which is a constant fraction of the current flowing through the resistor $R_s$, always flows in one direction, no matter what the polarity of the input voltage.

Provided that the open-loop gain of the amplifier is very high, the voltage developed across $R_s$ (equal to $R_s I_l$, where $I_l$ is the current flowing through the meter circuit) will be equal to:

$$V_{in}(R_3 + R_{m_1})(R_3 + R_{m_2})$$

Thus, $R_s I_l = V_{in}(R_3 + R_{m_1})(R_3 + R_{m_2})$ and the input voltage sensitivity for f.s.d. equals $R_s I_l$, where $R_s I_l$ is the current flowing through the meter and when the meter indicates f.s.d. The current flowing through the meter is thus directly proportional to the applied input voltage.

In order to overcome the non-linearity of the rectifying diodes the open-loop gain of the amplifier should be very high, allowing the large amount of negative feedback to be applied. The negative feedback effectively increases the output impedance as 'seen' by the meter circuit. Thus an alternative to a high open-loop gain is to increase the output impedance as seen by the meter circuit by some other means. The simple circuit of Fig.2 does not possess either a sufficiently high open-loop gain or high output impedance to satisfactorily overcome the non-linearity of the rectifying diodes. Hence, the more sophisticated circuit of Fig. 3 was developed. Fig. 3 is basically the same circuit as Fig. 2 except for the fact that $R_{C_1}$, $R_{C_2}$ and $R_{E_1}$ have all been replaced by constant current sources. By replacing $R_{C_1}$ with a constant current source, the output impedance of the output stage is made extremely high by virtue of the fact that the collector impedance of a transistor is very high. This, coupled with a reasonably high open-loop gain, satisfactorily overcomes the non-linearity of the rectifying diodes. This technique has been successfully carried out by the writer in a previous a.c. millivoltmeter design 1. By replacing $R_{E_1}$ with a constant current source, the current flowing through the differential pair of transistors is maintained at a constant level, regardless of the potential at the two transistor bases. It also has the advantage of increasing the common-mode rejection ratio of the differential pair. The replacement of $R_{C_1}$ by a constant current source is not really necessary from the point of view of greatly increasing the open-loop gain of the amplifier or its overall performance. However, by doing so, and by virtue of the fact that $R_{E_1}$ and $R_{C_2}$ have been replaced by constant currents, the d.c. stability of the amplifier becomes virtually unaffected by variations in the supply voltages of plus or minus one volt. Thus there is no need for stabilized voltage supplies.

The heart of the amplifier remains the transistors $T_r$ to $T_r$ which are all contained in the integrated circuit, CA3018A, which is a tighter specification version of the CA3018. The two isolated transistors in this package are matched for $h_{FE}$ to ±10% and $V_{BE}$ to ±2mV over a range of collector currents from 10µA to 10mA. This type 'A' version of the package was specifically used because $h_{FE}$ of the two isolated transistors is maintained to a minimum of 30 at a collector current of 10µA. The current gain of the Darlington pair is a minimum of 1000 at a collector current of 100µA and 2000 at 1mA. The collector currents of the differential pair are designed to be 10µA. For a minimum gain of 30, the input bias currents are thus a maximum of 0.33µA. By virtue of the fact that the basic sensitivity of the Avo Model 9 meter movement, when shunted with a 10kΩ resistor, is 50µA, and that two 5kΩ resistors are used in the bridge rectifying circuit instead of a further two diodes, the current indicated by the meter is half that which flows through the meter circuit. Thus, under zero input conditions, the meter registers a maximum of half the input bias current, i.e. 0.17µA, which is only 0.3% of the 50µA f.s.d. This is a negligible amount and may be totally ignored under normal measurement conditions.

The current through the potential divider network to the base of the transistor $T_r$ is designed to be 10µA. The constant current supplied by $T_r$ thus needs to be 20µA. The collector current of the Darlington

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Fig. 3. Practical circuit of meter amplifier, with higher open-loop gain and higher output impedance than those of the basic circuit (Fig. 2). In the i.c. the circled numbers indicate the conductors which are taken out to leads (with corresponding numbers) on the TO-5 12-lead package. No. 10 is the chip substrate. (The intersections of conductors with the shaded rectangle have no significance as connections.) $D_1$ and $D_2$ are type OA202.
ton pair is designed to be about 400μA and as the minimum gain of the pair is 1000 at a collector current of 100μA and 2000 at 1mA, the current flowing into the base of \( T_r \) will certainly be less than 0.4μA, which is negligibly small compared with the 10μA flowing through the potential divider.

Since the collector loads of \( T_r \) and the Darlington pair are effectively constant current circuits and since the input impedance of the Darlington pair will be very much less than 750kΩ, the current gain of the amplifier will be approximately equal to the product of the current gains of \( T_r \) and the Darlington pair. This product is likely to have a minimum value of between 30,000 and 60,000 depending on the minimum gain of the Darlington pair at a collector current of about 400μA. A minimum of 40,000 should be a fairly safe assumption. The typical figure will be about 200,000.

The current through the meter circuit for f.s.d. is 100μA, therefore the change in input bias current to the differential pair will be a maximum, for f.s.d., of 100μA/40,000 = 2.5nA. Now for reasonable accuracy the input current to the amplifier should be at least 100 times greater than this value, i.e. 250nA. However, the current gain of the amplifier falls off slightly under low input signal conditions because of the increase in impedance of the rectifying diodes. If it is assumed that the impedance of the meter circuit rises to become approximately equal to the output impedance of \( T_r \) under low input signal conditions, then the current gain will fall by about a half.

Thus the input signal current should not be less than about 500nA for f.s.d. of the meter. For a voltage sensitivity of 100mV this gives an input impedance to the amplifier of 2MΩ/volt, and values for the resistors \( R_1 \) and \( R_2 \) (Fig. 2) of 100kΩ.

The minimum voltage sensitivity of the circuit is basically determined by the ability to accurately balance the open and short circuit input conditions and the tendency of the zero to drift. The difference between \( V_{BE1} \) and \( V_{BE2} \) for the CA3018A, is not more than ±2mV and, since the gains of the differential pair are matched to ±10%, the input bias currents are less than about 500nA 

\[
\frac{100kΩ \times 0.33μA}{100kΩ} = ±3.3μV
\]

With 10μA flowing through the collectors of \( T_r \) and \( T_{r2} \), and hence their emitters, it is a simple matter to correct the open circuit input balance by inserting a 250Ω potentiometer between the emitters as shown in Fig. 3. This enables the difference between \( V_{BE1} \) and \( V_{BE2} \), to be corrected by ±2.5mV, i.e. a swing of 5mV.

In order to correct the short circuit input balance, it is necessary to ensure that the voltages dropped across the resistors \( R_1 \) and \( R_2 \) (of Fig. 2) are identical. This can, of course, be done by altering the value of either \( R_1 \) or \( R_2 \), but as this would upset the gain of the amplifier an alternative method is to alter the currents flowing through these resistors. These currents are the input bias currents to the bases of the differential pair of transistors and can be controlled by varying the collector currents of the differential pair. Since the emitters of these two transistors are fed with a constant current, the sum of the two collector currents will be very nearly equal to this value, i.e. \( I_c = I_{c1} + I_{c2} \), \( I_{c1} \) is a constant current and therefore so also is \( I_{c2} \), equal approximately to \( I_{c1} \). Thus, if \( I_{c2} \) is decreased by an amount \( dI_c \), current \( I_{c2} \) must be increased by the same amount. The 220kΩ fixed resistor and the 25kΩ variable resistor in the emitter circuit of \( T_{r2} \) are designed to vary the collector currents of \( T_r \) over the range of approximately 19μA to 21μA. Remembering that about 10μA of this current flows through the potential divider network consisting of the 360kΩ and 750kΩ resistors, this means that the collector currents of \( T_r \) and \( T_{r2} \) can be varied over a range of 9μA to 11μA.

This gives an overall ±5% control over the base currents of \( T_r \) and \( T_{r2} \). Only a control of 10% is required to compensate for the tolerances of the two transistors but the additional control will cope with the ±5% tolerances of the resistors, \( R_1 \) and \( R_2 \). The short circuit balance may therefore be controlled over a range of ±6.7mV (100kΩ × 1μA/30), i.e. a swing of 13mV.

The two balance controls thus provide a swing of not less than 13mV, which is small compared with a basic sensitivity of 100mV, and no difficulty will be encountered in obtaining an accurate zero.

The stability of the zero with changes in temperature is extremely good. The change in the difference between \( V_{BE1} \) and
The pair of f.e.t.s can be bought with their gate-to-source voltages matched to within \( \pm 100\text{mV} \) at a drain current of 1mA. As a result of this fairly coarse matching, it is not worth buying the tighter specification version of the CA3018. The value of 1k\( \Omega \) for the set-zero control allows for variations of 5% in the values of the source resistors and produces an overall swing of about \( \pm 200\text{mV} \). It should, therefore, be a ten-turn controller or, alternatively, an additional finer control of about 100\( \Omega \) may be placed in series with one of the 9.1k\( \Omega \) source resistors.

The circuit of Fig. 4 does have one disadvantage over the simpler circuit of Fig. 3 and that is a poorer upper frequency response due to the use of higher valued resistors for \( R_1 \) to \( R_3 \). This is to be expected since all resistors have a stray capacitance associated with them and the higher the value of the resistance the greater the effect. Consequently it is recommended that resistors of as high a wattage rating as is practical should be used in these positions as the effect is more reduced the greater the physical size of the resistor.

It was found necessary with the circuit of Fig. 4 to connect a capacitor and resistor, \( R_2 \) and \( C_2 \) of Fig. 4, in series between the collector and base of transistor \( T_2 \), in order to prevent high-frequency instability. By trial and error, practical values for the resistor and capacitor of 5.6k\( \Omega \) and 470\( \text{pF} \) were found to produce the desired effect.

Prototypes of both circuits, Figs. 3 and 4, have been constructed by the writer. The circuits were tested for linearity and frequency response at their basic voltage sensitivities of 100mV. The scale linearity of both circuits is excellent, there being no detectable error at 10% or even 3% of f.s.d. The circuit of Fig. 3 has a frequency
response, flat within ±2%, from d.c. to 100kHz. The circuit of Fig. 4 does not perform quite so well up at the 100kHz end of the scale, the reading falling off by about 10%. This frequency response, however, was achieved without making any attempt to compensate for the stray capacitances across the 10MΩ resistors.

The writer has constructed a prototype multimeter, with voltage and current ranges, using the circuit of Fig. 3 with a suitable attenuator as shown in Fig. 5. By suitable adjustment of the feed-back resistor in the meter circuit the multimeter was made to operate with a 100μA meter of the Japanese 'SEW' type. The multimeter was constructed in an aluminium die-cast box of dimensions, 4¾in wide × 7¾in long × 3¾in deep. The ranges provided by the attenuator are, 100mV, 1V, 10V, 100V, 1kV, 10kV, 100kV, 1mA, 10mA, 100mA, and 1A, a.c. or d.c. The inclusion of the two-pole switch and the associated components shown in Fig. 5 enables a.c. signals only to be measured with an r.m.s. scale for sinewaves. This is essential if a.c. signals are to be measured in isolation from the d.c. bias conditions of a circuit. If it is required that the meter should be able to give an indication of the polarity of a d.c. signal, a diode may be inserted in the meter circuit between the junctions of the resistors R1 and R2, and R3 and R4 of Fig. 2. This diode should normally be short-circuited by a single-pole, push-button switch. On pressing the switch and disconnecting the short-circuit across the diode, the meter will continue to indicate the measured voltage only if the polarity of the input signal is in one particular direction (determined by the 'way round' the diode is connected into the meter circuit). If the polarity of the signal is in the other direction, the reading will fall to zero. This technique maintains the amplifier's advantage of always giving a 'positive' meter deflection no matter what the polarity of the input d.c. signal, at the same time as providing a means of determining the signal's polarity. The input terminals must therefore be marked + and -.

Having discussed the theoretical aspects of zero drift for the circuit of Fig. 3, the writer should perhaps mention that in practice the prototype multimeter constructed by him has not exhibited a zero-drift of greater than 1% of f.s.d. over a period of use. In all cases it has only been necessary to adjust the short-circuit zero control to obtain a satisfactory zero.

The writer is not able to comment on the practical zero-drift of the circuit of Fig. 4, not having, at the time of writing, tested the circuit in a finally constructed form. There does not appear, however, any reason why it should be very much worse than that of Fig. 3, provided the two f.e.μs can be maintained at roughly the same temperature.

Finally, Fig. 6 is a suggested method of obtaining a number of voltage ranges using the more sophisticated circuit of Fig. 4.

Reference

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**Television Sound Adaptors**

**Methods requiring no electrical connection to the set**

Since publication of the leader in the October 1971 issue, some letters to the Editor in the November and December issues, and an article by J.C.G. Gilbert in the January issue, all on television receiver sound quality, we have heard of several methods of extracting the sound signal from the television set for feeding into hi-fi equipment. Of these, techniques which do not require a direct electrical connection to the receiver would seem the most attractive, first because they avoid the problem of the set's 'live' chassis, secondly because they do not invalidate the maker's guarantee if the set is owned by the user and thirdly because they can be used on rented sets without infringing the rental agreement. These methods depend on the presence around u.h.f. sets (or the u.h.f. part of dual-standard sets) of electromagnetic radiation from the intercarrier sound i.f. signal, which in the U.K. is at a frequency of 6MHz. The magnetic component of this radiation can be inductively picked up outside the cabinet, and the resulting signal, containing the sound information as frequency modulation, can be amplified and demodulated as required.

Two examples of this general technique are shown in the figures. Fig. 1 is extracted from an article in our associated Swedish journal Radio & Television, No. 11, Nov. 1971. This circuit picks up the intercarrier sound i.f. signal, which is 5.5MHz in Sweden, and converts it to a frequency of 10.7MHz to allow it to be fed into the i.f. section of a conventional f.m. tuner. Thus the tuner can be switched to

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give either the normal v.h.f. sound broadcasts or the television sound broadcasts. As can be seen there is a pick-up unit containing a ferrite-cored coil and capacitor tuned to 5.5MHz and an f.e.t., which forms the bottom half of a cascade input stage. The upper half of the cascode stage is the 2N2369 bipolar r.f. transistor, which is housed in the main unit of the adaptor. The final transistor, in conjunction with the 16.2MHz tuned circuit, is an f.e.t. oscillator/mixer. In this the local oscillation is heterodyned with the 5.5MHz signal from the cascade stage to give a difference frequency of 16.2 − 5.5 = 10.7MHz, which is selected from the sum and difference frequencies by the output tuned circuit and transformer in the drain circuit of the f.e.t.

Fig. 2 is a block diagram of a commercial adaptor called the Telefi which has just become available in the U.K. (makers: Dinosaur Electronics Ltd, 85 Victoria Street, Windsor, Berks). This device provides an audio output signal (typically 1V r.m.s.) for feeding into a hi-fi amplifier. The change-over switches seen on the right are simply to allow the user to select either the output of the adaptor or the output of his f.m. tuner. Here the pick-up unit is simply an encapsulated coil on the end of a screened cable, which is connected to the main unit containing all the electronics and switches. First the 6MHz intercarrier i.f. signal is amplified in a single-transistor stage, then the required band of frequencies on 6MHz is selected by a band-pass filter. Next the frequency modulated 6MHz is applied to an integrated circuit which includes an amplifier, a limiter and an f.m. discriminator (this being of the coincidence detector type described by T. D. Towers in the September 1971 issue, p.434). The audio amplifier includes a 'user' pre-set gain control, and the output a.f. signal is claimed to have less than 1% distortion at full modulation at a level of 1V r.m.s. Wireless World has tried out this device and found it completely satisfactory. The main unit is housed in a wooden case which can be stood on the top of the TV set or placed anywhere nearby. The encapsulated coil may be stuck to the insulating back cover of the set, as near to the sound i.f. amplifier as possible, or allowed to dangle loose near the back cover. Its position and orientation were found to be non-critical on a set containing a B.R.C. type 1500 chassis. The manufacturers say that with one of the five TV sets on which they have tried the adaptor the pick-up coil could be up to 3ft away!

Information on Standards

The British Standards Institution maintains an affiliated organization for standards engineers, metrication officers and others concerned with standardization in every industry. Known as B.S.I.'s Standards Associates Section it provides a meeting place for the exchange of views on standards matters and a feedback of information to B.S.I. on the effectiveness of standards. The subscription of £3 a year includes a copy of the magazine B.S.I. News each month which keeps members up to date on metrication and lists new and revised standards issued; early notification of standards conferences and courses at special rates, and, in most parts of the country, a regular programme of meetings. Information from the Secretary, Standards Associates Section, B.S.I. 2 Park Street, London W1A 2BS.
Current-limited Power Supply

A simple design of relatively low cost with dual range current and voltage supplies

by A. Royston*, B.Sc. (Hons), Ph.D.

A stable power supply with a controllable maximum current and a widely variable output voltage is an essential aid to experimental electronics. This article describes a unit which was designed to be simple and fairly cheap to build and yet have a performance comparable to those available commercially. It might be thought that the circuit is retrograde in that it does not use currently available monolithic voltage regulators, but apart from the slightly expensive Motorola MC1466, none of those known to the author gives a comparable performance, particularly with respect to current limiting. Also, few external components are saved by their use. Performance of the unit is given in Table 1 and a list of symbols used in Table 2.

Circuit description

Fig. 1(a) shows a constant-voltage source in which the error amplifier, IC1, feeds with the positive output terminal - this ensures that the only circuit component to have a high voltage across it is the series element Tr1,2.

Since the voltage drop across R1 is constant (Vref) and the input current to the error amplifier is assumed to be zero, the output voltage, \( V_{out} \), is

\[ V_{out} = V_{ref} \frac{R_2}{R_1} \]

This implies that the gain of the amplifier is infinite, and for this design the expression will not be more than 0.001% in error.

As it stands, the output current of this circuit will be limited only by the capability of the series element, and in any case it would be destroyed by short-circuiting the output so some means of controlling the maximum load current must be provided to protect both the supply and any load connected to it. Fig. 1(b) shows a simple constant-current source in which the voltage across a current sensing resistor, \( R_p \), is compared with the drop across part of the potential divider chain formed by \( R_1 \) and \( R_2 \). If we assume that IC2 is a perfect differential amplifier,

\[ I_{lim} = (R_1, V_{ref})/(R_p, R_2) \]

The two sources are combined in Fig. 1(c) such that the maximum load current will be \( I_{lim} \) and the highest output voltage will be that defined by \( R_1 \). (It should be remembered that an ideal current source has an unlimited output voltage capability to accommodate high load resistances.) This implies an OR function which is implemented in Fig. 1(c) by Tr. If the load current is less than \( I_{lim} \) the inverting terminal of IC2 is negative of the non-inverting input causing the output of this amplifier to be positive with respect to the positive output terminal. Because \( R_d \) supplies sufficient base current to Tr for this transistor to be saturated, the output of IC2 is effectively connected to the base of the series element Tr1,2, giving the equivalent of Fig. 1(a). If the load current is increased to near \( I_{lim} \) the output of IC2 will fall until the base current of Tr is insufficient to cause saturation so in attempting to compensate the output of IC1 will rise. When the output voltage has fallen by a small amount this amplifier will saturate positively and as Tr is now an emitter follower IC1 will have complete control of the output current. By following this argument in reverse, it can be readily seen that a subsequent increase of the load resistance to such a value that the load current is less than \( I_{lim} \) means that the voltage control amplifier will resume control over the output voltage. Thus we have a

* University of Durham
current limited voltage source, and a voltage limited constant current source for which the change from one characteristic to the other requires only a change of about 1mV and 100µA in the load voltage and current respectively. In the appendix a more detailed treatment of the theoretical performance is given.

Fig. 2 shows the implementation of the diagrammatic circuit of Fig. 1(c). Practical considerations make it necessary that the heat dissipation of $Tr_1$ be kept to reasonable limits by splitting the output voltage into two ranges, 0-25V and 25-50V, and having corresponding a.c. voltages of 30V and 48V supplied to rectifier $D_6$(Fig. 3). When the lower range is selected the reservoir capacitor $C_6$ has 36V across it (load current of 1.0A) so the greatest short circuit dissipation of $Tr_1$ is about 36W. For the upper voltage range the corresponding dissipation would be a somewhat excessive 36W. To reduce this re-entrant foldback protection has been incorporated by including $Tr_2$ and $Tr_5$, together with some associated circuitry which is inactive while the lower range is selected. If the voltage between $A$ and $B$ is more than 48V the zener current of $D_1$ flowing through $R_1$ causes $Tr_5$ to saturate and the junction of $R_6$ and $R_7$ to be 23.5V with respect to the negative output terminal. For as long as the output voltage is more than this value $D_1$ is reverse biased and no emitter current can be caused to flow from $Tr_5$, but if the supply current limits the output, voltage can drop below 23.5V. Diode $D_1$ now conducts and as the output voltage falls to zero $Tr_5$ saturates (collector-base voltage zero) causing $R_4$ to parallel the current setting potentiometer, $R_{30}$. Thus the short-circuit current is defined by this resistor combination. It is important that $R_4$ should supply only sufficient current for $Tr_5$ to just saturate or the supply would latch off, with a zero short circuit current, and the output voltage would not rise with removal of the load. Fig. 4(a) shows the current/voltage characteristic for both ranges, while Fig. 4(b) indicates the corresponding dissipation of $Tr_5$.

Fig. 3 shows all the internal supplies. The mains transformer supplies 30V or 48V a.c., selected by the range switch $S_1$, which is rectified and smoothed by $D_3$ and $C_6$, respectively. The positive and negative supplies, which float with the positive output of the circuit, are given by an additional secondary. To obtain a ripple-free reference voltage $D_3$ is fed indirectly by a constant current source $Tr_6$. For the negative voltage, less current is required so far simpler smoothing has been used, but

### TABLE 1. Performance

<table>
<thead>
<tr>
<th>Output voltage</th>
<th>Fully variable, two ranges</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output current</td>
<td>Fully variable, two ranges</td>
</tr>
<tr>
<td>Load regulation*</td>
<td>$r.A = 0.1A$</td>
</tr>
<tr>
<td>Current regulation</td>
<td>$10%$ means variation</td>
</tr>
<tr>
<td>Ripple and noise</td>
<td>Constant voltage mode, $r.I = 0.1A$</td>
</tr>
<tr>
<td>Voltage</td>
<td>Constant current mode</td>
</tr>
<tr>
<td>Current</td>
<td>Temperature coefficient</td>
</tr>
</tbody>
</table>

* These figures are the limit of the accuracy of measurements of the equipment available to the author. They should be better than 0.002%.

### TABLE 2.

<table>
<thead>
<tr>
<th>Symbols used</th>
<th>Typical value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A$</td>
<td>Voltage gain of differential amplifier</td>
</tr>
<tr>
<td>$dV$</td>
<td>Voltage between the input terminals of the differential amplifier</td>
</tr>
<tr>
<td>$I$</td>
<td>Load current</td>
</tr>
<tr>
<td>$i_{ca}$</td>
<td>Load current at limiting</td>
</tr>
<tr>
<td>$L$</td>
<td>Load regulation (%)</td>
</tr>
<tr>
<td>$V$</td>
<td>Negative reference voltage</td>
</tr>
<tr>
<td>$V_m$</td>
<td>Output voltage with no load current</td>
</tr>
<tr>
<td>$V_{oc}$</td>
<td>Output voltage with load current</td>
</tr>
<tr>
<td>$V_{ref}$</td>
<td>Positive reference voltage</td>
</tr>
<tr>
<td>$R$</td>
<td>Output resistance of differential amplifier</td>
</tr>
<tr>
<td>$R_1$</td>
<td>Base resistance of $Tr_1(V_i)$</td>
</tr>
<tr>
<td>$R_2$</td>
<td>Current sensing resistor</td>
</tr>
<tr>
<td>$R_3$</td>
<td>Current gain of $Tr_2$</td>
</tr>
<tr>
<td>$R_4$</td>
<td>Current gain of $Tr_3$</td>
</tr>
<tr>
<td>$A^V$</td>
<td>Change in output voltage when load current changes by 1 amp</td>
</tr>
</tbody>
</table>

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Fig. 2. Circuit of the regulator.
Fig. 3. Internal supplies.

Completed prototype, showing the layout of controls.

Fig. 4. (a) Output characteristic. (b) Tr, dissipation (W).

Internal view, showing the position of the larger components, p.c. board and heat sink at the rear.

Fig. 5. P.c. board, copper side.
short circuited suddenly the current control amplifier had an occasional tendency to latch-up, preventing current limiting. In their place, the more sophisticated SN74741Ns have been used as they do not display this characteristic. In addition, they are internally compensated and have such a high open loop gain, that it is difficult to measure the performance of the circuit.

The maximum voltage across \( T_{R2} \) is about 72V, so the ubiquitous 2N3055 with its \( V_{CE} \) of 65V would not be a good choice. The RCA 40636 is an improved version with a \( V_{CE} \) of 95V. Transistor \( T_R \) has a similar rating, and also recommends itself by having an integral heatsink sufficiently large to dissipate the heat generated by the base current of a worst case 40636 (\( I_B = 30 \) mA at 1.6W).

Diodes \( D_1 \), \( D_2 \), and \( D_3 \) protect the inputs of the integrated circuits.

To aid construction Fig. 5 shows the conductor side of a suitable printed circuit board which contains all but the largest of the components. Its layout is shown in Fig. 6, and Fig. 7 indicates the wiring from the edge connector to the various external components and controls. To obtain the stated performance it is essential that wiring carrying high currents be done carefully. Resistor \( R_{19} \) should be returned directly to the negative output terminal, and pin 9 of the edge conductor, together with the current sensing resistor \( R_{20} \), must be joined by a low resistance path to the positive terminal. Otherwise a wiring resistance of 5m\( \Omega \) can make the load regulation worse by a factor of twenty. This problem has been discussed in a Motorola application note, and could be solved by means of sense terminals.

The unit was built into a Contil Mod 2B case with the heatsink mounted externally. Total cost was about £17, although the cheapest components were not used.

References
1. Motorola application note AN497.
2. Motorola application note AN480.

As it provides the reference voltage for the limiting current it is perhaps unfortunate that the temperature coefficient of 8.2V zener diodes is somewhat higher than that of 5.1V zeners (0.006% per °C against 0.007% per °C). If this is unsatisfactory, \( D_4 \) could be replaced by one similar to \( D_3 \), giving a common mode voltage of about -2V to \( I_C \), and \( I_C \). This would probably not affect circuit operation.

To give the higher voltage range \( S_{10} \) switches an extra resistor \( (R_{21}) \) into the voltage sensing chain and the two current ranges are implemented by altering the current sensing resistor from 1\( \Omega \) to 10\( \Omega \) by means of \( S_{10} \). The last switch, \( S_{12} \), allows a meter to display the load current or voltage. Preset potentiometers \( R_{22} \) and \( R_{23} \) set the maximum output current and calibrate the meter current range, respectively, while \( R_{19} \) varies the limiting current. Adjustment of \( R_{19} \) causes the output to swing by about 26 volts.

Constructional details
Originally SN72702N amplifiers were used but it was found that when the supply was

![Fig. 6. Layout of p.s. board.](image)

Figure 6. Layout of p.s. board.

![Fig. 7. Connections to p.c. board.](image)

Figure 7. Connections to p.c. board.

Appendix—Theory
In the foregoing account it was assumed that the devices used in the circuit were ideal. Here the effect of using real amplifiers is examined.
It is assumed that $dV$ for each of the error amplifiers is zero when there is no load current, and that their input bias current is zero. Also the base-emitter voltages of $TR_1, TR_2, TR_3$ are ignored.

**Constant voltage mode** (see Fig. 1(a))

With no load current the output voltage is,

$$V_o = V_{ref} \frac{R_2}{R_1}$$

If the current is drawn from the supply an error voltage is developed across the input of $IC_2$ to compensate for the drop over $R_1$ and $R_2$ so,

$$V_{out} = \left( V_{ref} - dV \right) \frac{R_2}{R_1} - dV$$

where

$$dV = \frac{I(R_1 + R/\beta)}{A}$$

Thus the change in output voltage is

$$\Delta V_o = V_o - V_{out} = \frac{I(1 + R_2/R_1)(R_1 + R/\beta)}{A}$$

The load regulation, $L$, is defined as

$$L = \frac{\Delta V_o}{V_o} \times 100\% = \frac{I(1 + R_2/R_1)(1 + R_1/R_2)}{AV_{ref}} \times 100\%$$

and the output resistance is

$$R_{out} = \frac{\Delta V_o}{I} = \frac{(R_1 + R/\beta)(1 + R_1/R_2)}{A}$$

We can calculate the performance using the typical parameters:

<table>
<thead>
<tr>
<th>Current Range</th>
<th>0.1</th>
<th>1.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Voltage</td>
<td>5 V</td>
<td>50 V</td>
</tr>
<tr>
<td>$L$</td>
<td>0.0004</td>
<td>0.0002</td>
</tr>
<tr>
<td>$R_{out}$</td>
<td>0.20</td>
<td>1.1</td>
</tr>
<tr>
<td>$\Delta V_o$</td>
<td>0.02</td>
<td>0.11</td>
</tr>
</tbody>
</table>

The figures are considerably better than the measurement accuracy of the instruments available to the author, but the performance of a similar regulator confirms them.3

**Constant current mode** (see Fig. 1(c))

Here $TR_3$ acts as an emitter follower, now $dV = I_{i\text{em}}(R_1 + R_2/\beta)A$ and at limiting the drops across $R_1$ and $R_2$ differ by $dV_2$ so

$$I_{i\text{em}} = I_{i\text{em}}R_1 + (I_{i\text{em}}R_3 + I_2)\frac{R_3 + R_4}{A} + dV_2$$

after some re-arrangement

$$I_{\text{lim}} = \frac{R_3 V_2}{R_4 R_1 \left(1 - \frac{R_3 + R_4}{A} \frac{1 + R_2/\beta}{\beta} \right)}$$

For the simple case $A = \beta = \infty$

$$I_{\text{lim}} = \frac{R_3 V_2}{R_1}$$

so the actual current is higher than that given above by

$$(1 + R_2/\beta)(R_3 + R_4)/AR_1 \times 100\%$$

Since $R_2 = 8R_3$ ($I_{\text{lim}} = 100\text{mA}$ or 1.0A this error will be about 0.001%.

Current regulation is made worse by two factors. First, shunting of the load by $R_{12}$ when the meter is used to measure the output voltage, and secondly by the voltage sensing chain. This second effect can be removed by connecting the inverting input of $IC_2$ to the other end of $R_1$, instead of the output, leaving the two diodes between pins 4 and 5 of $IC_1$. This connection must be used only in the constant current mode, the output voltage before limiting will be

$$V_{out} = V_{ref} \frac{R_2}{R_1} - IR_1$$

Since the current and voltage are linearly dependent on the values of $R_{10}$ and $R_{21}$ these potentiometers could be replaced by helipots or decimally switched resistors.

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**Announcements**

**B.C.S. approval, Marconi Instruments have received British Calibration Service approval for their Electrical and Electronic Calibration Laboratories at Luton and the Microwave Standards Laboratory of the Sanders Division at Stevenage. The Measurement Standard Laboratory at the company's St. Albans headquarters received B.C.S. approval in 1968.**

Lampitt Electronics Ltd, 48 Princess Street, Manchester M1 6HR, have signed an agreement with SABA to distribute the German company's range of colour television, radio and hi-fi equipment.

Following the introduction in the U.K. of the "R remotecop" facsimile transmission equipment, Plessey Communication Systems have signed a long-term agreement with Visual Sciences Inc, of New York, for the world-wide marketing (outside the U.S.A. and Japan) of the American designed facsimile transceiver.

The Carrier Corporation, owners of the electronic components and potentiometer manufacturing organization Spectrol Electronics Corporation of California, have acquired Reliance Controls Ltd, of Swindon. The company name has been changed to Spectrol Reliance Ltd.

Emilus Microcomponents Ltd have signed an agreement under which the company will sell and distribute in the U.K. the range of standard m.o.s. integrated circuits manufactured by the Monst Corporation, of Carrollton, Texas, U.S.A.

ICI and Ciba-Geigy have acquired from CBS of America its interests in the EVR Partnership. The Partnership will be granted non-exclusive rights to exploit the EVR system in the U.K. and Canada.

Brandenburg Ltd, manufacturers of high-voltage power supplies, have opened a sales office in Germany. The address is: bei Edwards Hochvakuum GmbH, 6 Frankfurt am Main, Niederrad, Hahnstrasse 46, (Postfach 160).

Sims-World International Ltd, Hodford House, 17-27 High Street, Hounslow, Middx, have recently been appointed UK distributor for the range of electronic components supplied by Thomson-CSF United Kingdom Ltd.

Daly (Condensors) Ltd, electrolytic capacitor manufacturers, are to set up a subsidiary, Daly Processes Ltd, to specialize in the anodizing or forming of aluminium foil.

**Signatrol Ltd, Denmark House, Devonshire Street, Cheltenham, Glos, is to market in the U.K. the range of computer products manufactured by the Analogic Corporation of Massachusetts, U.S.A.**

**Seatronics (UK) Ltd, 22-25 Finsbury Square, London EC2A 1DT, have appointed the French company, International Passive Components - exclusive agent and distributor throughout France.**

**Guest International Ltd, of Thornton Heath, Surrey, have acquired the U.K. agency for the Linteronix range of solid state lamps and digital displays.**

**Advance Electronics Ltd, of Bishop's Stortford, and Ballantine Laboratories Inc., of Boonton, N. J., have announced an agreement in which Ballantine will in future market Advance instruments throughout the United States.**

**Taylor Instrument Companies (Europe) Ltd., of Stevenage, Herts, manufacturers of control systems for the process industries, have acquired Servonics Controls Ltd, of Crowborough, Sussex, who make analytical instruments for industrial and laboratory use.**

**Transducers (C.E.L.) Ltd., Trafford Road, Reading RG1 8JH, have been appointed exclusive U.K. agents for the ranges of precision dia-stamped foil strain gauges and temperature sensors manufactured by Dentronics Inc., of New Jersey, U.S.A.**

**Data Recognition's ranges of Datamec 3 and Datamex 2 optical mark reading data-capture systems are now available combined with any of the embossed-plate printing machines manufactured by Bradma Ltd, to enable semi-variable alphanumeric or bar-marked data to be easily printed onto o.m.r. forms.**

**Decca have received orders for radars totalling over £200,000 from BP Tanker Co Ltd. Twenty of the 22 tankers are to have dual 16-in installations.**

**The Post Office have placed a £1.5M contract for final development, supply and installation of large electronic exchanges of the TNE4 design with Standard Telephones and Cables Ltd.**

**Pye TVT Ltd have received the first u.h.f. colour television transmitter order, valued at £40,000, from the South African Broadcasting Commission. The order was placed with Philips Telecommunications, South Africa, Pye TVT's agents.**

**Frigates being built by Vosper Thornycroft Ltd, for the Brazilian Navy will be fitted with tactical action information and weapon control systems designed and developed by Ferranti's digital system division. The systems are valued at well over £5m.**

**Redifon Telecommunications Ltd have received an order, valued at £63,000, to supply h.f., s.s.b. receivers to the Civil Aviation Division of the New Zealand Ministry of Transport for use in air traffic control networks.**

**Standard Telephones and Cables Ltd have received a contract from the Post Office to supply over £1M worth of pulse code modulation equipment to increase the circuit capacity of existing junction cables between exchanges.**

**Tape Recorder Spares Ltd, Tape Recorder Components Ltd, and Audio Packs Ltd, members of the Tape Recorder group of companies, have moved to premises at 206/210 Hertford Road, London SE15 INS. (Tel. 01-639 7844).**
Corner horns

In the May 1970 issue of Wireless World, 'Toneburst' described a domestic corner horn which reproduced bass down to below 30Hz. Ralph West referred in the June issue to having measured it down to at least 40Hz. 'Toneburst' in the July 'Letters to the Editor' suggested that "the flaring rate made in conjunction with the walls is too rapid to be accommodated by horn theory for the results attained. Even for a 40Hz cut off the cross sectional area should double in about 16in".

It is worth considering further what is the flaring rate resulting from the walls and the floor. One may perhaps normally think of the expansion of a corner as the increase in the distance between corresponding points on the walls as distance from their junction increases. The useful expansion, however, is that of the trihedral corner formed by the walls with the floor. This can be represented as a succession of larger and larger equilateral triangles coming out from the apex of the corner down on the floor, their midpoints forming the axis of the expansion. The areas of a number of such triangles, at 1ft distances along the axis, are shown as crosses on the accompanying graph. The graph shows in addition the approximate equivalent of the volume occupied by 'Toneburst's' speaker, the expansion of the corner net of this being shown by the upper boundary of the diagonal hatching.

The superimposed lines showing the laws of expansion of exponential horns with minimum useful frequencies of 65, 40 and 30Hz indicate that we are dealing with something that is essentially a set of segments of horns of differing expansions succeeding each other imperceptibly as far as the corner flare is concerned. But there seems to be a bad junction with the loudspeaker itself. The attested performance is presumably obtained in the face of this disadvantage.

It would be fun to try energizing an array of speakers on an equilateral baffle about 20 sq ft in area and mounted in a corner in an effort to match more closely to the useful part of the corner flare. A large box baffle (over 20 cu ft) should pick up most of the advantages with less difficulty. Finally, one or more available speakers could be packed out to 20 cu ft with back issues of Wireless World to see the effect.

S. W. GILBERT.
Beckenham, Kent.

Electrostatic headphones

In reply to Mr. Halliday's letter, in the January issue apropos my article (November issue), and taking his points in turn: First, he is badly misguided if he believes plastic foam has no acoustic function! For a start there are many types of foam rubber or foam plastic — latex, polyurethane and polystyrene being only three of the better known ones. Depending on the process of manufacture these are again subdivided into many types, and each type has its own acoustical transmission/absorption spectrum, hence giving quite a distinct colouration of 'true sound' when used in any transducer. A specific example of the use of foam to alter the measured frequency response of an earphone is given in H. Souther's description of the Koss type ESP6, in Hi-Fi News, February 69; the use of a foam/foil combination in an acoustical cavity is clearly shown to remove several kinks from the response curve. Hence the properties of the transmission tunnel lined with latex foam might be somewhat easier to measure than calculate. This was done and the results are given in the article.

Secondly, I am afraid I must accept that the performance of the earphones is limited by the rest of the system, as the response taken on its own is extremely good. (It is a pity we cannot say the same for most electro-magnetic transducers). I would accept that 0.1% distortion is a good figure to aim for, and am pleased that Mr. Halliday has taken the second step in this direction, but I do not really believe that the amplifier proposed will ever be widely used or accepted for the same reason as I have designed a second amplifier. The reason is that any circuit involving valves nowadays is subject to noise, replacement costs, and inbuilt obsolescence.

As there is a lack of groundwork in audio high-voltage design of this type the amplifier is, of necessity, simple in construction and conception. The commercially available e.s. headphones that I know of overcome this problem with the use of a transformer, with its own special inbuilt distortion. Perhaps we shall see a rapid development of a suitable drive amplifier to the standards of perfection we have become accustomed to in audio work today. This development notwithstanding, I would be surprised if one could buy an audio system for much below £150 giving as linear and distortion-free an output from its speakers as one can obtain from the system described in the article.

Thirdly, one must certainly bear in mind the limitations of the application of, and the results given by, the electro-mechanical analogy. The -6dB frequency limits given are those calculated in terms of power and certainly become -3dB points if calculated in terms of current required for a given audio output. The lower limit is in practice a lot higher than 0.5Hz as is pointed out. This is not only to be expected, but welcomed, as the earphones are quite capable of reproducing any rumble or other low-frequency sounds emitted by modern decks. In fact would-be designers of an amplifier for these earphones ought to beware and cater for a very sharp roll-off below 30Hz!

The last paragraph of the letter is
somewhat incomprehensible to me as I fail to see any direct relation between small-signal sensitivity and bass response. In answer to the query, however, I hope the following will suffice. For normal listening conditions the earphones require quite large signal voltages — in fact something of the order of one fifth of the bias potential is required. In constant-voltage mode this would cause fairly severe distortion. In constant charge mode this is not so. To achieve a true constant-charge condition the diaphragm resistance would have to be infinite, which is physically impossible. A conflicting requirement is also present to provide a small current that flows through the diaphragm and small leakage paths to earth. The best compromise has been found to be $10^4$-$10^6$ ohms per square. If the value were to be any smaller than this the earphone's intrinsic $RC$ time constant would be such as to allow the charge on the diaphragm to fall during bass notes of the same order of time constant, thus reducing the sensitivity.

In conclusion may I thank Mr. Halliday for his interest in the article and the many points he has brought up.

PHILIP D. HARVEY,
Chelmsford, Essex.

**Swept-frequency audio oscillator**

I read the article on the construction of a swept-frequency audio oscillator written by R. J. Ward which appeared in your September issue with considerable interest. Mr. Ward is to be congratulated on the production of a really practical design.

I would like to offer the attached circuit as an alternative to the active filter used by Mr. Ward. The measured frequency response of a breadboard version is given in the graph. The response is not as flat over the pass band as the calculated response is, but as it is within the design limits (3% amplitude error) no fiddling was done to improve it.

The resistors used are from the Philips 5% range. It has been my experience that it is rare indeed for one of these resistors to depart from its marked value by as much as 2%. The capacitors are from the S.T.C. polystyrene and foil range — type 455-105 which are close tolerance and the transistors are BC 107s. There was no special selection of the components.

Naturally the source resistance has to be included in the $5.1k$ input element.

**T.T.L. trigger circuits**

The following comments may be of interest to potential users of the t.t.l. trigger circuits described by H. A. Cole in the January issue (p.31).

In order to operate reliably t.t.l. devices require a fast edge to the driving waveform. If this is not so there is a possibility that the gate will oscillate as the input voltage reaches a level which biases the gate as an amplifier. The result of this is that for one input voltage level change, several output pulses will occur.

To interface slow rise and fall times into t.t.l.-compatible speeds some manufacturers have introduced a dual four-input Schmitt trigger device, the SN7413. With this device input voltage changes of the order of volts per second can be interfaced with normal t.t.l. devices without risk of a multi-pulse output.

While it is feasible that many 'monostable' circuits of the type suggested by Mr. Cole may function without 'multi-pulsing', or that such malfunction will not be detrimental to the performance of certain systems, it is suggested that the SN7413 device may be better employed as the pulse output gate, particularly where pulse lengths in excess of a microsecond or so are contemplated.

It may be of interest to note that by connecting a 330 ohm resistor between the input and output terminals of the '13' gate and connecting the input to the zero volt rail via a capacitor the device will provide a most useful oscillator for driving such devices as bistables and counters.

Further details of the '13' gate and its uses are contained in one of the application reports published by, Texas Instruments Ltd.

D. B. STILES,
Bristol, Somerset.

**Differential discriminators**

I was surprised to read ('Letters', January issue) of the trouble R. W. Penny says he experienced when operating the type of differential discriminator described by me in the December issue, page 603. I have had a single-channel analyser based upon this design in quantity production for more than three years now without experiencing any of these difficulties. This instrument, described in references 1 and 2, is capable of operating at 10MHz within the dynamic range 25mV to 5V, and uses standard t.t.l. throughout.

I suspect Mr. Penny may have been trying to operate the comparator at bias levels set below the intrinsic backbias (about 5mV). If this is attempted, periodic latch-up will occur and meaningless channel output pulses will be produced. As far as t.t.l. switching spikes are concerned, I agree these can be troublesome but with a good earth plane and single-point earthing for the comparators (preferably at the signal input socket), they can usually be kept below an acceptable limit.

H. A. COLE,
Reading, Berks.


**Free bump-testing!**

My wife and I are planning an expedition, to take approximately a year, during which time we shall be travelling through the Sahara to South Africa, across to Australia and from there via Indonesia, Malaysia, Thailand, and Pakistan to India and then home. I am particularly interested in using modern equipment for our expedition.

May we please extend through your magazine a request to manufacturers to help us to provision our expedition. Any item would be welcomed especially equipment that is compact, light-weight and either has a small current drain at 12 volts or an internal supply. We can supply what must be one of the most rigorous field trial and 'bump-testing' services available.

D. M. WILLOUGBY,
Grafton, Balclava Hill, Pell Green, Wadhurst, Sussex.
Electronic Building Bricks

20. The function generator

by James Franklin

We have already seen that some of the operations we wish to perform on information have a specifically mathematical character — addition, subtraction, integration, for example. This means in practice that an electrical variable (or variables) representing information is fed into a 'building brick' and the output of the 'brick' is the input variable operated upon in some way in accordance with a mathematical rule. The general mathematical way of describing such situations is to say that the output, which will also be a variable, is a function of the input variable. Using mathematical symbols to express this, if the input is \( x \) (representing, say, a varying voltage or current) the output is \( f(x) \), a function of \( x \). This is shown schematically in Fig. 1 (top). In electronic terms we know that the output is dependent on or related to the input by some, as yet unspecified, law.

So let us be more specific and say what the operation performed on the input is. We will assume that it is the process of squaring. Then we can depict the situation as in Fig. 1 (middle). Here \( f(x) \), the function of \( x \) which is the output in this particular case is \( x^2 \). To simplify matters we can replace the symbols \( f(x) \) by a single symbol, say \( y \). Then the output variable \( y = x^2 \) as shown in Fig. 1 (bottom). This is an equation, which can be plotted as a graph to show the relationship between \( x \) and \( y \) — see Fig. 2 — and can be considered as a 'mathematical model' of the relationship.

We have made this incursion into mathematical concepts and terminology really to explain what is meant by a building brick described as a 'function generator'. Electronically we think of what goes on inside any of the boxes in Fig. 1 as the generation of a function — a function of the input variable. Actually the word 'generator' is a bit misleading because it suggests an electrical device that doesn't need an electrical input, whereas in fact this building brick is a sort of modifier of input signals. However, if one thinks of the mathematical operation as being 'generated' then the term is perhaps acceptable. Function generators are used in analogue computers and in other information processing equipment where analogue computing operations are performed on signals. For example there is one type of analogue multiplier which works on the 'quarter-squares' principle and this uses two 'squaring' function generators of the type illustrated in Figs. 1 and 2. Another application of the squaring function generator is to derive an output signal which is proportional to electrical power (Part 8) from an input signal representing electric current or voltage.

In communications systems one type of function generator is used for compressing the dynamic range of input signals. This means that the electrical signal, in passing through the device, is modified so that it is no longer proportional to the physical variable it represents (e.g. air pressure): as the physical variable and input signal are increased in value, the rate of increase of the output signal becomes progressively less and less. Often this compression is achieved by using a function generator with a logarithmic characteristic as shown in Fig. 3 (a) and (b). It is usually known as a logarithmic amplifier and the solid-line graph in (b) shows how its amplification or gain factor (indicated by the ratio of the output signal change to a given input signal change) alters with the value of the input signal in contrast with the gain factor of an ordinary amplifier (Part 9) which, as indicated by the chain line, is constant whatever the value of the input signal. At a later point in the communications system, the 'compressed' signal produced in such a way will probably have to be expanded, to return it to true proportionality. This would be done by a function generator with the opposite effect to that in Fig. 3 — that is giving an output signal which is an exponential function of the input signal.

Not all function generators are based on straightforward mathematical laws. Some have arbitrary functions — that is any kind of output/input relationship that the electronic engineer wishes to provide for a particular application. For example in an analogue simulator it may be necessary to introduce functions corresponding to backlash or dead-zone, as found in mechanical transmissions.

*The product \( a \cdot b \) is equal to a quarter of \( (a+b)^2 -(a-b)^2 \).*

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Fig. 1 Schematic representation of a function generator. The output variable is a function of the input variable.

Fig. 2. Graphical representation, or 'model', of the equation \( y = x^2 \), as given by the function generator in Fig. 1.

Fig. 3. Function generator with a logarithmic characteristic (a); and output/input characteristic of a logarithmic amplifier compared with that of a linear amplifier (b). A given change of input signal, \( dV \), results in different changes of output signal, depending on where it occurs on the log. curve.

*Type of analogue computer which acts as an electronic 'model' of some mechanical or other physical system.*
Solder Tests

Choice of solder for fine wires and thin copper films

by B. M. Allen*

Soft solders are among the most versatile of metal-bonding materials, mainly because of their ability to wet so many of the commonly-used metals and alloys. The wetting action is spontaneous and, for most metals, requires only the assistance of a flux and a moderate degree of heat for the joint to be completed in a few seconds. It is for this reason that the process has found almost universal application in the making of permanent electrical connections in electronic equipment. However, the action of wetting, on which the joint-making property depends, is only the first stage in a process by which the metal to be soldered is dissolved by the molten solder. The rate of penetration is not very rapid but, with the increasing use of fine wires and thin films, is becoming significant. The strength and reliability of soldered joints on such parts can be seriously affected if attention is not given to the proper choice of solder.

Previous work on the rate of solution of metals in solder has been based on the rate of wear of a soldering iron bit under simulated practical conditions, or on the reduction of diameter of wires immersed in solder. The former method is entirely empirical and the results are applicable only to the rate of wear of bits. The latter method is based on sectioning and polishing, supported by X-ray microprobe analysis of the sections.

In the investigation to be described, simple equipment was used to determine the time to breakage of lightly-loaded thin copper wire immersed in a globule of molten solder. Though results are somewhat variable and the test conditions do not at present allow of accurate calculation of a rate of solution, the method is inexpensive to set up, easy to perform, and the average of several results is sufficiently stable to show very significant differences in the performance of various solder alloys.

Method of test

The equipment, illustrated in Figs. 1 and 2, comprises an arrangement for holding a reel of fine copper wire, the wire passing over a smooth electric contact to a pulley supporting a balance pan, the weight of which is adjusted to provide a load sufficient to keep the wire taut and to break it abruptly when a certain amount of solution has taken place. The wire passes from the pulley over an iron-plated soldering bit (held at a controlled temperature) and secured to another contact. Timing by an electrically-operated seconds counter is started manually at the moment when pre-fluxed solder, applied to the bit, is seen to wet the wire, and stopped automatically when breakage of the wire interrupts the clock circuit.

Temperature control was obtained using commercial iron-plated bits incorporating a magnetic thermostat. Four temperatures were used, measured with an iron-constantan thermocouple standardized by taking the cooling curves of lead, tin and 60/40 tin/lead solder placed on the bit. The temperatures and their variation during the thermostat switching cycle were 272 ±17°C, 318 ±9°C, 370 ±7°C, and 430 ±10°C. The wires tested were soft-drawn of h. c. + copper of diameter 0.067 and 0.115mm. It was found that, even at the highest temperature used, the wires when laid across the bit without solder took at least 30 times as long to break as they did with any solder in position, so simple burning-through of the wire may be discounted. It was also found that the time to break was unaffected by the type of flux used, whether plain rosin, activated rosin or acid-salt type. Generally, activated rosin flux was used to ensure rapid wetting of the test wire by the solder. The quantity of flux used was varied from 1.5 to 3.5% of the weight of the solder, again without affecting the result. The pan weight was adjusted to 50g.

Interpretation of results

(a) Weight of solder. Tests on solders containing added copper showed that the rate of attack on the wire by the solder is reduced if copper is already present in it. The rate of attack will therefore be slowed down during the test, as the copper concentration in the solder rises, unless the weight of solder used is sufficient to make this change in concentration very small. From Fig. 3 it appears that with 0.067mm diameter wire this possibility hardly affects the measured time if the initial solder weight is over 75 mg. For most of the tests reported here, a constant volume of solder (9 cu. mm.) was used; the weight used therefore varied with the solder density.

(b) Variability of results. For a given test condition, five tests were made and the mean value taken. In most cases, the individual results did not differ from the
mean by more than $\pm 20\%$, but in some cases the longest time was double the shortest time. The cause of this variation is not known, but it seems possible that variation of temperature during the test may account for some of it, since the switching cycle (40 seconds at 272°C and 25 seconds at 430°C) is longer than some of the measured times. Another possible source of variation is in the grain structure of the copper wire; since it is likely that tin will penetrate more rapidly along grain boundaries than in other directions, grain boundaries transverse to the wire axis will cause more rapid breakage than boundaries parallel to it. Variable degrees of hardening produced in the wire when pulling it off the reel prior to lining it up for test may also have produced some variation in results.

(c) Effect of wire diameter. According to Beal\(^1\), the solution of solid metal in solder is a diffusion-controlled process and the thickness dissolved should therefore be directly proportional to the square root of the time. If timing could be continued until the wire was completely dissolved, the square root of the total time would then vary linearly with the initial wire diameter \(d\), passing through the origin and with the slope a function of the absolute temperature; 

\[
\sqrt{T} = kd
\]

Fig. 4. Relation of breakage time to wire diameter in 60/40 solder.

Table 1: Characteristics of solders tested.

<table>
<thead>
<tr>
<th>solder</th>
<th>approval or standard</th>
<th>melting range, °C</th>
<th>% tin</th>
<th>% copper</th>
<th>% silver</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pure tin</td>
<td>BS 3252. Type T2</td>
<td>232</td>
<td>99.8</td>
<td>0.01</td>
<td>—</td>
</tr>
<tr>
<td>60/40</td>
<td>BS 219, Grade K</td>
<td>183 - 188</td>
<td>59.1</td>
<td>0.005</td>
<td>—</td>
</tr>
<tr>
<td>70/30</td>
<td>BS 219, Grade V</td>
<td>183 - 277</td>
<td>15.2</td>
<td>0.005</td>
<td>—</td>
</tr>
<tr>
<td>H M P</td>
<td>BS 219, Grade S8</td>
<td>296 - 301</td>
<td>4.8</td>
<td>n.d.</td>
<td>1.63</td>
</tr>
<tr>
<td>Savbit No. 1</td>
<td>DTD. 900/4535</td>
<td>183 - 215*</td>
<td>50.3</td>
<td>1.50</td>
<td>—</td>
</tr>
</tbody>
</table>

*Registered trade mark. Multicore Solder Ltd.

As indicated by cooling curve. The temperature at which the dispersed Cu₃Sn₃ dissolves is much higher; see text.

and the temperature not being constant) complete re-solution of the Cu₃Sn₃ is not achieved below a temperature of about 350°C. Accordingly, copper-loaded solders such as Savbit continue to give protection against solution of copper up to this temperature, but at higher temperatures the rate of solution increases rather rapidly. In some applications a high-melting solder has to be used in order to permit operation of the joint at high service temperatures. Fig. 5 shows that in such cases, h.m.p. solder gives the best protection at the high soldering temperature needed. In this composition, the poor wetting qualities of a 5 tin 95 lead solder are in some degree rectified by the presence of 1.5% of silver.

Conclusion

Simple equipment has been described which permits comparison of the rate of attack of various solders on copper. The results show that for soldering temperatures up to 350°C copper-loaded solders offer the best protection; above this temperature one should use silver-bearing low-tin solders. Future work will be directed to improving the accuracy of the measurements to allow calculation of absolute solution rates, and to testing other metals, including silver and gold.

References


Fig. 5. Breakage time of 0.067mm copper wire in various solders as a function of temperature.
Circuit Ideas

Output transistor protection in class B amplifiers

The frequently used method of protecting output transistors $T_1$, $T_2$ by using the voltage developed by the output current across sensing resistors $R_e$, $R_e'$ to turn on base bypass switches $T_2$, $T_2'$, can be improved by the addition of resistors $R_1$, $R_1'$ from the bases of $T_2$, $T_2'$ to the power lines. The result is that the peak output current available is reduced as the impedance $R_1$ of the load is reduced, so that the output transistor dissipation when the amplifier is driven into a short circuit or other reduced load is kept to a value comparable to the dissipation when the amplifier is run into the specified load.

Suppose that the peak output current into the normal load is $I$, which is approximately $V/R_e$ in most cases. If the load were purely resistive then near optimum results would be obtained by resistors satisfying the equations

$$R_e = 0.6/I$$

and $R_1/R_e = R_1'/R_e$.

Since loudspeakers are not purely resistive and their impedance is usually not accurately known, and to allow the use of low tolerance resistors, it is advisable to reduce the value of $R_e$ and increase the value of $R_1$, by perhaps 20% in each case.

The circuit has been tested with the values shown. Under full drive it gave peak currents of approximately 2.6A into $R_e$ and 1.1A into a short circuit; in the former case, the clipping of the signal was due to the limitation on output current imposed by the supply voltage.

The circuit is equally suited to dual polarity or single polarity power supply amplifiers; in the latter case replace $V$ in the calculation of $R_e$ by half the supply voltage. The circuit cannot be used with class A output stages as it causes the output transistors to be cut-off for most of alternate half cycles.

It is not recommended that this circuit be regularly used with a loudspeaker of lower impedance than the amplifier was designed for, as the operation of the overload protection which would occur in this situation would cause unpleasant distortion at high volume levels.

J. R. I. Piper, Edinburgh.

Triangular waveform generator

This circuit uses three MC709 amplifiers to generate a triangular waveform, the amplitude of which may be independently varied either side of zero by adjusting the input voltages to the first two amplifiers. The frequency is determined by the 470kΩ and 1nF integrating network, and also by the amplitude of the output. With the components shown the circuit will give a 1kHz output 8V peak-to-peak symmetrically placed about zero. If the value of the 1nF capacitor is altered for another frequency the 22nF should also be altered so that the ratio between the two capacitors is about 20:1. The circuit has good linearity, and the amplitude stability is maintained over a wide temperature range.


Crystal marker identification

In a communications receiver, marker signals from a 500kHz crystal oscillator are not always easily distinguished in a welter of received signals. The circuit shows a unijunction transistor arranged to switch the supply to the crystal oscillator on and off. This gives a 'puffing' kind of identification at a frequency of 5 to 10Hz, or a slight warble when the b.f.o. is in use. The saw-tooth at the emitter of the unijunction is at high impedance; an emitter follower therefore drives the final switching transistor. Resistor $R$ may require adjustment to obtain a suitable degree of 'modulation', and a voltmeter connected between the collector of $T_3$ and earth...
Stable voltage-to-frequency converter

When a conventional free-running multivibrator is used as a voltage-to-frequency converter, the resistors in the base leads of the switching transistors are often substituted with two additional transistors used for constant current supply. The temperature variation of the emitter-to-base voltage of these transistors reduces considerably the frequency stability of the basic circuit.

A m.o.s. field-effect transistor can have a positive or a negative temperature coefficient, depending on its operating conditions. Accordingly, in a hybrid circuit containing a suitable combination of conventional and m.o.s. transistors, the temperature stability can be improved if the operating points of m.o.s.f.e.t.s are placed in the region of the positive temperature coefficient. Such a hybrid circuit for voltage-to-frequency conversion is shown. The m.o.s. transistors $T_{f100}$ are used as the constant-current sources and as the temperature compensation devices simultaneously. The proper operating conditions of these transistors can be determined from their temperature characteristics.

The experimental results regarding the temperature stability of the converter circuits are given in Table 1 for the three following cases: (a) the circuit without constant-current sources; (b) with conventional transistors BSX29 used as the constant current sources; and (c) with m.o.s. transistors F1100 used as the constant-current sources.

The temperature behaviour was checked for different values of the control voltage $V_x$ in the frequency range from 5-10kHz. The results in Table 1 are valid for $V_x$ in the middle of the range.

<table>
<thead>
<tr>
<th>Circuits specified</th>
<th>Percentage departure of frequencies in the text</th>
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</thead>
<tbody>
<tr>
<td></td>
<td>at $60^\circ$C</td>
</tr>
<tr>
<td>a</td>
<td>0.9</td>
</tr>
<tr>
<td>b</td>
<td>2.1</td>
</tr>
<tr>
<td>c</td>
<td>0.2</td>
</tr>
</tbody>
</table>

Comparing the obtained results, it is evident that the frequency stability of the hybrid circuit is considerably better than that of the conventional one. Moreover, with careful adjustment of m.o.s.f.e.t. currents for a fixed control voltage, the frequency stability of the hybrid circuit can be further improved.

Table 1

S. TESIC,
University of Belgrade,
Yugoslavia.

High-level low-distortion tone control

This circuit was developed to provide a voltage gain of about 20dB around a tone-control circuit, with the object of running a 1V nominal output to drive a low-impedance power amplifier from a 100mV source. Absolute maximum output voltage into a 1kΩ load is 15V peak, and the tone controls have an impedance of about 10kΩ in the flat position, so that the maximum additional load is limited to about 1kΩ for 3V r.m.s. at clamping.

The circuit is equivalent to a one-transistor stage but distortion is reduced by operating $T_{r2}$ in common-base mode and $T_{s2}$ is operated at an almost constant collector voltage, hence reducing 2nd harmonic distortion. Transistor $T_{r2}$ is a constant current source to allow full current into the load on the positive half cycle, which is about 5mA.

Since the input impedance is low the circuit may also be used as a virtual earth mixer with a nominal 10kHz input impedance.

The presence of $C_1$ may cause some i.f. instability and the value may need adjustment to allow the collector voltage of $T_{r2}$ to rise in a critically damped manner, especially if the supply line has a high impedance for some reason. Omission of $C_1$ and combining the 51kΩ resistors to one of 100k reduces the gain to $\sim$14dB and also reduced the dynamic tone control range. H.F. overshoot can be eliminated by either a 1nF capacitor across the 6.8kΩ resistor or an 820Ω plus 2.2nF from $T_{r1}$ collector to ground, connected in series, or both as found necessary on the 'scope. This will also depend on the previous stage response to some extent.

J. N. ELLIS,
Canterbury,
Kent.

Inverted Wien bridge oscillator

If the reactive arms of a Wien bridge network are 'turned upside down', with the output taken across the series arm, transmission is minimum at the zero-phase-shift frequency. If this inverted network is now used as a frequency-sensitive potential divider in a negative feedback path, gain is maximum at the usual tuned frequency. Broad-band positive feedback via a resistive network can then be used to obtain oscillation. The arrangement lends itself to lamp-stabilized oscillators with simple circuitry, like the example shown. Measured distortion for this circuit was 0.3% at 1kHz.

G. W. SHORT,
South Croydon,
Surrey.
High-power Audio Amplifier Design

Obtaining 100W and more from a fully-protected class-B circuit

by R. B. H. Becker

The purpose of this article is to look at the major problems of high-power amplifier design with particular reference to overload protection, and to offer solutions which are exemplified in a design for a fully-protected 100W amplifier.

A serious design restriction appears with the acknowledgment of secondary breakdown. Sometimes this is not acknowledged in short-form data. For example, an 80V, 150W transistor may be listed but reference to individual data sheets reveals that the device has been de-rated to 50W at 60V and a gain originally 25 at 1A has fallen to 15 at 3A or 5 at 10A!

To dissipate heat effectively, a power transistor has a large chip and a correspondingly large junction area. The whole device can conveniently be imagined as a number of smaller transistors connected in parallel so that they all have the same base-emitter voltage. Assuming that they are identical, the current flowing will be equally distributed between them.

Stability is ensured so long as the transistors agree amongst themselves that their base-emitter junction temperatures are identical. This is essential, because of the very large positive exponential temperature coefficient of the collector current for a given base-emitter voltage; which means that if one transistor does get hotter than the rest, then it will take a larger share of the current, making it even hotter, and at an increasing rate hotter still until it is destroyed. This description is a model of what happens in discrete areas of the large chip of a power transistor where a certain amount of non-uniformity is bound to exist to promote the effect. In any event, if it were possible to produce a completely homogeneous construction, once the overall dissipation exceeded a certain critical value the current distribution would become unstable. Hot spots on the chip would lead to the permanent destruction of the junctions — the deceased would then be said to have suffered from secondary breakdown.

Life is made even more difficult for the circuit designer by the fact that the power level at which secondary breakdown occurs is reduced considerably at large values of collector-emitter voltage. It is therefore important when selecting a transistor for a particular function, to check the operating conditions of the device in the circuit against the published curves giving the maximum safe operating region of potentially suitable devices.

A typical set of 'safe area' of operation curves is shown in Fig. 1. From the model explaining the mechanism of second breakdown it will be evident that total duration of the 'dangerous overload' condition is of significance in determining the survival of the device. This is a function of the power magnitude and the thermal resistance of the chip and thus larger safe areas are shown for narrow power pulses. This is no consolation to the audio designer who is concerned with low frequencies where the time the device spends under high dissipation conditions is relatively long, restricting operation to the d.c. area only.

It is to ensure that the output transistors operate within their specified safe area under any condition to which the output of the amplifier is subjected that the protection circuits which will be described in the latter part of this article have been developed.

Designing back from the output

When designing an amplifier it is conventional to start at the output, to consider the requirements for the final stage and select output transistors using suitability, price and availability as the main criteria.

A 100W amplifier feeding an 8Ω load will be expected to produce 80V p-p, and hence a current of 5A peak. A bridge arrangement for the output circuit has already been designed by Ian Hardcastle. This is a very effective way of using non-exotic transistors to achieve the required 80V swing and has additional advantages of a more efficiently employed power supply (50V) and no output coupling capacitors. Because the load is being driven from each end, the output stages will each see a load of half the total load impedance, i.e. 4Ω for an 8Ω load, and on drawing the load lines for the output transistors the peak dissipation is seen to be about 39W per transistor. This load line assumes that the load is purely resistive which unfortunately is far from true for a loudspeaker. Here very large reactive impedances are present as indicated by the typical impedance/frequency graph of Fig. 2. On drawing load

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**Fig. 1. Areas of safe operation for a small power transistor.**

**Fig. 2. Impedance/frequency relationship for a typical 8Ω speaker.**
lines (Fig. 3(a)) for loads of 8Ω in series with reactances the results are quite alarming as the common tangent to all the curves drawn crosses the axes at I_{max} and V_{ce} resulting in peak dissipation of about 78W per transistor. To make matters worse, the nominal 8Ω impedance specified for a speaker is not the d.c. resistance of the voice coil. This is always lower in value and usually lies in the region 5.5 to 6.5Ω.

Taking the worst case condition of 5.5Ω when the applied frequency approaches zero, the current in the load will approach 9A while at audio frequencies peak dissipation can exceed 113W. This implies that to produce a reliable amplifier with single ended output it would be necessary to use devices with power handling capabilities in excess of 226W while the required supply voltage of 100V would restrict the choice of these devices to types with exceptionally good second breakdown characteristics.

By using the bridge output system these problems are alleviated sufficiently for there to be suitable types available 'off the shelf' such as TIP35A/36A (125W) or 2N3715/91 (150W). From the point of view of safety margins the metal can devices are the better choice but there is a third option; that of using parallel pairs of lower powered plastic transistors such as TIP33A/34A (160W total).

All forms of transistor show a reducing cost-versus-time curve as production efficiency and production expertise increases and the market expands. Cost comparisons between the metal can devices, which have been with us a long time, and the TIP33A/34A, which are comparatively new, show that at the moment the prices per set are approximately equal. In the medium term, however, it is almost certain that the plastic devices will show a greater rate of price reduction hence making desirable their selection. In the short term, the plastic devices, when used as a pair, have the advantages of higher gain at high currents and a less inconvenient second breakdown region.

A redesign of the 100W circuit to use paired TIP33A/34A output pairs will now be described.

The gain of the TIP33A/34A is quoted as 20/100 at 3A but to find out what is happening to the driver transistors Tr_{3a}, in Fig. 4 we need to know the gain of the output devices Tr_{a,b}. Tr_{a,b} at different currents which is shown in Fig. 5. Transistors Tr_{r} and Tr_{3} have approximately the same collector-emitter voltage, while the collector current of Tr_{r} is approximately equal to the base current of Tr_{a,b}. Thus the peak dissipation in Tr_{r} will be equal to the dissipation in Tr_{r} divided by the h_{fe} of Tr_{r}. Peak dissipation will occur at 12.5V V_{ce} with a load current of 4.5A, i.e. 2.25A collector current per transistor at which current the typical h_{fe} is 52. Reducing this figure by 50% for worst-case transistors, the peak dissipation in Tr_{r} will consequently be about 4.3W, which is well within the limits of TIP29A/30A as these are rated at 30W.

The question remaining is what current is required to drive Tr_{r} for peak signals, i.e. 7.3A load current as will occur when there is 80V p-p being applied to a 5.5Ω load. Referring again to the typical gain/current characteristic of TIP33A/34A and applying the 50% derating to get a minimum gain, the effective gain for Tr_{r} is 13 at 7.3A load current. As a result 590mA is being demanded from Tr_{r} which would then have a typical gain of 52 and assigning it a minimum gain of 26 gives us a maximum base current of 23mA. It is unlikely that such a magnitude of current will ever be required, particularly if a matched set of transistors is used, but to allow for the possibility a current of 25mA in the constant current sink was taken as being a reasonable value. However, this does mean that the peak power dissipated in Tr_{r} will be almost 1.25W because most of our supply voltage can appear across it but the current is unchanged. Consequently, a TIP29A is required in this position as is a TIP30A for Tr_{r} which is subjected to similar conditions when there is a normal load and conditions twice as bad when overload occurs — as will be seen later.

The selection of these devices for these positions is further justified by their high gain at the currents used.

Building in protection circuits

High-power amplifiers usually have a very arduous service life, being frequently moved and re-installed. There is a very high probability of shorting the outputs by breaking the cables, connectors, etc., or of connection to unsuitably low-impedance loads occurring from time to time. The purpose of overload protection circuits is, of course, to ensure that the occasion on which one of these faults arises is not the last. As secondary circuits...
breakdown is capable of destroying a transistor within milliseconds it is essential that the device is prevented from entering the second breakdown region in any circumstances, hence the requirement is for a means of controlling the device such that this does not occur.

A number of protection arrangements are shown in Fig. 6. First, Fig. 6 (a) shows a very simple form of protection — a current limit once a certain current in the output stage is exceeded: $T_R$ turns on and removes any extra base drive from the output transistor. This circuit produces the line AB in Fig. 7 above which the operation of the output transistor cannot stray; however, for this limit to be effective, the maximum current would have to be very low. Figs. 6 (b) and 6 (c), show the use of diodes to achieve the same result. An improvement over this form of limit is the load line limit CD characteristic of the circuit in Fig. 6 (d), which works very well for protection of audio amplifiers as demonstrated by Dr. Bailey. In this circuit $R_1$ produces a voltage proportional to the emitter current of $T_R$. When this is great enough, the voltage across $R_1$ is sufficient to turn on $T_R$, removing further drive current from the base of $T_R$. Also fed to the base of $T_R$ is a current from $R_2$, proportional to the voltage between the supply line and the output line (a close approximation to $V_{CE}$ of $T_R$). In this way the emitter current of $T_R$ required to turn on $T_R$, is reduced for increasing values of $V_{CE}$ of $T_R$ and at point D no current at all is necessary for cut off.

Figs. 6 (e) and 6 (f), show developments of the load line, bending it with non-linear devices for applications where using a straight load line would result in either the maximum current permitted being exceeded or entry into the second breakdown region occurring. Linearized approximations to these curves are EFG, HJK, in Fig. 7. However, as we saw earlier, reactive loads give rise to operation in a region bounded by a straight line making logical the use of a protection line which is straight also.

Reverting to Fig. 6 (d), it can be seen that to protect the output stage, $R_1$ is required to protect each protection transistor from the otherwise large currents which would flow down it from the turned-on $T_R$ of Fig. 4, down to the shorted output line. Unfortunately, $R_1$ is in the signal path and will reduce the open loop gain of the amplifier so there is a compromise between how much current the protection transistor is allowed to sink and how much signal drop there is to be across $R_1$. This drop will also limit the voltage swing of the amplifier, when maximum power is being delivered, making undesirable a high valued $R_1$. Looking at the bottom half of the output stage the problem is already solved for us by $T_R$, being a constant-current sink. Thus the maximum current in the protection device is 25mA which leads us to the possibility of also limiting the current of $T_R$. Transistor $T_R$ always draws 25mA from $T_R$, when the upper half of the output stage is conducting hence a current limit of 50mA will restrict the current in the top protection transistor to 25mA to be symmetrical with the bottom one. The current limit can be achieved very readily by placing a small resistor in the emitter circuit of $T_R$, and two diodes, in series between the base and the supply line. When the current reaches 50mA the voltage across the resistor together with the base-emitter voltage of $T_R$ is sufficient to turn on the diodes preventing any further increase in current. Because of their well defined forward voltage drop IS 920 diodes were chosen.

In Fig. 8, which is drawn on a linear scale, the safe operating limits for paired TIP33A devices are shown with the protective limits.
tection line, which is displaced from the worst case limit of operation to ensure that premature limiting can never occur with normal use. From the intersects CD we can find suitable values for resistors $R_2$, $R_3$, $R_4$, if a value for $R_1$ is assumed, but this value is somewhat of a compromise. It needs to be large to give temperature stability of the quiescent current since, at normal temperatures, the temperature coefficient of a silicon p-n junction is about 2mV per °C so that the variation in quiescent current would be about 2mV (resistance of $R_1$ m& per °C temperature difference between the junctions of the driver transistor $T_r$, and the 'amplified diode', $T_v$. However, if its value is too large the maximum voltage swing in the output stage will be restricted and also, on peak currents, the protection transistor on the non-conducting transistor pair will be turned on in its reverse voltage mode hence producing a false overload condition. Balancing these criteria 0.2A was arrived at for the value of $R_1$, and germanium diodes which have a low forward voltage drop, were added to obstruct the reverse voltages on the protection devices. For correct operation of the protection circuitry, at point C on Fig. 8 where 10A are assumed to flow, there are 2V across $R_1$ and the protection transistor is just turning on with 600mV on its base. If we allow 10mA to flow down $R_4$ and $R_2$ their values are 140Ω and 60Ω respectively. At point D there is 600V across the parallel combination of $R_2$, $R_1$ and the current down these comes via $R_2$ which has 53.4V across it hence the value of $R_2$ is 3.74kΩ.

Adjusting the calculated values of the resistors to preferred values we now have a set of components which are shown in the circuit diagram in Fig. 9.

In this modified circuit the original system of obtaining high-frequency stability has had to be changed because the capacitors which were between the base and emitter of the pre-driver transistors will lose their effect when the diodes turn on, causing instability under overload conditions. The networks $C_{10}$, $R_{11}$, and $C_{11}$, $R_{12}$ control the closed loop gain of the amplifier resulting in a defined 3dB bandwidth of 50kHz whilst stability with reactive loads is ensured by the Zeobel networks $C_{13}$, $R_{63}$ and $C_{14}$, $R_{64}$ at the output terminals. The choke in the output circuit is not now necessary.

The rearrangement of the potentiometer on the quiescent current setting circuit was considered an advantage. This is now failsafe in that the biasing transistor will turn fully on, turning the output transistors fully off (instead of the other way round) during a loss of contact on the potentiometer. The capacitors across the transistors $T_{910}$ are a bypass for high frequencies where the conductance of the transistor is reduced. Potentiometer $R_{43}$ has also been connected in a similar manner.

This modified driver stage is then attached to the voltage amplifier and phase splitter stages of the original amplifier. The value of $R_1$ in the phase splitter stages has been altered because the current down the resistor chain supplying the base current

Fig. 9. Circuit of 100W amplifier with comprehensive overload protection.
for $T_{r9}$ and $T_{r10}$ has been increased to cater for the greater current requirement of these devices. Another change is a substitution of the diode-connected transistor for the diode in the chain. This is to facilitate the thermal interconnection of the diode and $T_{r9}$ or $T_{r10}$ to ensure constancy of current in the driver stages, with variations in the junction temperature of the current sink transistors. Also the 10kΩ resistor in the tail of the differential pair has been replaced by a transistor acting as a constant current source to achieve equality of gain for the two sides of the amplifier.

**Avoiding instability problems**

When building bridge-output or stereo amplifiers, a fairly common problem which arises is that of instability, and the answer generally comes down to excessive neatness. Although cable forms and edge connectors look very attractive, they give rise to extra resistance in the interconnections and stray interconnection between various parts of the circuit. All high-current leads must be kept as short as possible and no signal allowed to be transferred at low signal level to the next stage using, as the return path, a lead carrying current along that path to a high level stage. Also undesirable is any form of closed loop in the common line. These requirements can easily conflict as indicated in Fig. 10(a) where there is either an earth loop or else the signal return is down a current path. One way of easing the problem here is to connect up as though closing an earth loop, but inserting small-value resistors in the earth leads to the pre-amplifier or phase splitter, as in Fig. 10(b), so that the low-impedance loop is broken but supply and signal levels are not appreciably affected. For this amplifier two 10Ω resistors $R_{13}, R_{14}$, are used as shown.

When wiring up the heat sinks one attractive feature of the TIP range of transistors soon becomes apparent; that of being single-hole fixing and of flat construction simplifying drilling and enabling them to be bolted onto the inside of the sink, eliminating the usual exposure of circuitry at the back of the amplifier. Silicone grease should be used to ensure good thermal conduction. The common line problems are reduced if all four n-p-n devices are mounted on one heat sink and all four p-n-p types on the other so that the supply leads can now come directly from the power supply output capacitor. None of the other transistors needs to be mounted on a heat sink, but small tabs are desirable on $T_{r2}$ and $T_{r6}$ and also on $T_{r11}$ and $T_{r17}$, if air circulation to the printed wiring board is rather restricted. The cooling tabs may be strips of thin aluminium as displayed in the photograph of the prototype amplifier (Fig. 11). This also indicates the method of interconnecting the circuit board, heat sinks and the power supply. For increased stability of quiescent current, albeit non-critical, the junctions of the driver transistors and $T_{r_{10}}$ should be maintained at similar temperatures as can be achieved by using the flat of $T_{r9}$ to the cap of $T_{r1}$, and likewise for $T_{r_{10}}$.

**Meeting the power supply requirement**

Most regulated power supplies are prone to destruction if their outputs are shorted—as can easily happen when working on an amplifier. This shortcoming can be simply cured by adding a few extra components to form a protection circuit, similar to those in the amplifier, but using a straight line characteristic intersecting the voltage axis at a point lower than the output voltage. Without affecting normal operation, this enables the power supply (Fig. 12) to trip out and supply almost zero current when the output is shorted or otherwise reduced to a low level. No overheating can occur and the amplifier cannot be used again until the mains supply is switched off and approximately 20 seconds allowed to elapse before reconnecting it. To reduce the cost incurred by the use of a 150W regulator transistor, two TIP3055 devices have been paralleled using 0.22Ω resistors to divide the current. These transistors are 70V, 90W TO-3 compatible plastic replacements for metal can 3055s.

A final degree of protection has been added to the power supply to save the amplifier from destruction under inadequate ventilation conditions. If the amplifier is run under prolonged high dissipation conditions with an insufficient flow of cooling air to the sinks, then failure of the output devices can occur, due to either straight over dissipation or to secondary breakdown. This latter effect can still occur in spite of the protection circuits in the amplifier because the safe area of operation of the output transistors is reduced if they are operated at elevated case temperatures. To protect against disaster arising from this cause the temperature of one of the heat sinks is sensed by a miniature bead-type thermistor $T_{h}$, which is glued to it. The thermistor is part of a potential divider placed across the reference zener. When the temperature of the heat sink rises to 70°C the voltage on the potential divider rises to a point where zener diode
Fig. 12. Power supply incorporating short circuit protection and a thermal trip.

\[ D_2 \] breaks down and allows enough current to flow through it to forward bias the base-emitter junction of transistor \( T_{r15} \). Transistors \( T_{r15,15} \) form a thyristor equivalent circuit which, once turned on, latches on and collapses the reference voltage reducing the output voltage to a low level where the supply shuts down and will not start up again until the heat sinks have cooled and the mains switched off about 20 seconds. This 20 second wait is to allow \( C_{14} \) to discharge sufficiently for the power supply protection circuit to be released from its tripped out state.

**Setting up**

After building, the circuits should be carefully checked and confirmed to be correct, preferably also by an independent witness, before switching on. The power supply voltage must be set to 50V by varying the potentiometer \( R_{a4} \) before the output of the supply is connected to the amplifier. To set up the thermal cut-out temperature, \( R_{a4} \) should be set to its minimum value and the thermistor heated up in a beaker of water. When the water is at 70° C, adjust \( R_{a4} \) until the supply just trips. Remove the thermistor from the water, reset the supply by switching off the mains and then recheck the operation of the trip.

The circuit board can be tested without connecting the output devices. To do this the potentiometers should be rotated for maximum resistance, and the 50V supply taken directly to the board. After switching on, the 50kΩ potentiometer is set to give an output offset voltage as low as possible but the 2.5kΩ potentiometer must not at this stage be touched. The current taken will be about 70mA. With the output transistors in circuit one of the 2.5kΩ pots is carefully rotated until the consumed current increases by 50mA. After leaving time for the amplifier to stabilize thermally and for the current to settle to its final value, this procedure is repeated for the other potentiometer. After checking the offset voltage, the load may now be connected.

Should 100W be insufficient the use of more powerful transistors in the pre-driver stage will permit the output to be raised to 200W by the simple expedient of adding more output transistors, forming parallel quadruples, and also of halving the values of resistors \( R_{21,25,29,30} \) and \( R_{31} \) through to \( R_{a4} \). To cater for the increased base current called for by the output devices it is advantageous to substitute TIP31A/32A for \( T_{r17} \) through to \( T_{r30} \). Quirescent current of 100mA will now be necessary and high performance heat sinks essential. A suitable brand of sink is the Redpoint 6W which will readily accommodate eight transistors of the TIP variety.

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**References**


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Sixty Years Ago

February 1912. The Marconigraph carries an account of experiments performed in America by a P. Ludewig on fitting a radio receiver to a balloon. The form of the apparatus can be seen in the reproduction of the original drawing. The receiver was a Schloenmich cell which was patented in America in 1903 and is a variation of the positive point electrolytic detectors invented, independently, by Ferrié, Fessenden and Vreeland in 1900 all of which were based on work done by Wehnelt in 1899.

The 1900 inventions consisted of a fine point of platinum wire about 0.025mm in diameter. This formed the positive electrode and was dipped into a dilute solution of sulphuric or nitric acid contained in a lead cup which was the negative electrode. The device was connected in series with a pair of high-resistance phones and a variable resistor. A battery power supply was required.

Schloenmich substituted zinc for platinum so the electrolytic detector also became a cell and generated its own power supply. The device was never widely used.
Multivibrators with Seven-decade Range in Period

Astable and monostable voltage-controlled oscillators using f.e.ts and controlled current generators

by D. T. Smith*

Monostable and astable multivibrators are very useful and widely used, but in their simple form give a limited range of frequencies or periods unless components are switched. Various modifications have been suggested to increase the range, for example Bowes† has described a multivibrator whose frequency can be controlled by a voltage over a hundred-to-one range. Circuits are described here of a multivibrator with a voltage-controlled frequency, and monostable with a voltage-controlled period, each covering a range well in excess of a million to one.

Astable circuits
The basic multivibrator circuit (see Fig. 1) uses a pair of f.e.ts in a conventional manner, but with controlled current generators replacing the usual gate discharge resistors. This gives a linear discharge of the capacitors as shown in the waveforms. After switching, the capacitor must discharge by $V'$ volts before drain conduction starts and the circuit switches over. This gives an oscillation frequency for a symmetrical circuit of $f = 1/2CV'$.

To obtain a wide frequency range without changing the capacitors, $I$ must be varied over a wide range. This is achieved with bipolar transistors, using the well-known exponential variation of collector current with base-emitter voltage $I_C = I_S \exp (V_{BE}/V_T)$ (where $V_T \approx 25mV$ at room temperature and $I_S$ is a constant for a given transistor at a fixed temperature). With modern transistors this relation holds well for a range of currents typically from below one nanoamp to the milliamp region.

The full circuit uses p-n-p transistors for the current generators, with series resistors to limit the current to a safe value, Fig. 2. If a symmetrical output waveform is required these transistors should be matched for equal currents. If the transistors have different values of $I_S$, the waveform will be asymmetrical, but the mark-space ratio will remain constant as frequency is varied. The measured performance of the circuit is plotted in Fig. 3, and shows a frequency range of $10^7$ to 1.

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The range is limited at the l.f. end by leakage currents, and at the h.f. end by the time taken for the drain to switch with its capacitive load.

The h.f. performance can be extended by driving the capacitors from a low impedance (rather than direct from the drain) and emitter follower buffers can be used as shown in Fig. 4. This modification does not affect the l.f. performance, but increases the maximum frequency as indicated in Fig. 4. With the values shown, the h.f. limitation is given by the current limit set by the 2.2kΩ safety resistors.

Mono-stable circuits

The mono-stable circuit follows from the astable circuit and uses the same technique to obtain a wide range of period variation, Fig. 5. The measured variation of period with input voltage is plotted in Fig. 6, and this corresponds with the frequency variation of the multivibrator. Because temperature coefficient of the base-emitter voltage of a transistor at constant current is typically -2 to -3mV/deg C, the input must be reduced by about +2.5mV/deg C for constant period. (With a constant input, frequency will rise by about 10% per deg C).

Most silicon planar transistors and junction f.e.t.s would be suitable for these circuits. The types used were the cheapest to hand. These were 2N5172 for n-p-n transistors, 2N4061 for p-n-p and 2N3819 for the f.e.t.s.

Book Received

Colour Television Theory by G. H. Hutson is primarily intended for those who already have a working knowledge of monochrome television principles and is mainly concerned with the PAL development of the N.T.S.C. system. The contents cover the subjects of coding, transmitting and receiving, demodulation, subcarrier reference oscillators. There is also a chapter on e.h.t. systems and receiver design and development. The book contains a large number of illustrations. The aim is to give a clear understanding of the way in which colour television works, without working through complex mathematical concepts, and provide the reader, who wishes eventually to explore certain topics in greater depth, with comprehensive information on basic principles. Pp. 326. Price £3.85. McGraw-Hill Publishing Company Ltd. Maidenhead, Berks.

Signals and Information by C. C. Goodyear. The main emphasis is on signals and the ways in which they may be processed for transmission. The book is aimed at final year undergraduate level, specializing in telecommunications and other fields, such as control engineering where a knowledge of processing techniques is required. After a brief introduction to signals and signal processing, there is a detailed chapter on signal and system analysis using Fourier techniques, ranging from the simple Fourier series to Fourier transforms applied to filter response analysis. A section on modulation theory describes types of modulation, detection and the implementation of the basic relevant systems. Other chapters are devoted to probability and statistics, noise, signals with noise, information and coding and communication channels. Pp. 310. Price £4.20. Butterworth & Co. (Publishers) Ltd, 88 Kingsway, London WC2B 6AB.
Educational Television Network
Broadcast system with talkback serving colleges and industry
by Aubrey Harris

Stanford University, in Palo Alto, California, operates a scheme of postgraduate study in the disciplines of engineering, applied physics, computer science, and statistics. The Stanford Honors Co-operative Program (H.C.P.), as it is called, enables graduates from industrial, governmental, and research organizations to attend courses of instruction on a part-time day release basis.

Students enrolled under this plan are able to obtain the master's degree in two years by attending two courses per academic quarter. In the American system of higher education, in contrast to the European system, the student may choose any of the courses from a large number offered, provided that certain basic requirements are met. In the graduate programme of study at Stanford, 46 courses and six seminars are scheduled. The student is thus able to tailor his study directly to his particular educational or vocational needs.

At the present time, there are about 32 organizations sending personnel to study under the Stanford H.C.P.; some of them are within two or three miles of the University, others are over 40 miles away by road. The travelling time for students to and from their place of work to the classroom in many instances exceeds the instruction time by almost two-to-one. As an example, a typical case is as follows: in order to attend a 50-minute lecture the student leaves his place of work 45 minutes before the scheduled class starting time, and returns 45 minutes after the end of the lecture. He is away from work for two hours, 20 minutes for a lecture lasting 50 minutes. In order to cut down on such time-wasting commuting the Stanford Instructional Television Network was set up in 1969. Basically, the system is a means of transmitting by television up to four lectures, simultaneously, from studio classrooms at Stanford to remote receiving classrooms set up at students' places of work. A sense of participation by students is maintained by providing audio talkback from the remote students to the lecturer at Stanford University. This allows the students, in their remote classrooms, to ask direct questions of the lecturer and receive an immediate reply via the TV audio. The talkback system is designed so that all students, whether they are in the origination classroom or the remote classrooms, can hear the students' questions and the replies. Course assignments and homework are delivered and picked up daily by a courier.

The facilities at Stanford include three television studio classrooms, and a television studio auditorium, each with associated studio control rooms, where the lectures are originated, a master control centre and transmitter terminal. At each receiving location, a down-converter is provided, which heterodynes the received signals down to the normal v.h.f. television frequencies. The output of the down-converter is then fed to standard television sets in the receiving classrooms.

The transmission of four simultaneous TV-channels is made possible by the allocation by the U.S. Federal Communications Commission of thirty-one 6-MHz television channels, and one 4-MHz channel in the 2,500 MHz to 2,690 MHz band. The 4-MHz channel is not used for television but for audio talkback as described later (Fig. 1).

![Fig. 1. The I.T.F.S. band 2500-2690MHz is divided into thirty-one 6MHz TV channels and thirty-two 125KHz audio response channels.](image)

![Fig. 2. Approximate coverage of directional and semi-omnidirectional aerals.](image)
The area covered by the I.T.F.S. transmission channel is shown in the map (Fig. 2). This indicates the receiving location in each area and the site of these relative to the Black Mountain transmitter. Most of the reception points are within 20 miles of the transmitter, although there are three 35 miles away and one 39 miles distant. I.T.F.S. may also be used for relaying the same type of programming on a point-to-point basis for re-transmission to locations which may be out of line-of-site of the main transmitter (see Fig. 3).

The I.T.F.S. frequency band is split into interleaved groups of four channels (see Fig. 1). A station operating four transmission channels is assigned a group containing four alternate channels, for example, B1, B2, B3 and B4, which are separated from each other by 6 MHz.

The vision is amplitude modulated and the sound is f.m. with the same characteristics and vision-sound separation as the regular domestic television broadcasts— the only difference is in the carrier frequency. The down converter has been mentioned above; it is basically a carrier frequency-changer, heterodyning the 2,500MHz I.T.F.S. channels down to the v.h.f. broadcast TV frequencies (54-88 and 174-216 MHz).

This enables normal broadcast television receivers to be used in the receiving classrooms without additional electronics. The reason for using interleaved alternate channels will be apparent: on down-conversion, the output channels will also be alternate, thus reducing the problem of adjacent channel pick-up in the receiver. For any particular installation, the down-converter output channels are chosen to interleave with the broadcast channels in the area. As an example, if in a certain town, v.h.f. channels 8, 10, and 12 were in use for regular TV broadcasting, the I.T.F.S. transmissions would be converted to 7, 9, 11, and 13.

In the San Francisco-San Jose area, where Stanford is located, v.h.f. channels 2, 4, 5, 7, 9, and 11 are already in use by broadcast stations. The I.T.F.S. frequencies used for transmission are in the range 2,596 to 2,638 MHz (I.T.F.S. channels E1, E2, E3 and E4). By heterodyning with a difference frequency of 2,428 MHz, the top three channels fall into unused v.h.f. channels 8, 10 and 12. Unfortunately, the lowest channel (2,596 to 2,602 MHz) when heterodyned, falls below channel 7 into 168-174 MHz: this frequency band is one which does not appear on the tuner dial. For this reason, an extra conversion is carried out on this channel to place it into another, unused, channel — channel 3 (see Fig. 4).

One big advantage of the I.T.F.S. system, is the low cost at the receiving location. Aerial systems and down converter costs (excluding receivers and distribution equipment) are in the range of $1200-$1500. Most of the expense is at the transmitter: a four-channel I.T.F.S. transmitter plus aerial system costs between $40,000 and $50,000 ($16000-$20,000), this does not include studio or microwave link equipment. The total cost of the whole Stanford transmission and studio system was approximately $625,000 (£240,000). The system was financed by assessing each of the participating organizations for a portion of the total cost; the share paid being dependent upon the size of the organization. The additional cost of operating the instructional television system compared to normal classroom instruction is estimated to be only 25%. The operating costs of the HCP programme, including television, are covered by the fees paid by the sponsoring organization for each student enrolled. Each participating organization is responsible for providing and maintaining the equipment and classrooms at its remote receiving location.

### Studio classrooms

In order to make the studios seem like ordinary classrooms, steps were taken in their design to avoid their appearing too much like TV studios; cameras and TV control equipment are not obtrusive. Small television monitors are recessed into the students' desks, one between each two students. Of course, the students in the classroom can see the lecturer well enough, but various graphic materials, which are picked up by an overhead vidicon camera and transmitted to the remote students, can be seen also by the students in the originating classroom on the television monitors.

Associated with each monitor is a talk-back microphone and press-to-talk switch. A student wishing to ask a question of the lecturer uses the microphone, which is integrated into the audio transmission path, so that students at the remote locations may hear clearly all questions asked.

Apart from the overhead camera mentioned above, another vidicon camera is installed on the rear wall to provide a close-up view of the lecturer or of the blackboard, or to show a view of the entire classroom (see Fig. 5). Both cameras have remote controls for adjustment of pan, tilt, zoom, and focus. A control room is associated with each studio classroom. This houses the camera control, video switching, audio and intercom equipment.

The video and audio outputs from each studio classroom are fed to the master control room where the signals are routed.

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**Fig. 3. I.T.F.S. transmitter may be used for direct transmission to schools and colleges (A, B, C) or via repeater (D) to location (E) not directly in view of transmitter due to natural obstruction.**

**Fig. 4. 'Down conversion' of the I.T.F.S. channels brings the signals into the v.h.f. TV broadcast band with same spacing of bandwidth. E1 would drop out of normal TV channel; additional conversion places it in v.h.f. low-band.**
to the four-channel microwave link transmitters. The microwave link serves to relay the signals to the I.T.F.S. transmission equipment located eight miles S.W. and 2,100 feet higher than the master control building on the campus.

The microwave link from the campus to the transmission site operates in the 12-GHz band and utilizes a single 4-foot diameter paraboloidal aerial at each end for all four microwave channels. At the I.T.F.S. transmission site, the signals from the link receivers are demodulated to base-band and remodulated in the I.T.F.S. transmitters.

A triple aerial system is used to radiate the 10-W output of the transmitters. A modified omnidirectional aerial is fed with seven watts of the output power. This radiates over a 160° arc to the nearest receiving points and, with its restricted vertical radiation pattern, has a gain of about 17dB. A 10-foot diameter parabolic dish aerial fed with 1W is used to reach the San Francisco receivers (35 miles distant). This has a beam angle of about three degrees. A six-foot diameter parabolic dish radiates the remaining 2W from the transmitter towards the Berkeley-Emeryville area. These points are 39 air miles away and are within the four-degree beam width.

At each receiving location, the 2,500 MHz transmitted signal is picked up by a parabolic dish (Fig. 6) and converted to v.h.f. by the down-converter bolted nearby. An r.f. cable carries the output from the down-converter into the building for distribution to the classrooms. Where only one or two receivers are to be connected, no further amplification is necessary as the signal level is approximately 8 mV (±18 dBmV). An r.f. wideband amplifier is used to raise the r.f. level if required to feed more receivers.

In the remote classrooms standard, high-quality receivers are used modified to select talkback transmitter frequencies as discussed below.

The audio talkback system transmissions take place in the upper 4 MHz of the I.T.F.S. spectrum. Each of the 31 television channels has exclusively associated with it a 125 kHz return audio channel. At each receiving location, a 250mW f.m. audio transmitter, with a modulation passband of 300 Hz to 10 kHz, is used to relay the talkback speech signals to the television transmitter site; the signals are here demodulated and fed to the originating classroom. There is only one talkback transmitter at each television receiving location, but this can be switched to operate at the frequency associated with the television channel being used. The switching at the receiving end is done by an extra bank of switch contacts on the receiver TV channel selector control. The transmitters are keyed on, remotely, from switches mounted on the students' microphones.

Separate aerial dishes are used for the talkback transmitters and the receiver at Stanford, in order to reduce crosstalk between the television transmission and the talkback transmission; the latter aerials are cross-polarized relative to the television aerials.
Miniature Automatic Telephone Exchange

Constructing a telephone system for private use

by G. F. Goddard

Telephones designed on the standard G.P.O pattern which can be bought very cheaply are shown in a simplified form in the circuit of Fig. 1. The audio circuit consisting of a carbon microphone and magnetic earpiece (plus induction coil etc.) is completed through the line by lifting the handset from its cradle. The microphone operates by modulating the d.c. passing through it with the speech waveform and consequently modulating the line current. The bell is designed to operate on a.c. only and is isolated from d.c. in the line circuit by a capacitor. The dial operates by breaking the line circuit the number of times corresponding to the number dialled.

Thus the requirements of a telephone exchange are basically d.c. for the line, a counting system to count the dial impulses, a connecting system to join the two relevant telephones together, and an a.c. source to ring the bell.

If, however, only two telephones are to be used in a system, the requirements can be simplified.

Connecting two telephones

The telephones are wired in parallel and supplied with d.c. from a 50V supply in series with a 1kΩ resistor (R, Fig.2). When both handsets are on their cradles no current is drawn. If one handset is lifted and the dial turned to its fullest extent, the impulses produced will be transmitted to the bell of the other telephone via its built-in capacitor, and cause it to ring. Then, on lifting that handset, both d.c. circuits will be complete and speech modulation produced in the microphone of one telephone will be transmitted to the receiver of the other. The resistor drops the voltage across the telephones to about 5V while talking, but allows the capacitors to charge up to the full battery voltage for ringing. Some telephones may have three or four wires coming out. If this is the case it is necessary to remove the base of the telephone and refer to the circuit diagram and notes within. This will show which two wires are used and which to join together.

The automatic exchange

The operating sequence of the exchange is as follows:
1. The handset is lifted from the cradle of a calling telephone and d.c. passes through it to energize a relay at the exchange. This causes a uniselecter to lock onto that particular line and inactivate all the other lines in the system. When this has happened a dialling tone is connected to the calling line to indicate it is clear to dial. If the exchange was already being used by another telephone the dialling tone would not appear until the exchange was free.
2. The calling telephone dials the number of the called telephone and the impulses from the dial are counted by a second uniselecter.
3. At the completion of dialling an a.c. ringing signal is sent to the called phone and a ringing tone to the calling telephone.
4. When the called telephone is answered the d.c. that passes through it activates a relay which disconnects the ringing current and connects the telephone to the calling telephone.
5. When the handsets are returned to their cradles, the second uniselecter returns to zero and the exchange switches itself off.

Fig.1. Simplified telephone circuit.

Fig.2. Method of connecting two telephones. R should be selected to suit the battery employed as follows: for 50V use 1kΩ, for 67.5V use 1.5kΩ and for 90V use 1.8kΩ.

The switch contacts on the uniselecter, a large rotary switch rotated by an electromagnetic mechanism (see Fig.6), are mounted in semicircular banks. Each impulse to the electromagnet moves the wiper contacts round one position. Each wiper has two contacts; when one comes off the end of the bank of contacts the other makes contact at the beginning. This means that the wipers can go round and round over the contacts without any discontinuity. There is a slow release relay attached to the electromagnet, which, if wired in series with it, will cause the wiper to rotate continuously when a supply is applied.

Circuit

The complete circuit of the automatic exchange is shown in Fig.3. The relay and uniselecter coils are represented as lettered blocks, and the contacts and wipers are lettered and numbered. For example contact RLB/2 is the second contact of relay RLB.

The unit is powered by 50V d.c. from the mains via a transformer and rectifiers, but, to prevent excessive power consumption and for safety reasons the mains is switched on only when the exchange is in operation. A small battery is included as an auxiliary power supply.

Referring to Fig.4 it can be seen that when no telephones are in use there is no current being drawn from the battery. As soon as one handset is lifted a current passes through RLA and the resistor R for the particular telephone. RLA/1 operates switching on the mains and RLA/2 energizes the coil of uniselecter 1. The uniselecter coil is wired in series with its contact breaker so that the current sets the wiper in motion around its contacts. When the wiper reaches the contact to which the particular telephone is connected current passes through RLC and this stops the wiper at that point. All other telephones are therefore rendered inactive and no dialling tone will be heard if they are lifted.

In Fig.3 RLC is a fast acting relay and will respond to impulses produced by the dial. Thus, to hold the rest of the circuit operational during the impulsing, RLC/1 energizes the slow release relay RLD. RLE is another slow release relay, which is energized by RLD/1. The purpose of this relay is to sense when the dial
impulses have stopped, and then actuate the ringing circuit.

When RLE is energized RLE/1 switches the dial tone through to the calling telephone via U2a, C1, and U1 indicating that the exchange is ready for dialling pulses. In addition RLE/2 switches the coil of U2 to the negative supply, thus making it responsive to the dial pulses which also de energize RLC.

The uniselector U2 will move round one position for each dialling pulse. There are 25 positions on the uniselector enabling the telephones to have the numbers 3 to 9, 01 to 09, and 001 to 004.

RLE initially gets its supply via tank c of U3 but as the impulses send the wiper round away from the homing position, the relay remains energized only by pulses from RLC/2. As soon as the impulses stop RLE will turn off unless 0 or 00 has been dialled, in which case current will again pass through U1c before RLE has a chance to turn off.

When RLE de energizes RLE/3 energizes RLF via RLG/1, RLF/1 and RLC/1. RLF/3 switches the ringing current to the telephone being called and RLF/2 feeds the ringing tone to the calling telephone. RLF and RLG are both very slow release relays, and together with RLF they break up the ringing, producing a ring for a few seconds, etc. This ringing pattern is not the same as the G.P.O.'s, and can easily be distinguished. The timing can be altered by adjusting the tension of the spring contacts of RLF and RLG.

On lifting the called telephone d.c. passes through it and energizes RLF using winding A, and it holds itself on by energizing winding B via RLF/1. Its action is to stop the ringing signals and connect the called telephone to the calling telephone with RLJ/1. The purpose of RLF winding B is to stop the relay turning off when the called telephone's handset is replaced, as in some circumstances this would otherwise cause the ringing current to switch on again, and the bell would again ring. The final steppath pathway is shown in Fig.5.

The power supply is 50V d.c. zener stabilized. The mains transformer is switched on initially by RLA/1 via a surge limiting resistor R2. The rectified output is smoothed by C2 via R2p and stabilized by D1 and TR1. The coil of RLB forms the bias resistor of TR1, and the relay operates only when C2 has charged up to 50V, i.e. when the bias current starts rising. When RLB/1 closes R1c and R2b are short circuited (RLF/1 and RLF/2) and RLB/3 connects the power to the rest of the circuit.

RLB/2 bypasses the 45V battery by applying the 50V to it in a reverse direction from the power supply. Thus the battery is used only for the first few seconds of a call.

The dialling tone is derived from the rectifying circuit, and is the coarse half wave, rectified, unsmoothed 50Hz. The ringing tone is full-wave rectified 50Hz appearing as 100Hz. Both are passed to the calling telephone via C1 and U1.

On completion of a telephone call the exchange starts to switch itself off when the calling telephone's handset is returned to its cradle. RLC de-energizes as do RLF and RLA. A supply passes through U3c and the coil of U1, and this returns the wiper of U2 to its homing position by the action of the contact breaker. Bank b of U1 breaks the supply of the coil of RLB, which switches off the mains. RLB is a slow release relay so that the short breaks as the uniselector wiper passes between contacts, or short breaks between calls do not turn it off. U1 does not have a homing position; the wiper stays where it was left at the end of the call.

Other points
There are no telephones with the numbers one or two because they could easily be accidentally called by rapid movement of the cradle switch. Instead, dialling one or two causes the dialling tone to reappear.

The purpose of some of the contacts in the circuit still remain unexplained; they are to maintain consistent operation of the exchange however the telephones are used.

Components and construction
Details of the Post Office 3000 type relays employed are given in Table 1. The only other relay is RLC; this should be a high-speed relay with a total resistance of about 3kΩ, operating at about 24V. This may be more easily achieved by using a lower voltage and lower resistance relay with a series resistor. If RLC is not

---

**TABLE 1**

<table>
<thead>
<tr>
<th>No.</th>
<th>Coil resistance (kΩ)</th>
<th>Contacts</th>
<th>Change in voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td>RLA</td>
<td>8000</td>
<td>3</td>
<td>±1</td>
</tr>
<tr>
<td>RLB</td>
<td>8000</td>
<td>3</td>
<td>±1.5</td>
</tr>
<tr>
<td>RLD</td>
<td>4000</td>
<td>3</td>
<td>±1.5</td>
</tr>
<tr>
<td>RLE</td>
<td>4000</td>
<td>1</td>
<td>±1.5</td>
</tr>
<tr>
<td>RLF</td>
<td>4000</td>
<td>1</td>
<td>±1</td>
</tr>
<tr>
<td>RGF</td>
<td>4000</td>
<td>1</td>
<td>±1.5</td>
</tr>
<tr>
<td>RLH</td>
<td>4000</td>
<td>2</td>
<td>±1</td>
</tr>
<tr>
<td>RLJ</td>
<td>3000</td>
<td>1</td>
<td>±1</td>
</tr>
<tr>
<td>RLA</td>
<td>3000</td>
<td>1</td>
<td>±1</td>
</tr>
</tbody>
</table>

---

Fig.3. The complete exchange.
sensitive enough it will not stop $U_1$, rotating, on the other hand if it is too sensitive it will not de-energize at the end of a call. It is therefore an advantage to be able to adjust the tension of the relay contacts to give optimum sensitivity.

Uniselector $U_1$ needs one bank and $U_2$ four banks of contacts. The coils should be 75Ω and the uniselectors 50V working.

It does not matter in which order the line wires are connected to the bank of $U_1$, they need not be matched with those on $U_2$. There have been references to a homing position on $U_2$; this is for convenience, the first contact in the bank.

**Construction**

Layout is not important although one should bear in mind that relays and uniselectors are noisy components and therefore the exchange should be mounted where the noise is not going to be a nuisance. Do not use a common earth line when connecting telephones to the private exchange as a faint conversation will be heard on all telephones using the same earth line; use two wires for each telephone.

**Testing**

After completing the construction, the exchange should be tested on the bench using two or three telephones. Connect up the batteries and the mains, and lift a handset. RLA should operate, then, after a few seconds pause as $C_1$ charges, RLB will also operate. $U_1$ should rotate, then stop when RLC energizes, RLD and REL should then energize and the dialling tone should be heard. Problems may arise due to the sensitivity of RLC as mentioned earlier.

Dial the number of another telephone. $U_2$ should rotate, then REL should break and RLF, RLG and RLH should jump in. The called telephone should ring. The tension in the spring contacts of RLF and RLG should be very low so that their slow release action is made as long as a few seconds each.

RLF, RLG, and RLH will continue clicking, as they break up the ringing signal, until the called telephone is answered. RLJ should then operate and RLF, RLG and RLH should cease operating and speech should be possible between the two telephones.

When the handsets are replaced $U_2$ should return to its homing position and everything else should switch off which will be indicated by the neon.

It may arise that one or two of the relays buzz instead of operating properly. This may be caused by a contact which is acting too early or too late relative to the rest of the system. The problem can be solved by bending the contact nearer or further away, as the case may be. Note that change-over contacts should 'break before make' to prevent this occurring.

The frequency of the ringing current used is 50Hz a.c. This is higher than normal so it is necessary to move the two gongs of the bells closer together to make the bells ring smoothly. This is easily done by loosening the gong-retaining bolts and rotating the gongs, and then re-tightening the bolts.

Editor's Note: It is illegal to connect any private equipment such as described here to the public telephone system without Post Office approval.
About People

Bernard J. O’Kane, Ph.D., B.Eng., M.I.E.E., chairman of Marconi-Elliott Avionic Systems Ltd., who has been serving as vice-chairman of the council of the Electronic Engineering Association for the current year, has been elected president of the Association in succession to W. R. R. Holmes, formerly of the Piccsy Company, who has resigned from the industry. Dr. O’Kane, who is 59 and a graduate of Liverpool University, joined the Research Laboratories of the General Electric Company where he served until 1941 when he was transferred on loan to the Telecommunications Establishment at Malvern. He rejoined the G.E.C. Radio Department in 1945. During 1947 Dr. O’Kane moved to Interna
tional Aerial Ltd. as chief engineer and in 1952 joined the Marconi Company as chief air radio engineer. He became managing director of Marconi-Elliott Avionic Systems Ltd in 1969 and is now chairman.

E. T. Goodwin, Ph.D., has been appointed a deputy director at the National Physical Laboratory. He will have executive responsibility for the Engineering Sciences Group, which comprises those divisions which are concerned with the hydrodynamics of ships and other marine vehicles and their operational characteristics: computing and computer applications; engineering, building, and acoustics. Dr. Goodwin graduated in mathematics at Cambridge in 1934, and obtained his doctorate for research in theoretical physics. He became a lecturer in applied mathematics at the University of Sheffield in 1937, but at the outbreak of the war returned to Cambridge, where he worked for the Ministry of Supply until 1945. After 18 months with the Admiralty Signal Establishment at Witley, Surrey, he transferred to Bath to supervise the work of the Admiralty Computing Service. His duties and staff of this group were transferred to the National Physical Laboratory in 1945, and in 1951 Dr. Goodwin became superintendent of this Division of Numerical Analysis and Computing.

Captain Roger L. Cloyd, F.I.E.E., divisional secretary, responsible for broadcasting, computers, industrial electronics and maritime matters in the Electronic Engineering Association, has retired. He joined the E.E.A. in 1965 after 40 years’ service in the Royal Navy. During the major part of his service in the R.N. he specialized as an electrical/electronics officer. He was at one time Fleet Electrical Engineer with the Nato Naval Forces and then controller of the Naval Air Electrical School at Lee-on-Solent. Capt. Cloyd subsequently held the post of Director of Naval Electrical Training and then the Director of Radio Equipment at the Admiralty.

I. G. Philpott, who graduated in the electrical sciences tripos at the University of Cambridge last year, has been awarded a three-year B.B.C. Research Scholarship to undertake research in the Department of Engineering at the University of Cambridge, under the supervision of Professor P. S. Brandon, M.A. The subject of Mr. Philpott’s research will be ‘ways of reducing the channel capacity required by a television signal, or improving the quality of a television image within a given channel capacity, by the use of digital electronic techniques’. The Engineering Division of the B.B.C. maintains six such Research Scholars at United Kingdom Universities.

The title ‘apprentice of the year’ has been awarded by E.M.I. to the top apprentice in each of three grades — graduate, technician and craftsman. In the graduate category the award was won by Christopher Haynes, B.Sc., aged 23, who is in the chief scientist’s department of the Radar & Equipment Division and has just completed a year on Birmingham University’s Bosworth Master of Science Course in radio and radar. Best technician apprentice was 20-year-old Richard Moon who works at the Wells, Somerset, factory of EMI’s Systems & Weapons Division, where he has been specializing in production engineering during his C.B. & Guilds Electrical Technicians course. Martin Sanders, aged 20, received the award for top craftsman apprentice. He is working in the model shops of the Radar & Equipment Division at Hayes.

Percy Allaway, director of EMI Ltd and chairman of EMI Electronics Ltd has been elected chairman of the National Council for Quality and Reliability. He is also chairman of the newly formed Defence Industries Quality Assurance Panel.

M. T. Ranade, B.Sc., M.I.E.R.E., recently returned to the London office of the I.E.R.E. after completing two years' tour of duty as the technical officer and editor of the I.E.R.E.-I.E.R.E. Proceedings, India, and K. B. Kapasi, M.Sc., has joined the I.E.R.E. office in Bangalore as his successor. Mr. Kapasi graduated in 1959 from Delhi University and obtained his M.Sc. degree in 1961. In 1962 he joined the Council of Scientific and Industrial Research and in 1966 took up appointment with the Institution of Telecommunications Engineers (India) and was the editor of the Institution’s journals.

Dolby Laboratories have announced the appointment of Elmar Stetter as applications engineer who will be concerned primarily with users of the Dolby System in recording, broadcasting, and motion pictures. Mr. Stetter, who will be based at the Dolby office in London, was trained in communications technology at the Technical University of Stuttgart, specializing in electronics and audio frequency electronics where he obtained his Diplom-Ingenieur. He comes to Dolby from the headquarters of BASF in Ludwigshafen, Germany.

Newmarket Transistors Ltd have announced the appointment of Malcolm Singer as field force manager in the physics graduate of London University. Mr. Singer joined NT as a sales engineer in 1970 having spent five years with G.E.C. at Stanmore in their aero-space and guided weapons systems division, working on research and development of circuit and systems design.

Jim Crossley, who joined Dynamic Technology Ltd of Wembley, Middx. last year as director of engineering, after many years as head of production engineering for London Weekend Television, has been appointed managing director. The company, which is concerned with control equipment for lighting, has also appointed Stan Giddings sales and marketing director and overall general manager.

Michael Cook is appointed sales manager of Davian (Instruments) Ltd, the Luton manufacturers of digital logic equipment. He was previously a sales engineer with Techmation Ltd.

NEW YEAR HONOURS

Recipients of honours in the New Year List include the following: C.B.E.


O.B.E.


Sing-yui King, J.P., Ph.D., B.Sc., F.I.E.E., professor of electrical engineering, University of Hong Kong.

C. E. Tate, F.I.E.E., director of research, Pesley Telecommunications Co. Ltd.


Commander G. W. K. Whitaker, M.I.E.R.E., Royal Navy, M.B.E.


C. H. Wolff, B.Sc., M.I.E.E., head of data handling division, Pesley Telecommunications Research Ltd.

British Empire Medal

A. J. Ashenden, assistant (equipment). Television Outside Broadcasts, B.B.C.

C. E. Holmes, superintendent, manufacturing department, Guided Weapons Division, British Aircraft Corporation.

D. C. Stanley, technical officer. Defence Communications Centre.

OBITUARY

David Sarnoff, for many years chairman and chief executive of the Radio Corporation of America, died on December 12th at the age of 80. Born near Minsk, Russia, he went to the United States in 1900 becoming a member of the staff of the Marconi Company of America and was, incidentally, for a time our New York correspondent. When in 1919 the R.C.A. was formed, in which the American Marconi Company was absorbed, Sarnoff was appointed commercial manager. In 1926 he set up the National Broadcasting Company as a subsidiary of R.C.A. to establish a national network of broadcasting stations. During World War II he served as special consultant on communications on Gen. Eisenhower’s SHAPE headquarters in Europe.

Ernest Milner, market development director of A. B. Electronic Components Ltd of Abercynon, Glamorgan, died recently after a long illness. He was 45. Mr. Milner joined the company in 1959 as chief development engineer — switches.
More light on transsequatorial propagation

Recent observations by the Australian Ionospheric Prediction Service (see *Nature Physical Science*, November 22, 1971) on paths between Japan / Okinawa and Queensland, Australia, have strongly supported the view, frequently expressed in the past by amateurs, that transsequatorial propagation involves chordal hop or super-mode paths without intermediate ground reflection points even on paths over 5000 km long, and correlates fairly closely with the existence of range-spreading (Spread-F) conditions over the tropics. A number of amateur v.h.f. beacon stations should help to make possible more regular observations on this interesting form of F-layer propagation which has been shown to extend to beyond 70 MHz, to occur most frequently at the equinoxes, and apparently to include two main forms — one in the afternoons at high signal levels, and the other (with fading) in the late evenings. Unfortunately, the U.K. is too far north for frequent transsequatorial openings, but it can be observed occasionally, particularly on 28 MHz, in the second form.

Among the useful beacons for t.e.p. work are GB3SX at Crowborough, Sussex, now operating regularly on 28.185 MHz and on 70.669 MHz (on the latter frequency the aerials point north to detect auroral openings); ZSSVHF near Johannesburg on 50.1 MHz from Friday evenings to Monday mornings; and, from about February, a new beacon ZC4VHF on Cyprus which plans to radiate 20 watts on 50.5 MHz using an omni-directional vertical aerial.

Over 17,000 U.K. amateurs

The number of British amateur licences has now risen to over 20,400 — representing over 17,000 individuals if the 3200 mobile licences are excluded. This compares with some 10,500 licences (9500 plus 1000 mobile) ten years ago. The impressive growth rate undoubtedly owes much of its impetus to the introduction of the Class B (v.h.f-only, phone-only) licence which now accounts for some 3000 (17%) of the total amateurs. A further large increase could result from the present investigation by the Ministry of Posts & Telecommunications into the issue of some form of novice licence in the U.K. All the evidence from Japan (see below) and elsewhere suggests that there could well be 100,000 or so people in the U.K. who would like to participate in amateur radio were it not for the examinations — on the other hand, there can be little doubt that where licence qualifications are relaxed the turn-over of amateurs increases to a very marked extent.

But many amateurs will have learned with regret that the Post Office charge for conducting the Morse Test has again been increased. After many years at ten shillings (50p) the cost went up in October 1970 to £1. Now, from January 1st, it is £2, and the Post Office says the charge will be reviewed again in 1972/73. Since some candidates have to travel considerable distances to take the examination, the new charges could pose a real deterrent to youngsters wishing to take out a Class A licence. One would have thought — as clearly many Eastern European countries think — that the existence of a body of experienced radio telegraphists would have formed a useful national asset. One can appreciate that the Post Office finds responsibility for the Morse Test burdensome — but would it not be possible for M.P.T. to seek alternatives? Councils (other than 14MHz) and societies can conduct such tests under the aegis of the local education authorities?

Amateur radio in Japan

The tremendous increase in the number of amateurs in Japan is underlined by a report in *I.A.R.U. Region I News*. The total number of people now licensed as amateur operators has reached 256,655 although only about 54% of these actually hold callsigns and have their own stations. First and second class operator licences (corresponding roughly to our own Class A licences) represent less than 5% of the total. Curiously enough the Japanese phone-only licence (issued to some 86% of amateurs) permits operation on all h.f. and v.h.f. bands (other than 14 MHz) with a 10W limit, and would thus seem to contravene the international radio regulations. For the low-power licences, it is possible to qualify simply by taking a 40-hour training course conducted by the Japanese national society, J.A.R.U., without formal examination. Two bands are severely restricted in Japan: 160-metres comprises only 1907.3 to 1912.35 MHz and 80-metres is limited to 3500 to 3575 MHz.

The problems of s.s.b.

The popularization of s.s.b. — to the virtual exclusion of a.m. on some h.f. bands and now increasingly used on v.h.f. — has been a major feature of amateur operation in recent years. But s.s.b. brings its own problems, including the need for a high order of linearity throughout the system, both in transmitters and receivers. In practice, ‘flat-topping’ of power amplifiers too often results in excessive bandwidth which cannot always be readily distinguished from the effects of non-linearity in the receiver. A useful test is to be able to aurally compare two transmitters who would have formed part of the second form. The Bangor & District Amateur Radio Society has recently re-activated the GB3GI 145.96 MHz beacon from a new location near Salisbury in County Down (XO 41J) with 8 watts output into two four-element beams, one beaming east of north, the other to the south east, with automatic switching between beams.

In brief

The Bangor & District Amateur Radio Society has recently re-activated the GB3GI 145.96 MHz beacon from a new location near Salisbury in County Down (XO 41J) with 8 watts output into two four-element beams, one beaming east of north, the other to the south east, with automatic switching between beams. Reports would be appreciated (Eric Sandys, G1ZFH, 25 Moira Park, Bangor, County Down). . . . Believed to be the first ‘Top Band’ contact between the U.K. and Hong Kong was made by A. R. France, (G3WRF) and VS6DO at 23.00 G.M.T. on November 18. . . . The 1803kHz transmissions of 8P6DR, Barbados, have been heard in the Netherlands . . . Microwave activity periods have been proposed for March 25-26, April 29-30, May 27-28, July 29-30 and August 26-27. . . . There are now 37 member societies in the I.A.R.U. Region 1 Division, latest recruit is the Liberian Radio Amateur Association . . . An international slow-scan TV contest is planned for February 5th and 13th . . . Brigadier A. J. Jackson, Director of Telecommunications (Army), has become president of the Royal Signals Amateur Radio Society — the current membership list includes 635 licensed amateurs.

** PAT HAWKER, G3VA **
Random noise generator

Gaussian random noise generators type NS110, from ADM Electronics (a division of Siliconix), operate from a 9V (12mA) supply, and provide an output of typically 150mV r.m.s. Spectral uniformity is ±1dB (model NS110S) and ±3dB (model NS110G) over the range 60-300Hz, ±5dB and ±10dB respectively over 20Hz-3MHz. Asymmetry is better than 15%. The output amplifier presents an output impedance of 600Ω. The module measures approximately $2 \times 1 \times 0.62$in deep $(51 \times 29 \times 16$mm). ADM Electronics, Sketty, Swansea, SA2 8BA.
WW340 for further details.

Marine Lincompex radio equipment

Marine Lincompex speech-processor TM20 from STC replaces the TM1B system for shipboard use. Both transmit and receive circuits are contained in a single case with built-in loudspeaker, and provision is made for a remote loudspeaker and telephone. There are models for desk and bulkhead mounting (which incorporate vibration dampers), and a bay mounting version is also available. The equipment can be used in the radio room in conjunction with an existing radio transmitter and receiver provided these conform to certain standards which include adequate short-term stability. An in/out switch is incorporated in the equipment which enables the operator to revert to conventional radio communication. Standard Telephones and Cables Ltd, 190 Strand, London WC2R 1DU.
WW332 for further details.

Microwave semiconductors

Three Toshiba microwave semiconductors are available from Erie Electronics. The first product, an n-p-n silicon bipolar transistor type S1072 (illustrated), is designed primarily for small-signal low-noise microwave amplifier applications. At 4GHz it has a noise figure of typically 5dB and 3dB at 2GHz. The insert power gain at 2GHz is typically 8dB. The second product is a Gunn oscillator type E4054A. It replaces the 2K25 reflex klystron which has been in wide use as the local oscillator in marine radar receivers. The E4054A is also suitable as a local oscillator in microwave receivers, an a pump source in parametric amplifiers, in transponders and test equipment. Frequency is adjustable, using a single tuning screw, from 8.5 to 9.6GHz and is 70kHz voltage tunable with a tuning sensitivity of 2MHz/V. Power output is about 40mW for the entire frequency range. The third product, a Gunn diode type S3020 is for use primarily in microwave power generation. Output of up to 200mW at X-band is obtainable, with a suitable cavity, in the frequency range 8.2 to 12.4GHz. Erie Electronics Ltd, South Denes, Great Yarmouth, Norfolk.
WW337 for further details (transistor)
WW338 for further details (Gunn oscillator)
WW339 for further details (Gunn diode).

5W regulator diode

An axial-lead 5W sub-miniature plastic-encapsulated voltage regulator diode has been introduced by I.R. The new series, designated 5Z5 (1N5333A to 1N5778A JEDEC code), is available in selected voltage grades between 3.3 and 100V in both ±5% or ±10% tolerances. Operation is possible in the range —65 to +120°C. Size: 9mm long, 4mm diameter. International Rectifier, Hurst Green, Oxted, Surrey.
WW335 for further details.

Solid-state suppressors

Six families of solid-state devices designed for the suppression of large electrical transients have been announced by Mullard; they will give protection against transients up to 100kW for 1mS. The extreme stand-off voltage values for each range are shown in the table below; intermediate values correspond to the E24 series.

<table>
<thead>
<tr>
<th>Type</th>
<th>Stand-off voltage range (for 1mS)</th>
<th>Dissipation rating</th>
</tr>
</thead>
<tbody>
<tr>
<td>BZW70</td>
<td>5.6 to 62V</td>
<td>400W</td>
</tr>
<tr>
<td>BZW86</td>
<td>7.5 to 62V</td>
<td>12kW</td>
</tr>
<tr>
<td>BZW91</td>
<td>5.6 to 62V</td>
<td>5kW</td>
</tr>
<tr>
<td>BZW93</td>
<td>5.6 to 62V</td>
<td>450W</td>
</tr>
<tr>
<td>BZW95</td>
<td>7.5 to 62V</td>
<td>450W</td>
</tr>
<tr>
<td>BZW96</td>
<td>3.9 to 7.5V</td>
<td>90W</td>
</tr>
</tbody>
</table>

The use of these devices to suppress transients is described in an article published in Mullard Technical Communications. Entitled 'Transient voltage suppres-

Miniature vidicon tube

A miniature vidicon television camera tube type 20PE13, announced by Mullard, has a resolution of 25l lines. Electromagnetic focusing and scanning are used. When subject to an illumination of 10 lux from a lamp with a colour temperature of 2854°K the tube provides a signal current of 150nA. Its dark current is 20nA. External measure-


**Polypropylene capacitors**

A range of polypropylene capacitors from Steatite Insulations has been designed specifically to overcome the soldering problems often experienced with polystyrene capacitors. The capacitors, type KP1834, have a temperature range of -25 to +85°C. They are available in values ranging from 100pF to 0.033μF and with tolerances of 10%, 5% and 2.5%. Ratings are 63, 160 and 630V. The lead spacing on all values is 5mm. Steatite Insulations Ltd, Hagley House, Hagley Road, Edgbaston, Birmingham B16 8QW. WW336 for further details

**Voltage dependent resistor**

Steatite Insulations have introduced a new voltage dependent resistor, with a performance claimed to equal that of a zener diode. The ZNR varistor is composed principally of zinc oxide, and is supplied as an encapsulated disc 25mm diameter, with axial leads. The device can operate within the temperature range -25 to +70°C without derating, and up to 120°C with derating. Units can be supplied in the voltage range 22-330V. Steatite Insulations Ltd, Hagley House, Hagley Road, Birmingham B16 8QW. WW302 for further details

**Rugged miniature oscilloscope**

The Tektronix type 211 solid state miniature oscilloscope has double insulation, and is designed for use in severe environments. It will operate from a.c. mains or for up to 5 hours on internal batteries. Bandwidth is 500kHz, and deflection factors are 1mV/div to 50V/div. Sweep rates extend to 1μs/div. When operated from batteries the input may be raised to 700V above ground and to 250V r.m.s. above ground on a.c. Deflection factors and sweep rates are clearly indicated on the front panel, where they are related easily to the c.r.t. display. There is a single trigger control. Internal and external trigger circuits provide stable displays from about 7Hz to at least 500kHz. Turning the control clockwise causes the scope to trigger on the positive slope of the triggering waveform. Rotating the control further clockwise causes the 'scope to trigger on the negative slope of the triggering waveform. The type 211 is equipped with an integral flip stand which tilts the 'scope to a convenient viewing angle for bench-top operation and the integral probe and power cable wrap around a recessed area in the case. Size is 76 x 133 x 229mm, Weight 1.36kg. Price £262.00 including batteries; duty £29.60. Tektronix U.K. Ltd, P.O. Box 69, Harpenden, Herts. WW303 for further details

**Slider potentiometers**

A range of slider potentiometers with either single or double tracks is announced by ITT. Three standard slide lengths are available (30, 45 and 60mm) and track resistances range from 470 to 2.2MΩ in linear, log, reverse log and 'stereo balance' laws.

Various mountings are provided including printed circuit or direct wiring lugs. Dust covers and a range of knobs are also available. ITT Components Group Europe, Resistor Product Sales, Edinburgh Way, Harlow, Essex. WW304 for further details

**Fast recovery silicon diodes**

The Semiconductor Division of Westinghouse have introduced two new fast recovery silicon diodes. Both are low-cost plastic encapsulated types available in the voltage range 100-500V (\(V_{RRM}\)) with 200ns reverse recovery time. Current ratings are up to 1.5A for the SFxM1 type and up to 3A for the SFxM3. Prices: SF1M1 (100V), £17.50 per 100; SF1M3 (100V), £28.34 per 100. Westinghouse Brake and Signal Co. Ltd, 82 York Way, King's Cross, London N1 9AJ. WW310 for further details

**Micro wave signal generator**

The type 830 pulsed signal source from Microtest will generate pulsed microwave power with an output of 15W peak. The unit contains a pulse driver and modulator,
Wireless World, February 1972

highly stabilized power supplies, controls and a magnetron r.f. head tunable over the band 9.05 to 9.60GHz. It requires 115/230 a.c. mains input and a 50Ω r.f. load at the output. An external signal can be used to synchronize the internal pulse generator and/or an internal trigger generator output is available for synchronization of external instruments such as oscilloscopes.

**Specification:**
- Frequency range: 9.05 to 9.60GHz
- Frequency temperature coefficient: -1MHz per °C
- Power output: 15W min (peak)
- Pulse rate: 1 to 1000 p.s.s. in three ranges, continuously variable
- Pulse width: 1 to 100μs in two ranges, continuously variable
- Rise and fall times: 100ns
- Trigger output: 3V min. positive pulse at leading edge of main pulse, 1 to 10μs long.
- External sync.: 3V min. to 20V max. positive or negative pulses, 1 to 10μs long.
- Microtest Ltd, 28 Walker Lines Industrial Estate, Bodmin, Cornwall. WW301 for further details.

**Soldering iron with temperature control**
A solid-state variable-power supply controls the temperature of the bit of the Litesold ETC/I lightweight soldering iron. The operating temperature can be set within the range 150-400°C. The plug-in soldering iron contains no control components except the temperature sensor. A range of bit sizes is available. Light Soldering Developments Ltd, 28 Sydenham Road, Croydon CR9 2LL, Surrey. WW308 for further details.

**Tuning indication instrument**
Tuning indicator, type TIU 10, developed by Plessey Instrumentation, augments the tuning facilities of the Plessey demodulator TFS 50, which provides facilities for converting frequency shift a.f. or i.f. signals from a communications receiver to ω.c. signals for direct operation of, for example, teleprinters. The tuning indicator is operated via isolating buffer stages from the output of the discriminating filters. The signals are passed through amplifying and rectifying stages to provide a 'V' display for signal frequency shift signals transmission or an 'X' display for double transmission shift signals transmission on a 3in c.r.t. The unit is designed to mount in 19in international racks, with a panel height of 3.5in and a depth of 10.5in. Plessey Instrumentation, Sopers Lane, Poole, Dorset. WW307 for further details.

**Dual-in-line package sockets**
Sockets are available from Viking Industries for 14- and 16-pin dual-in-line package devices. The sockets are for wire-wrap and dip-solder applications, with a choice of three platings. Contacts are phosphor bronze and insulators glass-fibre-reinforced nylon. Large front lead-in chamfers on the entry surface of the sockets provide automatic self-straightening of leads. The phosphor bronze terminations are at 0.100in centres in two rows 0.300in apart, and 0.600in long for multiple wire wrapping. Viking Industries (UK) Ltd, Barton Industrial Estate, Faldo Road, Barton-le-Clay, Beds. WW306 for further details.

**Miniature high-voltage rectifier**
A range of miniature high-voltage cartridges from International Rectifier offers a large voltage-to-size ratio. The 5AV series covers the voltage range of 2 to 6.5kV ($V_{RRM}$), is rated at 50mA at 25°C, and measures 10 × 5mm. The voltage range is continued above 6kV through to 13kV ($V_{RRM}$) in a larger package size, 40 × 8mm. This series is designated 48V and is rated at 40mA at 25°C. Both sizes of cartridges are capable of operating at up to 20kHz into a resistive inductive load with an operating temperature range of −40 to +75°C. International Rectifier, Hurst Green, Oxted, Surrey. WW309 for further details.

**Ceramic disc capacitors**
A range of ceramic disc capacitors, type CSC, from Seatronics (UK) range between 1000pF and 5000pF (E6 series) with ±2% tolerance. The dielectric strength is 250% of the rated working voltage.

The capacitors may be used at temperatures from −55 to +85°C. Tan delta is 0.02 maximum. Seatronics (UK) Ltd, 22-25 Finsbury Square, London EC2A 1DT. WW323 for further details.

**Small power operational amplifier**
Raytheon Semiconductor have introduced an operational amplifier, type RM4132, which has a power consumption of 1mW at 20V and has typical current drain range of 22 to 35µA. Input offset current is reduced to 0.3nA typical, 2nA maximum. It maintains a unity gain bandwidth of 150kHz. The device is pin-
compatible with the RM709, 741 and 4131 and has high common-mode and differential-voltage range, and 20MΩ input impedance (typically). The units are supplied in d.i.l., TO-5 and flat packages. Eurosem International Ltd, Haywood House, 64 High Street, Pinner, Middx. WWW329 for further details

Large-screen oscilloscope
An automatic brightness control, which enables the timebase range to be changed from 1ms to 30s without adjustment to the brightness control, is provided on the large screen display oscilloscope type 383-2 from Racal Instruments. Up to four orange traces can be displayed simultaneously on the 43cm (17in) tube and each has individual Y-shift and gain controls. The display is visible up to a distance of fifty feet under normal lighting conditions. The input amplifier has a bandwidth of d.c. to 10kHz. Remote control facilities are available as an optional extra. Racal Instruments Ltd, Duke Street, Windsor, Berks. WWW324 for further details

Solid-state high-voltage modules
Brandenburg are producing high-voltage power supply modules with outputs from 7 to 15kV at up to 500μA. Two basic styles are available: model 576 with an unregulated output, and model 776A with regulation better than 0.05%. Brandenburg Ltd, 939 London Road, Thornton Heath, Surrey, CR4 6JE. WWW318 for further details

Logic power-driver i.c.
Motorola Semiconductors have introduced type MCH 2890 integrated circuit dual logic output power driver. Its function is to boost the outputs of d.t.i. and t.t.i. i.c.s to drive paper-tape punches, relays, step motors, lamps and other loads which are digitally controlled but require high current drive pulses. The i.c. is a hybrid construction of a monolithic dual two-input t.t.i. AND gate and a pair of Darlington power transistors. With normal d.t.i. and t.t.i. voltage level inputs, it can provide 6A output on a 10% duty cycle with 25ms pulse width. The continuous output current rating is 1A maximum. To be able to drive inductive loads at high current, the output Darlington transistors are each rated for a minimum collector-to-emitter breakdown voltage of 120V. The maximum rating of the collector-to-emitter voltage at saturation is 2.5V at a collector current of 6A. A 10-pin aluminium casing, similar to the commonly-used TO-3 style, has been designed for the MCH 2890 to give it a high power dissipation, its thermal resistance being 0°C/W. Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middx. WWW316 for further details

Thumbwheel switch
The EECo 8000 series thumbwheel switch from Plessey is the result of a licensing arrangement with EECo of the U.S.A. The switch is supplied in kit form, the parts being assembled directly on to the user's printed circuit board, which will incorporate the switching pattern as part of the overall printed circuit design. Assembly is carried out in seconds, without soldering or wiring leads. The switches are less than 1in high, and less than 1in wide. Six standard colours are available: black, red, yellow, green, blue and orange. Sealed and unsealed versions are available. Service life is better than 1,000,000 detent operations. Fifteen standard codes are described in a three-page publication (No. TIS22053) which includes specifications and applications information. Plessey Switching Controls, Abbey Works, Tichfield, Fareham, Hants. WWW325 for further details (switch) WWW326 for further details (publication)

Pulse generator
A general purpose pulse generator, type TF2025, from Marconi Instruments, is capable of delivering positive or negative double-pulse waveforms, either delayed or undelayed, at any repetition frequency from 0.2Hz to 25MHz. The pulse width is variable from 10ns to 1s and the rise time is 5ns. Pulse duration is continuously variable and the output pulse amplitude is adjustable between zero and 10V. Internal or external triggering may be used, the internal triggering being produced from a source frequency adjustable over the instrument's full repetition range. External triggering can be effected in the form of a sine, square or pulse waveform from d.c. to 30MHz. The instrument may also be used manually in a single shot mode. Price £215. Marconi Instruments Ltd, St. Albans, Herts. WWW315 for further details

Fail-safe 1kW transmitter
The Racal TTA 1860 solid-state 1kW h.f. transmitter terminal employs two 500W linear amplifiers, each with its own independent power supply and cooling unit. Each 500W amplifier is made up of four 125W plug-in modules which can be removed from the front while the equipment is in operation. The modules are based on

www.americanradiohistory.com
the design of the TA940 linear amplifier. The MA1720 exciter unit used in the
transmitter gives coverage from 1.6 to 30MHz in 100 steps. It is equipped for
operation in s.s.b., c.w. and a.m. modes with options for i.s.b. and r.a.t.t. Facilities
are also available for extended and remote operation, and it is capable of fast frequency
changes with a lock time of under 10ms. The overall height of the complete trans-
mittter terminal, including amplifier, exciter and feeder matching unit, is under 153cm.
Racial Communications Ltd, Western Road, Bracknell, Berks. RG12 1RG.

Add-on r.f. a.m. amplifier
A new solid-state amplifier from GEC Mobile Radio will raise the power of
a low-power a.m. radiotelephone to 80W 'without distortion'. No modifications to
the mobile radiotelephone are required. The unit has been approved by the Ministry of
Posts and Telecommunications for use with the GEC RC660 TR and RC665 TR
for mid-band and commercial bands. Marconi Communication Systems Ltd, Marconi House, Chelmsford CM1 1PL, Essex.

Plugging power-supply regulators
Roband have launched a range of card regulators which give a stable d.c. supply
from a 'raw' d.c. source. Metal covers which surround the plug-in card protect
the components and act as heat sinks. The Rokard range gives 3 to 8V or 6 to
30V positive or negative at 1A, 2.5A, 5A or 10A. Twin regulator cards give
±4 to ±25V. Each card is fully protected against excess current and all have provi-
sion for the Roband EPO over-voltage protection module. Except for the 10A and
twin units, the size of the Rokards (105 ×
121 × 41mm) is no greater than existing cards having much lower power capabili-
ties. The p.c.b. track is gold plated. Prices range from £8 for 1A to £15 for 10A. Twin
units are £10.25 (25W) and £21 (60W). Roband Electronics Ltd, Charlwood
Works, Charlwood, Horley, Surrey.

High-stability voltage reference units
Semitron have introduced a range of units designed to provide a very stable voltage
reference from unregulated inputs over extended periods of time and a wide range
of temperatures. The reference units have their own built-in power supply which is
temperature compensated. The result is stability of ±0.005% and temperature
coefficients as low as ±0.0005% per °C. Voltage regulation is 0.002% for an input
variation from 12 to 50V d.c. Units up to
100V are available on special request. All
types are available in positive or negative
outputs. Encapsulated in an epoxy package
measuring 25.5 × 25.5 × 13mm. Semitron
Ltd, Creekside, Wilts.

Fourteen-track tape recorder
An f.m.-anologue tape recorder designated the Ana-Log 14, and available from Pye
TVT, will record up to 14 channels of
10kHz bandwidth information. The design
is similar to that of the seven track Ana-Log
7 introduced about 3 years ago. There are
two switch selected tape speeds — $\frac{1}{2}$, $\frac{1}{2}$, 15 and 30in/s. The recorder is portable
and may be battery operated. It uses re-
usable cartridges containing standard
8-inch reels of 1-inch tape, and is also pro-
vided with a voice channel as standard. Pye TVT Ltd, Coldhams Lane, Cambridge.

Silicon junction varistors
A new series of silicon junction varistors, manufactured by General Semiconductor
Industries and marketed in the U.K. by
Auriema, consist of two matched silicon
junctions connected in parallel and oppos-
ite in polarity. The GSV series is rated for
125A peak-pulse current for lms. Breakdown voltage is typically 0.35V
to 0.50V at 10µA and 0.74V to 0.85V
at 100mA. Units are contained in an axial
lead package 0.200 × 0.370in. Auriema
Ltd, 442 Bath Road, Slough SL1 6BR.

Polyester capacitors
A range of capacitors with polyester film
dielectric and aluminium foil electrodes
is available from Jermy. The capacitors
have a low dissipation factor, high insu-
lation resistance, and low self inductance.
Capacitance and voltage ranges are:
0.01µF to 0.47µF 160V d.c.
1000pF to 0.47µF 400V d.c.
1000pF to 0.1µF 1000V d.c.
Standard capacitance tolerance is ±10%,
but ±5% or ±20% are available to order.
Jermy Industries, Vestry Estate, Seven-
oaks, Kent.

Phase-lock amplifier
Type A310 phase-lock amplifier from
Prosser Scientific Instruments is designed for
phase-locking signal generators and
function generators to other sources. It is
possible to lock the slave generator to the
signal when the slave frequency is har-
monically related. Both odd and even
harmonics may be locked in this mode.
The equipment is battery operated, but a
mains-powered version (type A311) is available. Prosser Scientific Instruments
Ltd, Lady Lane Industrial Estate, Hadleigh,
Ipswich IP7 6DQ.

Transistor/thyristor
insulating caps and sockets
The MS 501 and MS 502 series of insulating caps from Waycom have been
made of ceramic to insulate exposed surfaces of
TO-3 and TO-66 transistors or thyristors
when mounted on heatsinks. Manufactured
by Schaffner in black glass-filled nylon,
the caps have a maximum operating tem-
perature of 120°C. Also introduced is
the MS 510 socket which facilitates ease
of insertion and replacement of TO-3
transistors or thyristors. Waycom Semi-
conductors Ltd, Wokingham Road, Bracknell, Berks.

Precision voltage reference
diodes
Stable voltage reference diodes from Mul-
lard can provide a performance better than
that of a Weston standard cell. Types
BZX90 to BZX94, they have a nominal
voltage of 6.5V with temperature coeffi-
cients ranging from ±0.01 to ±0.005% per °C;
a typical value for the temperature co-
efficient of a Weston cell would be about
-0.004% per °C. The reference diode
is 4.25mm long and 1.85mm in diameter.
Mullard Ltd, Mullard House, Torrington
Place, London WC1E 7HD.

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Mullard Ltd, Mullard House, Torrington
Place, London WC1E 7HD.

Wireless World, February 1972
February Meetings

Tickets are required for some meetings: readers are advised, therefore, to communicate with the society concerned.

LONDON


1st. IEE — "Ceramic dielectrics for microwave devices" by R. C. Kelly at 17.30 at Savoy Pl., W.C.2.

2nd. IEE — Discussion on "Telecommunication stored programme control; large scale logic or microprocessors; which?" at 14.00 at Savoy Pl., W.C.2.


3rd. RTS — "Anchor - an electronic character generator" by R. J. Taylor at 19.00 at the I.T.A., 70 Brompton Rd, S.W.3.


8th. IEE — Discussion on "Are sandwich courses dying?" at 17.30 at Savoy Pl., W.C.2.

8th. AES — "Matrix systems for four-speaker stereo" by Michael A. Gerzon at 19.15 at the Mechanical Engineering Dept., Imperial College, Exhibition Rd, S.W.7.

9th. IEE — "The changing pattern in postgraduate education" by Prof. Hugh Ford at 17.30 at Savoy Pl., W.C.2.

10th. I.E.E. — "Conduction and magnetic signalling in the sea" colloquium at 18.00 at University College London, Torrington Pl., W.C.1.

11th. IEE — Discussion on "What does engineering management want from business schools?" at 17.30 at Savoy Pl., W.C.2.

14th. IEE — Discussion on "Automatic test equipment for r.f. communications" at 17.30 at Savoy Pl., W.C.2.


16th. IEE — Colloquium on "Aids to vehicle movement in low visibility conditions" at Savoy Pl., W.C.2.

16th. I.E.E. — "The control of innovation" by T. G. Clark at 18.00 at The Engineering Lecture Theatre, University College London, Torrington Pl., Gower St, W.C.1.

16th. SERT — "Digital processing of radio and television signals" by Dr. B. Moffatt at 19.00 at the I.T.A., 70 Brompton Rd, S.W.3.

17th. BCS — "Speech and synthesis" by A. J. Pourcin at 18.30 at the Engineering Lecture Theatre, University College London.

23rd. I.E.E. — Symposium on "The future of electronic displays" at 09.00 at the Royal Aeronautical Society, 4 Hamilton Pl., W.1.

23rd. I.Phys. — Applications of u.a. spectroscopy symposium at 11.00 at Imperial College.


ABERDEEN

23rd. I.E.E. — "Radio astronomy" by A. Flett at 19.30 at Robert Gordon's Institute of Technology, Physics Dept. Lecture Theatre, St. Andrew's St.

BIRMINGHAM

16th. RTS — "Special effects" by Mike Cox and Ray Knight at 19.00 at A.T.V. Centre, Broad St.

BRADFORD

10th. I.E.E./I.E.E — "Low-light television tubes" by P. A. Batey at 19.00 at The University, Dept. of Electrical and Electronic Engineering, Lecture Theatre C7.

BRIGHTON

24th. I.E.E.TE — "Principles and applications of lasers" by Dr. L. Allen at 19.30 at Royal Albion Hotel.


BRISTOL

15th. I.E.E. — "Digital techniques applied to the measurement of low-frequency noise" by Prof. H. R. Sutcliffe at 19.00 at the University, School of Chemistry, Lecture Theatre No.4.

CARDIFF

2nd. SERT — "Modern a.h.f. aerial design" by R. S. Roberts at 19.15 at Llandaff College of Technology, Western Avenue.

9th. I.E.E. — "The teaching of electronics" by Prof. A. Gambling at 18.30 at the University of Wales Institute of Science and Technology.

CHATHAM

24th. I.E.E. — "Budgetary control" by R. P. Reynolds at 19.00 at Medway College of Technology.

CHELMSFORD


CONNAUGHT QUAY, DARTFORD


EDINBURGH

8th. I.E.E./I.E.E. — "Electronics in crime detection" by A. T. Torlesse at 19.00 at Napier College of Science and Technology, Colinton Rd.

19th. I.E.E. — "Machine intelligence" by Prof. D. Mitchie at 18.00 at Mountbatten Buildings, Heriot-Watt University.

EVESHAM

14th. I.E.E. — "Domestic television recording techniques" at 19.00 at the B.B.C., Woodnorton.

EXETER

8th. I.E.E.TE — "Electricity and electronics in modern gas production" by W. Arkless at 19.30 at Imperial Hotel.

FARNBOROUGH, Hants

10th. I.E.E. — "A simplified atomic frequency source" by T. J. Bennett at 19.00 at the Technical College.

GLASGOW

17th. I.E.E./I.E.E. — "Electronics in crime detection" by A. T. Torlesse at 18.00 at The Institution of Engineers and Shipbuilders, Rankine House, 183 Bath St.

GUILDFORD


LIVERPOOL

9th. I.E.E. — "Instrumentation and the cardiac patient" by A. Lee at 19.00 at the University, Dept. of Electrical Engineering & Electronics.

LLANDAFF

2nd. RTS — "The technical future of television" by J. S. Sansom at 19.00 at the B.B.C., Broadcasting House.

LOUDBOROUGH


MANCHESTER

17th. I.E.E. — "Automatic landing — the instrument landing system" by J. W. B. McIvor at 18.15 at the U.M.I.S.T., Renold Bldg, Sackville St.

24th. SERT — "The Dolby B system" by J. N. Ayres at 19.30 at the U.M.I.S.T., Renold Building, Sackville St.

MIDDLESBROUGH

29th. SERT — "Electronic fuel injection" by J. T. Davies at 19.30 at Cleveland Scientific Institution.

NEWCASTLE-UPON-TYNE

2nd. SERT — "Dual in-line resistor network technique" by D. G. Fowler at 19.15 at the University of Newcastle Technical College, Mapple Terrace.

9th. I.E.E. — "Sensitivity capacitance micrometers and their applications" by Dr. C. S. Richards at 18.00 at Ellison Building, The Polytechnic.

NORWICH

9th. RTS — "Loudspeakers for the sound quality assessment of broadcast programmes" by R. E. Cooke at 19.30 at Anglia Television, Anglia House.

15th. SERT — "Decoders and C.D.A.s" by L. Briggs at 19.30 at Norwich City College, Ipswich Rd.


PLYMOUTH

3rd. I.E.E./I.E.E. — "Digital communication systems" by Prof. K. W. Cattermole at 19.00 at the Polytechnic.

READING


STAFFORD

28th. I.E.E./I.E.E./IPOEE — "The applications of electronics in large electrical machines" by Dr. R. D. M. Whitwell at 19.00 at North Staffordshire College of Technology, Beaconside.

SWINDON

8th. I.E.E./I.E.E. — "The computer in management: master, parasite or servant?" by D. Finberg at 18.15 at The College.

WOLVERHAMPTON

22nd. I.E.E. — "Teaching electronics at sixth form level" by Mr. Foxcroft at 19.15 at the Technical Teachers' College.
ACTIVE DEVICES
A data sheet from M.C.P. Electronics Ltd, Alperton, Wembley, Middx, HA0 4PE, gives information on the PT5741 f.f.c power transistor from TRW Semiconductors. The PT5741 is a silicon n-p-n interdigitated transistor for h.f./s.s.b. ultra linear application. WW401
We have received a leaflet from International Rectifier on their two new 40A medium power silicon transistors from IR-120S. International Rectifier, Hurst Green, Oxted, Surrey. WW402
A 23 x 21in wallchart detailing Motorola's plastic encapsulated small-signal transistors is available from GDS (Sales) Ltd, Michaelmas House, Salt Hill, Bath Rd, Stough, Bucks. WW403
Data sheets on five linear circuits are available from N.C. Semiconducotr GmbH, Lechaztra 255, 891 Landsberg, West Germany. LM112 (micropower operational amplifier) WW404
LM113 (temperature compensated low-voltage reference diode) WW405
LM114 and LM115 (transistor pairs with higspeed breakdown voltage and good d.c. characteristics) WW406
LM216 (high input impedance op.amp. for minimal input current errors) WW407
An electronic component stock catalogue from Celldis, 37/39 Lovelock Rd, Reading, Berks, RG3 1ED, lists all types of components (active and passive) available from Celldis. WW408
A comprehensive components catalogue for December 1971/March 1972 is available from R.S. Components Ltd, P.O. Box 427, 13/17 Epworth St, London, EC2P 2HA. WW409
Rastr Electronics Ltd, 275 King St, Hammersmith, London W6, have published price lists for integrated circuits and discrete components. WW410
A 1972 catalogue of electronic components has been received, from Sasco, P.O. Box 2000, Crawley, Sussex, RH10 2RU. WW411

PASSIVE COMPONENTS
We have received a sheet describing the series 899-3 resistor package from Beckman Instruments International, S.A., News bureau, 17 Rue des Pierres-du-Niton, 1207 Geneva, Switzerland. The d.d.l. package features seven thick film resistors of equal value available from the Hewlett Packard Division of Beckman Instruments. WW412
Model L-604-1 adjustable time delay equalizer (equalizes an L-band system time delay of 30ns variation to better than ±3ns over greater than 100MHz bandwidth) is described on a sheet from Wavecom Inc, 9036 Wilmot Avenue, Northridge, California 91324. WW413

APPLICATIONS
Data Recognition Ltd, Lovelock Rd, Battle Farm Estate, Reading, Berks RG3 1DX, have sent us a bulletin entitled 'The role of optical mark reading in accountancy'. WW414
A booklet entitled 'Simple stabilized power supplies' (R-271/1281) has been sent to us by Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD. WW415
TRW Semiconductors have a publishing description the design of a linear 100W solid-state amplifier intended for s.s.b. transmitter applications. MCP Electronics Ltd, Melbourne, Vic, Australia. HAO 4PE WW416

Two application notes are available from Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

- Numerical indicator tubes dynamic display systems operating from high-voltage supplies. WW417
- 'Drive circuits for Pandidian indicator tubes' WW418

A guide to automatic and semi-automatic op-amp testers has been sent to us by Teledyne Philbrick Nexus, St. Peter's House, Chichester, Sussex. It provides circuit descriptions and diagrams for testing calibration components which are discussed in terms of two Philbrick Tasters (Models 5104 and 5107). WW419

EQUIPMENT
New data sheets from Ferranti, Electronic and Display Equipment Division include data on the OLA/1049 c.r.t., a head-up display tube designed to meet adverse environmental conditions. Also included are details of the latest measurement techniques, and an information sheet 'Photographic data recording'. Ferranti Ltd, Gem Mill, Chaderton Oldham, Lancs. WW420

A leaflet is available on model AN2532 digital panel meter, featuring a 3½ digit display, 0.1% accuracy and a floating differential input. Available from Analogic, Audubon Rd, Wakefield, Massachusetts 01880, U.S.A. WW421

Information on the 'M' range fractional horsepower induction motor is contained in a leaflet from Comex. The 'M' range may be larger than the standard range but will develop 50% more power. Comex Ltd, Senate House, Tysen St, London EC 2ND. WW422

'Hospital Headsets for Patients' is the title of a leaflet describing and illustrating the headsets and also the audio faceplate units and also a 'pillowphone' produced by Danavox (Gt. Britain) Ltd, 'Broadlands', Bagshot Rd, Sunninghill, Ascot, Berks, SL5 9W. WW423

We have received an information sheet on an infra-red viewer type IRV-M for use in the medical field. The IRV-M was designed primarily for investigating by direct viewing of the effect upon spontaneous and induced nystagmus when fixation is removed by total darkness. Level Developments, 7-10, Hildreth Mews, London SW1 1RZ. WW424

Zam and Kirshner, Inc, 570 Seventh Ave, New York, N.Y.10018, have produced a brochure describing specifications and applications of three miniature, portable, real-time spectrum analyzers produced by Federal Scientific. WW425

A mixer intended for use in radio stations and the smaller television and recording studios is described in a leaflet from Audix. Designated MXT-800, the mixer has up to 16 input channels with two or four output groups and a wide range of ancillary modules. Audix Ltd, Stansted, Essex. WW426

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorks, LS22 5TR, have sent a leaflet describing several of their products including signal generators, power supplies, printer, digital controller and a 1.5GHz millimivoltmeter. WW427

Venner, a division of AMF International Ltd, have issued a brochure on their Time Control equipment. A brief summary is given of their more popular range of timers for heating, lighting and process control for industrial and domestic applications. Venner, Kingston By-Pass, New Malden, Surrey. WW428

An indicator with 4½ inch high characters called the 'Maxitron' is described in a leaflet from KG 51 Electronics. Specifications and description of the M107B and M107D Maxitrons are given. KG5 Electronics Ltd, Clock Tower Rd, Isleworth, Middlesex. WW429

Advance Electronics Ltd, Raynham Rd, Bishop's Stortford, Herts, have sent in their information on the OS2100 and OS2200 series oscilloscopes, the DPM 300 digital panel meter range and the PG 58 wide-range pulse generator. WW432

A brochure on the 'Slimpak' 19in racking system has been published by Contaut Electronics Ltd, 3 Trafford Rd, Radford, Coventry. WW433

We have received a brochure describing the 58 different Kepco type 'PRM' voltage stabilizers covering 4.5 to 2400V d.c., with ratings from 60 to 300W. Also included is a guide to the ferroresonant technique of voltage stabilization. Kepco Inc, Mamaroneck, N.Y. (UK representatives), 442 Bath Rd, Slough, Bucks. SL1 6EB WW436

Available equipment and services provided by Astley Electric Ltd, Ardriale Ave, South Ockendon, Essex, RM15 5SR, are given in a publicaion WW438

We have received leaflets on variable speech control which enables speech to be adjusted up or down, or down-converted speech with any gramophone or tape recording machine without any change in pitch. Cambridge Research and Development Group, 2 Newbridge Square, Westport, Con 706880, U.K. WW439

Descriptive literature of a Konkati product 'Positiv 20' — used for the production of individual or small series printed circuits — and other Konkati products are available from Specialist Products Distributors Ltd, Piccadilly, London W1Y 0HL. WW441

GENERAL INFORMATION
A bulletin on short and long courses in technology, management studies and commerce in the London and home counties region gives a short description of the content of each available course. London and Home Counties Regional Advisory Council for Technological Education, Tavistock House South, Tavistock Square, London WC1H 9LR Price 60p.

A list of short courses in electrical engineering and physics at Hatfield Polytechnic ranges from control and environment engineering to electronics for executives. An application form for enrolment is included with the leaflet. Hatfield Polytechnic, Hatfield, Herts. WW443

Details of services and background information on Rediffusion Radio Holdings is given in a booklet from Rediffusion Radio Ltd, 11th Floor, Baring Division, P.O. Box 451, Carlton House, Lower Regent St, London SW1 4LS. WW444


A leaflet from Firth Cleveland Fastenings Ltd, Treforest, Glamorgan, provides details of a range of self-locking nuts based on the 'Clevloc' design. WW445

A list of new films from the Central Film Library, (Industrial section) can be obtained from Government Building, Broomy Hill, Axton, London W3 7JB. Also available is the current 'Films for industry' catalogue. WW448

We have been sent a folder containing data sheets of Goodmans loudspeaker units and associated accessories, together with recommended cabinet designs. Goodmans Loudspeakers Ltd, Downley Rd, Havant, Hants. PO9 2NL. WW450

A leaflet on a range of glass-to-metal seals has been issued by Catheodon Ltd, Nuffield Rd, Cambridge CB4 4TF. These seals can be used for taking electrical and electronic equipment from continuously evacuated enclosures. WW450

Two publications are available from British Standards Institution, 2 Park St, London W1A 2BS:

- 'Rules for the preparation of detail specifications for semiconductor devices of assessed quality' microwave mixer diodes (c.w. operation) Price 50p.

For further information on any item include the WW number on the reader reply card.
This and that: but mostly Raudive

By the time my regular reader gets around to this page, it will almost be February and we shall probably be sliding through snow and ice and piling the poor aerial riggers on a day like this. At home, the post-Christmas bills will have fluttered down like leaves in a winter's gale, while at work the axemen will be honing a fine edge on their economy choppers. In short, a dismal outlook. A time of year when not a single prospect pleases and man, having got the spirit of Christmas out of his system for another year, is being more than usually vile. But enough of gloom, for as I write its Christmas Eve; the portcullis is lowered, the curtains are drawn, the 2kW yule log is blazing merrily and the horrors of January are light-years away. So don't expect any mordant comments on the electronics scene in this issue; just a few random reflections on what has passed.

But first, let me remind you that this is the age of the instant carol, a tough competitive business indeed. A few nights ago a thunderous pounding on the door brought us out post-haste. One beetle-browed small boy stood on the threshold holding out a paw. 'Carols?' he demanded brusquely. We said rather apologetically that we would prefer a sample before payment, whereupon, without another word, he disappeared into the darkness.

Last night we were brought to the door again. This time, two small boys were there, one with a tray suspended in front of him by means of a string around his neck. As the door opened we got our carols, a dreadful travesty of the choir of King's College, Cambridge, no less—a battery-driven tape recorder perched on the tray. We paid up to have it switched off.

What better time than Christmas Eve to talk of ghosts, spectres and things that go bump in the night? You may remember that some months ago I wrote about the so-called Raudive voices—you know, voices that are supposed to emanate from an ordinary tape through non-human agency. I recall that at the time the Editor, in his ineffable wisdom, predicted that the subject would produce a large post-bag. And, so help me, he was right; more letters resulted from that topic than from the rest of the year's topics put together.

At the time I mentally congratulated myself on having sat rather neatly on the fence without letting personal beliefs intrude. I put forward a suggestion that possibly the phenomenon might be accounted for within the framework of known natural laws. In sum, all I said in effect was that the matter ought to be thrashed out one way or another. Truly, pride goeth before a fall! Wasn't it Buster Keaton who, in an American Civil War film, walked unscathed between the two opposing armies because he bore a flag with the Union insignia on one side and the Confederate colours on the other? I tried the same trick and all I got for my trouble was to be caught in withering fire from both sides. For, understandably perhaps, all the letters derived from one of two camps. There was the 'spiritualism is bunk' school which berated me for my gullibility in falling for such palpable rubbish, and there was the 'confirmed believer' school which reproached me for lack of faith in what (they said) had been thoroughly proven times without number. If there was a lesson to be assimilated from the exercise it was, surely, that you can't win.

What I had hoped for at the outset was that there would be some uncommitted readers who would try the matter out on an impartial basis. But although a number expressed intention to experiment, these, without exception, held spiritualist beliefs. In the event, nobody subsequently wrote to report either success or failure.

As for me, I tried a limited number of times (about ten) using the simplest possible approach, namely, just an ordinary tape recorder left switched on in an empty room, with no attempt to create any form of seance atmosphere. Absolutely nothing came through on any occasion. Oh yes, there were voices and noises on play-back, but all of them traceable to known natural sources. The weather was hot and the windows were open and with an omnidirectional microphone its amazing how voices carry from a distance on a still summer evening. One notable feature was a barking dog and much as I'd like to attribute the noise to a spectral hound, honesty compels me to admit that it was readily traceable to a Great Dane who lives a quarter of a mile away.

I wouldn't have brought the subject up again except that I have received a copy of an article on the 'Raudive voices' from an Australian journal, *Electronics Today*, (with a reproduction of part of the relevant *W.W.* page). This article gives an account of two experiments which were performed at unspecified dates, possibly after the *Wireless World* article: at any rate I didn't know about them at the time it was written.

One of these is reported as taking place under carefully controlled conditions under the supervision of Mr Kenneth Attwood, chief engineer of Pye Recordings Ltd. The other (says the journal) was undertaken at the laboratories of Belling and Lee and was under similarly rigid conditions. Now, whatever your attitude toward the possibility of voices from the dead, none will question the integrity of the two organizations mentioned; any pronouncements therefrom demand at least a respectful hearing.

The experiment carried out by Mr Attwood was (says the Australian journal) performed over a period of nine hours, from 5.30 p.m. to 2.30 a.m. and was done in the presence of 25 witnesses. Extreme measures were taken to ensure that there was no possibility of electromagnetic pick-up from any external source; in other words, every effort was made to prevent extraneous voices or sounds appearing on the tapes. Yet, it seems, at playback upon conclusion of the session approximately, two hundred voices had manifested themselves. Of these, 27 were so clearly audible that they could be directly played back to the audience.

No detail is given of the Belling and Lee experiment except that one of those taking part was Mr Peter Hale, 'scientific advisor to the British Government'. Mr Hale is quoted as saying:—'From the result we obtained... I must state that something is happening which I can no longer explain in normal physical terms'.

Well, there you are. Where do we go from here? Clearly, if there is anything in this at all, the microphone doesn't play its normal role. If sound waves were impinging on it to produce the voices these would be heard at first hand by those present. So presumably the microphone could be replaced by a dummy load, or could it? Is the 'interference' occurring in the electrical part of the microphone or at the recording head? Because of the feeble nature of the alleged signals, the latter would seem the more likely. Has any experiment been performed to isolate the place of incidence?

If this should happen to catch the eye of Mr Attwood, Mr Hale or Mr D. Ellis of Trinity College, Cambridge, who has been given a postgraduate scholarship to investigate the phenomenon, perhaps one of them at least could be prevailed upon to let us know the present state of the art.

**Correction.** Regrettably, my indecipherable writing led to an error in the August 1971 piece 'On Stopping the Home Fires Burning', which was reproduced in the August 1972 issue of *Wireless World*. The approximate title of the report referred to in the centre column of p.410 should have been 'Jotro (Joint Fire Research Organization of the Dept. of the Environment and the Fire Offices Committee) and not Jitco.'