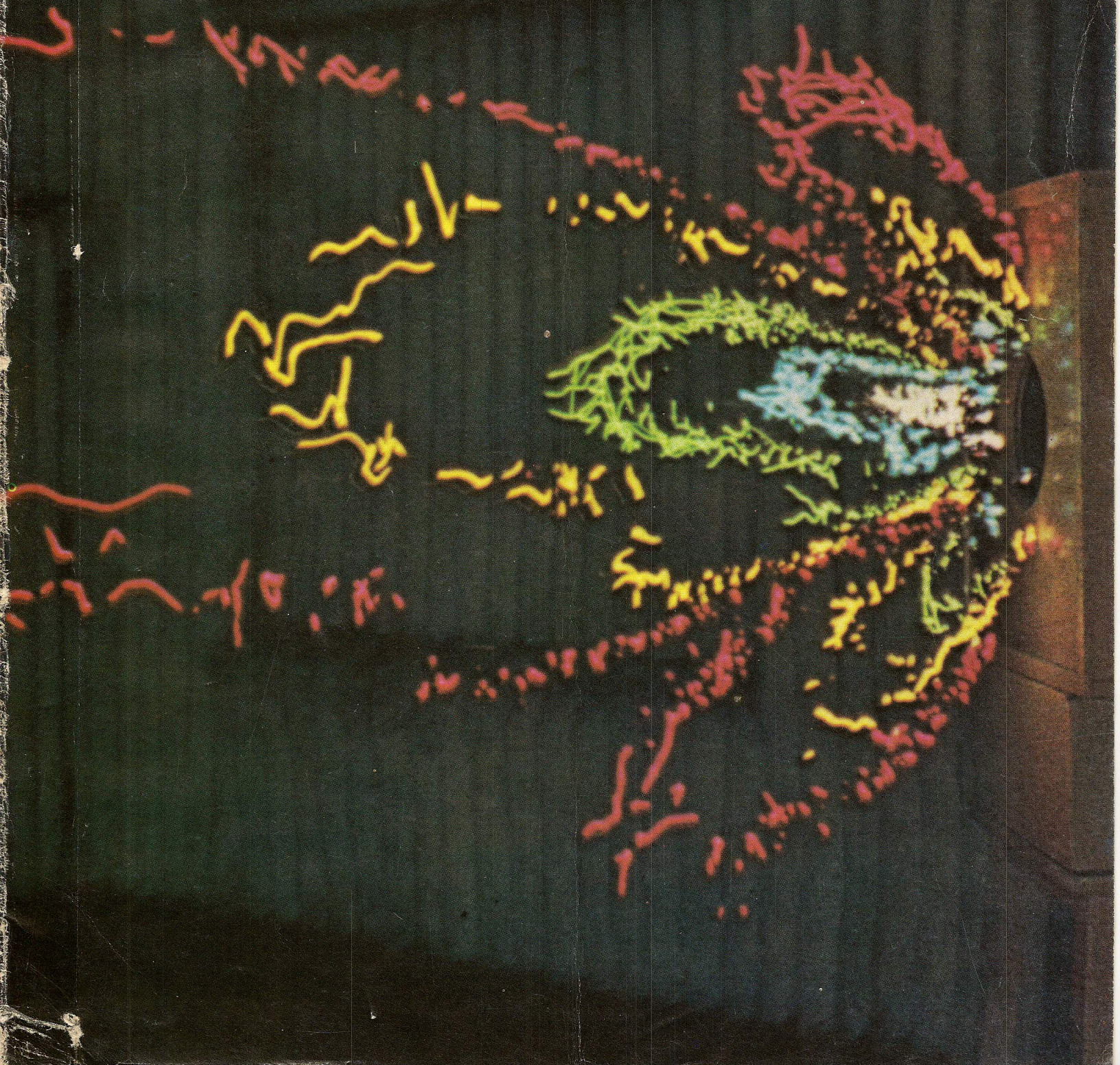


Wireless World

February 1972 20p

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Wireless World

Walter Cocking retires

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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 wobblator 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. Tusting. 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 quadrasonic 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 number¹. 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 article², 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, quadratizing 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 difectionality 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 Tappan³, 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 quadrasony 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

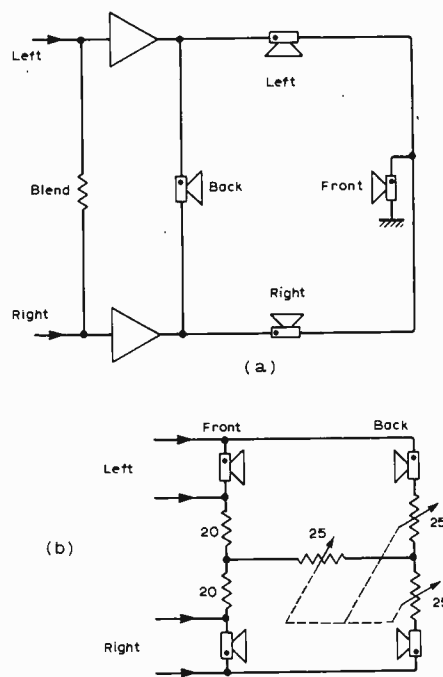


Fig. 1 Speaker matrix used with Dynaco 'diamond' array (a) and Dynaco 'square' array (b). Circuit (b) is that of the 'Quadtator'.

*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 6dB. (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

$$\begin{bmatrix} L'_F \\ R'_F \\ L'_B \\ R'_B \end{bmatrix} = \begin{bmatrix} 1 & 0.71 & 0.71 & 0 \\ 0.71 & 1 & 0 & 0.71 \\ 0.71 & 0 & 1 & -0.71 \\ 0 & 0.71 & -0.71 & 1 \end{bmatrix} \begin{bmatrix} L_F \\ R_F \\ L_B \\ R_B \end{bmatrix}$$

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 7.7dB, and in the mono mode the back signals are 7.7dB down. With the E-V matrix, left and right front crosstalk is 10dB in the two-channel mode, and in mono back signals are 8.3dB 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 — 10dB as opposed to 6.5dB.

A matrix proposed by Tappan³, 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,

TABLE 1

code	decode	outputs with coded material		
		four-channel	two-channel*	mono
Dynaco — 'diamond' array				
$L_T = L + F + B$	$L' = L_T$	$L + F + B$	left-right	$L + R + 2F$ front + 6dB back zero
$R_T = R + F - B$	$R' = R_T$	$R + F - B$	crosstalk null	
	$F = L_T + R_T$	$L + R + 2F$		
	$B' = L_T - R_T$	$L - R + 2B$		
Tappan modification				
$L_T = L + 0.7F + 0.7B$	$L' = L_T$	$L + 0.7F + 0.7B$	left-right	$L + R + 1.4F$ front + 3dB back zero
$R_T = R + 0.7F - 0.7B$	$R' = R_T$	$R + 0.7F - 0.7B$	crosstalk null	
	$F = 0.7(L_T + R_T)$	$F + 0.7L + 0.7R$		
	$B' = 0.7(L_T - R_T)$	$B + 0.7L - 0.7R$		

numbers recorded off to one decimal place.
*equations as under 'code'

but with back signals emphasized by 3dB. and in mono back signals are 7.8dB down.

One way of overcoming the severe mono incompatibility, proposed by Scheiber⁴, is to introduce a differential phase shift between the two coded signals of 90° (say $\phi + 45^\circ$ on one channel and $\phi - 45^\circ$ 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 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 advocating a new system — as from October last — probably the later matrix developed by Scheiber and announced at about that time. We hope to give more details later. (It is for this latest system that Electro-Voice are offering a new i.c. 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 system⁵. 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

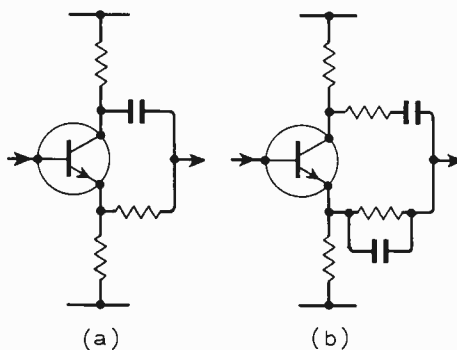


Fig. 2 All-pass filters provide a frequency-dependent 90° phase shift. Circuit at (a) used in Sansui decoders and at (b) in CBS decoders.

(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 $+90^\circ$ 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 $+90^\circ$ 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 is zero, and centre front-to-back side-to-side crosstalk is 7.7dB as for the basic matrix without phase shifters. Also as with the basic matrix, 'front' left-right crosstalk in the two-channel mode is 7.7dB (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 two back signals will emerge shifted by $+90^\circ$ 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 ϕ_1 , at a certain frequency, and the back channels a phase shift of $\phi_1 - 90^\circ$ and $\phi_1 + 90^\circ$ at that frequency. At another frequency the front phase might be ϕ_2 but so long as the rear phase shifts are $\phi_2 - 90^\circ$ and $\phi_2 + 90^\circ$, 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

		Outputs with coded material					
Code	Decode	Four-channel	Crosstalk (dB)	Two-channel*	Mono		
DYNACO 'square' array							
$\begin{bmatrix} 1.0 & 0.25 & 1.0 & -0.5 \\ 0.25 & 1.0 & -0.5 & 1.0 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0 \\ 0 & 1.0 \\ 0.64 & -0.36 \\ -0.36 & 0.64 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0.25 & 1.0 & -0.5 \\ 0.25 & 1.0 & -0.5 & 1.0 \\ 0.64 & -0.36 & 0.55 & -0.2 \\ -0.2 & 0.55 & -0.68 & 0.82 \end{bmatrix}$		Front crosstalk 12dB	$\begin{bmatrix} 1.25L_F + 1.25R_F \\ +0.5L_B + 0.5R_B \end{bmatrix}$	Back -8dB	
ELECTRO-VOICE original system							
$\begin{bmatrix} 1.0 & 0.3 & 1.0 & -0.5 \\ 0.3 & 1.0 & -0.5 & 1.0 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0.2 \\ 0.2 & 1.0 \\ 0.76 & -0.61 \\ -0.61 & 0.76 \end{bmatrix}$	$\begin{bmatrix} 1.06 & 0.5 & 0.9 & -0.3 \\ 0.5 & 1.06 & -0.3 & 0.9 \\ 0.59 & -0.38 & 1.06 & -0.99 \\ -0.38 & 0.59 & -0.99 & 1.06 \end{bmatrix}$		Front crosstalk 10.4dB	$\begin{bmatrix} 1.3L_F + 1.3R_F \\ 0.5L_B + 0.5R_B \end{bmatrix}$	Back -8.3dB	
ZENITH decode							
	$\begin{bmatrix} 1.0 & 0 \\ 0 & 1.0 \\ 0.68 & -0.53 \\ -0.53 & 0.68 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0.3 & 1.0 & -0.5 \\ 0.3 & 1.0 & -0.5 & 1.0 \\ 0.52 & -0.33 & 0.95 & -0.81 \\ -0.33 & 0.52 & -0.87 & 0.95 \end{bmatrix}$		As above	As above	As above	
SCHEIBER original system							
$\begin{bmatrix} 0.92 & 0.38 & 0.92 & -0.38 \\ 0.38 & 0.92 & -0.38 & 0.92 \end{bmatrix}$	$\begin{bmatrix} 0.92 & 0.38 \\ 0.38 & 0.92 \\ 0.92 & -0.38 \\ -0.38 & 0.92 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0.71 & 0.71 & 0 \\ 0.71 & 1.0 & 0 & 0.71 \\ 0.71 & 0 & 1.0 & -0.71 \\ 0 & 0.71 & -0.71 & 1.0 \end{bmatrix}$		Front crosstalk 7.7dB	$\begin{bmatrix} 1.31L_F + 1.31R_F \\ +0.54L_B + 0.54R_B \end{bmatrix}$	Back -7.7dB	0dB with phase shift mod.
TAPPAN							
$\begin{bmatrix} 0.71 & 0 & 1.0 & -0.71 \\ 0 & 0.71 & -0.71 & 1.0 \end{bmatrix}$	$\begin{bmatrix} 1.41 & 1.0 \\ 1.0 & 1.41 \\ 1.0 & 0 \\ 0 & 1.0 \end{bmatrix}$	As above	As above	Front crosstalk none	$\begin{bmatrix} 0.71L_F + 0.71R_F \\ +0.29L_B + 0.29R_B \end{bmatrix}$	Back -7.8dB	
SANSUI							
$\begin{bmatrix} 0.92 & 0.38 & j0.92 & j0.38 \\ 0.38 & 0.92 & -j0.38 & -j0.92 \end{bmatrix}$	$\begin{bmatrix} 0.92 & 0.38 \\ 0.38 & 0.92 \\ -j0.92 & j0.38 \\ -j0.38 & j0.92 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0.71 & j0.71 & 0 \\ 0.71 & 1.0 & 0 & -j0.71 \\ -j0.71 & 0 & 1.0 & 0.71 \\ 0 & j0.71 & 0.71 & 1.0 \end{bmatrix}$		Front crosstalk 7.7dB	$\begin{bmatrix} 1.31L_F + 1.31R_F \\ +j0.54L_B - j0.54R_B \end{bmatrix}$		
CBS							
$\begin{bmatrix} 1.0 & 0 & -j0.71 & 0.71 \\ 0 & 1.0 & -0.71 & j0.71 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0 \\ 0 & 1.0 \\ j0.71 & -0.71 \\ 0.71 & -j0.71 \end{bmatrix}$	$\begin{bmatrix} 1.0 & 0 & -j0.71 & 0.71 \\ 0 & 1.0 & -0.71 & j0.71 \\ j0.71 & -0.71 & 1.0 & 0 \\ 0.71 & -j0.71 & 0 & 1.0 \end{bmatrix}$		Front crosstalk none	$\begin{bmatrix} L_F + R_F \\ +L_B(-135^\circ) + R_B(+45^\circ) \end{bmatrix}$	Centre back signal suppressed	
Alternative forms of decode		$\begin{bmatrix} 1.0 & 0.25 \\ 0.25 & 1.0 \\ 0.71(0.25+j) & -0.71(1+0.25j) \\ 0.71(1+0.25j) & -0.71(0.25+j) \end{bmatrix}$	Improved centre front to back crosstalk		As above	As above	
Notes		$\begin{bmatrix} 1.0 & 0.1 \\ 0.1 & 1.0 \\ 0.71(0.4+j) & -0.71(1+0.4j) \\ 0.71(1+0.4j) & -0.71(0.4+j) \end{bmatrix}$	Centre front to back crosstalk 7dB		As above	As above	

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 reverberant 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 system⁶ 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 20-dB 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 8-dB rear crosstalk and a centre front-to-back differential of 7dB. 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

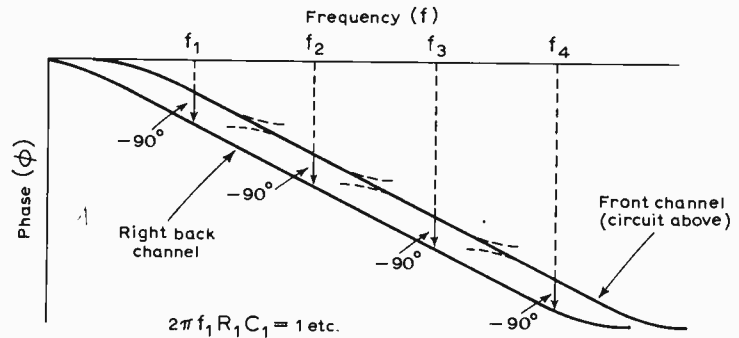
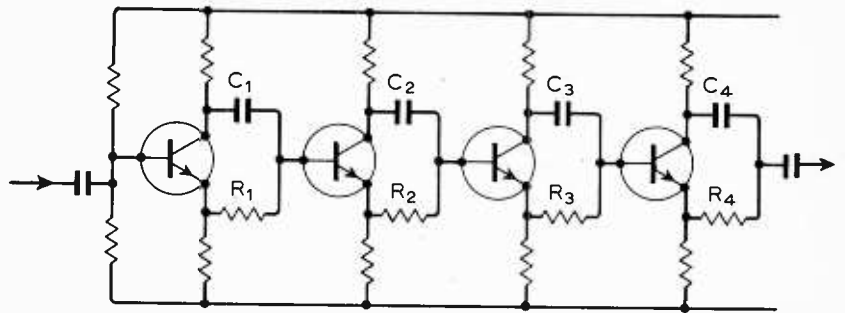


Fig. 3 Phase modulator circuit used in Sansui decoders to rapidly vary phase shift either side of the 90° norm.

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.

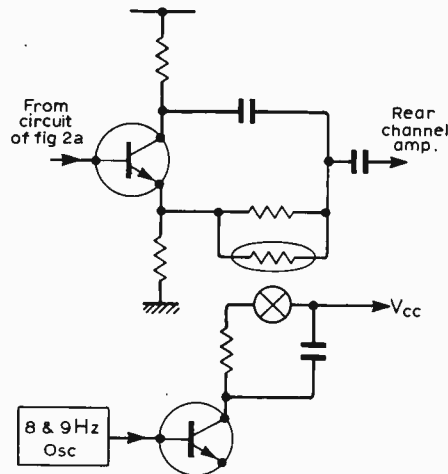


Fig. 4 Two cascaded all-pass filters one in each of two channels can provide a wideband 90° phase difference between the two channels.

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.
2. G. Shorter, "Four-channel stereo" *Wireless World* vol. 78 1972 pp.2-5.
3. P. Tappan "One plus one equals four". 39th A.E.S. convention (1970).
4. P. Scheiber "Four channels and compatibility" *J. Audio Eng. Soc.* vol. 19 1971 pp.267-79.
5. R. Itoh "Proposed universal encoding standard for compatible four-channel matrixing" 41st A.E.S. convention (1971).
6. "Quadraphony & home video steal the Berlin show" in *Wireless World* vol. 77 1971 pp. 486-8 includes a description of the CBS SQ system, described in full by B. B. Bauer, D. W. Gravereaux and A. J. Gust in *J. Audio Eng. Soc.* vol. 19 1971 pp.638-46.

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. B6 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 it 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'.

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.

*At mid 1971 prices

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



Twenty million tests per second on logic integrated circuits can be made by this computer controlled logic test system type S-3160 introduced by Tektronix. Up to four test bays can be controlled by the one computer.

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 Ettringen in the Wertach valley (31 miles south west of Ausberg). 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 from 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-

ers. Strength variations ranged from 3.2 to $6.6 \times 10^9 \text{ N/m}^2$. Each laboratory was supplied with a random selection of specimens together with a random order of testing for 10 specimens of each substrate. Samples were cleaned and dried; the relative humidity of the air was recorded; and the samples tested. Each 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 were fed 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. Suffice 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.

Push-button telephones using m.o.s.ts

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.

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. Viczian 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.

*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.ms 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 no real 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 $10k\Omega$ resistor to produce the required damping, of $50\mu A$ and a voltage sensitivity of $125mV$.

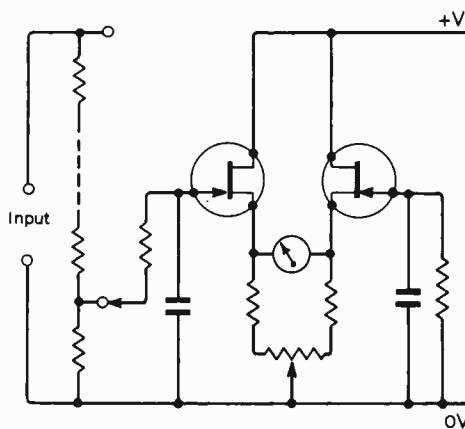


Fig. 1. Type of amplifier, for d.c. only, suitable for increasing input impedance.

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 $2M\Omega/volt$ and $200M\Omega/volt$. 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.

Circuit design

Fig. 2 shows the basic circuit from which the final designs evolved. Use of the RCA integrated circuit, CA3018, determined the transistor configuration used. It contains a pair of isolated matched transistors and a Darlington-connected pair. The input stage of Fig. 2 is a differential amplifier of the long-tailed pair variety. The output from the differential amplifier is fed to the Darlington-connected pair of transistors via a potential divider network in order to match the differing d.c. potentials at the collector of Tr_1 and the base of Tr_3 . The emitter of Tr_4 is held at a negative voltage by the zener diode, enabling its collector potential to vary above and below the zero

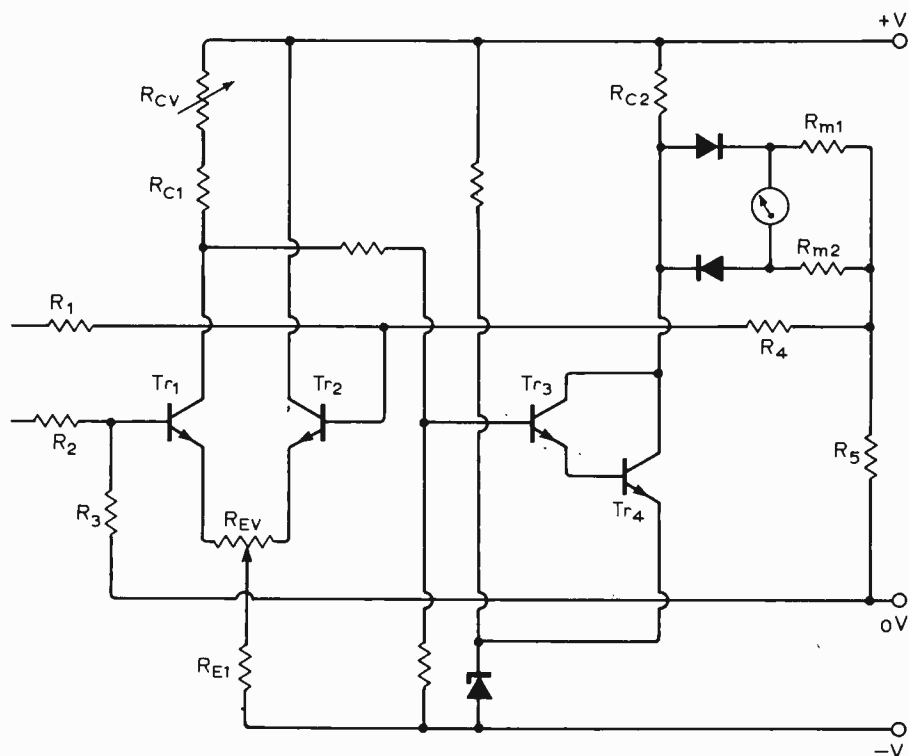


Fig. 2. Basic circuit from which amplifier designs were evolved. The two pairs of transistors are provided by the CA3018 integrated circuit.

volt line. The meter circuit is connected between Tr_4 collector and the zero volt line. Negative feedback is achieved by connecting the base of Tr_2 , via R_4 , to the junction of R_5 , R_{m1} and R_{m2} .

With no input signal to the amplifier the base of Tr_1 will be at some voltage slightly negative of the zero volt line due to the voltage dropped across R_3 by Tr_1 's input bias current. Provided that the differential amplifier is balanced, under open-circuit input (by R_{EV}) and short-circuit input (by R_{CV}) conditions, for minimum deflection of the meter, the base of Tr_2 must be at the same potential as the base of Tr_1 and the voltage dropped across R_4 must be the same as the voltage dropped across R_3 . Thus the potential at the junction of R_5 , R_4 , R_{m1} and R_{m2} must be equal to zero volts. A portion of the input bias current to Tr_2 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 Tr_1 and Tr_2 are initially upset, causing the voltage at the collector of Tr_1 to shift according to the polarity of the input signal. The voltage shift at the collector of Tr_1 is amplified by the Darlington pair, producing an even greater and opposite shift in the collector voltage of Tr_4 . A voltage is thus de-

veloped across R_5 , 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_5 , 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_5 (equal to $R_5 I_L$, where I_L is the current flowing through the meter circuit) will be equal to:

$$V_{in} (R_3 + R_4) / (R_1 + R_2)$$

Thus, $R_5 I_L = V_{in} (R_3 + R_4) / (R_1 + R_2)$ and the input voltage sensitivity for f.s.d. equals $R_5 I_L (R_1 + R_2) / (R_3 + R_4)$ when I_L equals the current flowing through the meter circuit 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 a 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 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_{C1} , R_{C2} and R_{E1} have all been replaced by constant current sources. By replacing R_{C2} 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 design¹. By replacing R_{E1} 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 transistors' bases. It also has the advantage of increasing the common-mode rejection ratio of the differential pair. The replacement of R_{C1} 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_{E1} and R_{C2} 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 Tr_1 to Tr_4 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 $\pm 10\%$ and V_{BE} to $\pm 2mV$ over a range of collector currents from $10\mu A$ 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 $10\mu A$. The current gain of the Darlington pair is a minimum of 1000 at a collector current of $100\mu A$ and 2000 at $1mA$. The collector currents of the differential pair are designed to be $10\mu A$. For a minimum gain of 30, the input bias currents are thus a maximum of $0.33\mu A$. By virtue of the fact that the basic sensitivity of the Avo Model 9 meter movement, when shunted with a $10k\Omega$ resistor, is $50\mu A$, and that two $5.1k\Omega$ 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. $0.17\mu A$, which is only 0.3% of the $50\mu A$ 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 Tr_3 is designed to be $10\mu A$. The constant current supplied by Tr_5 thus needs to be $20\mu A$. The collector current of the Darling-

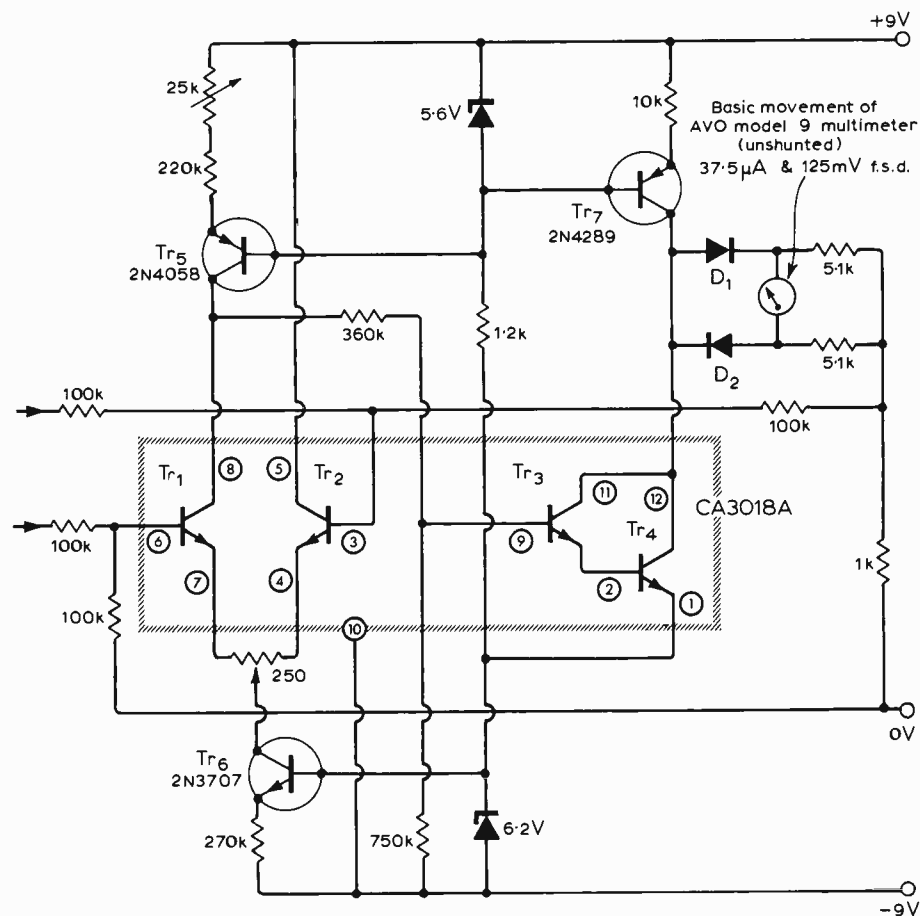


Fig. 3. Practical circuit of meter amplifier, with higher open-loop gain and higher output impedance than those of the basic circuit (Fig. 2). In the i.c. the circled numbers indicate the conductors which are taken out to leads (with corresponding numbers) on the TO-5 12-lead package. No. 10 is the chip substrate. (The intersections of conductors with the shaded rectangle have no significance as connections.) D_1 and D_2 are type OA202.

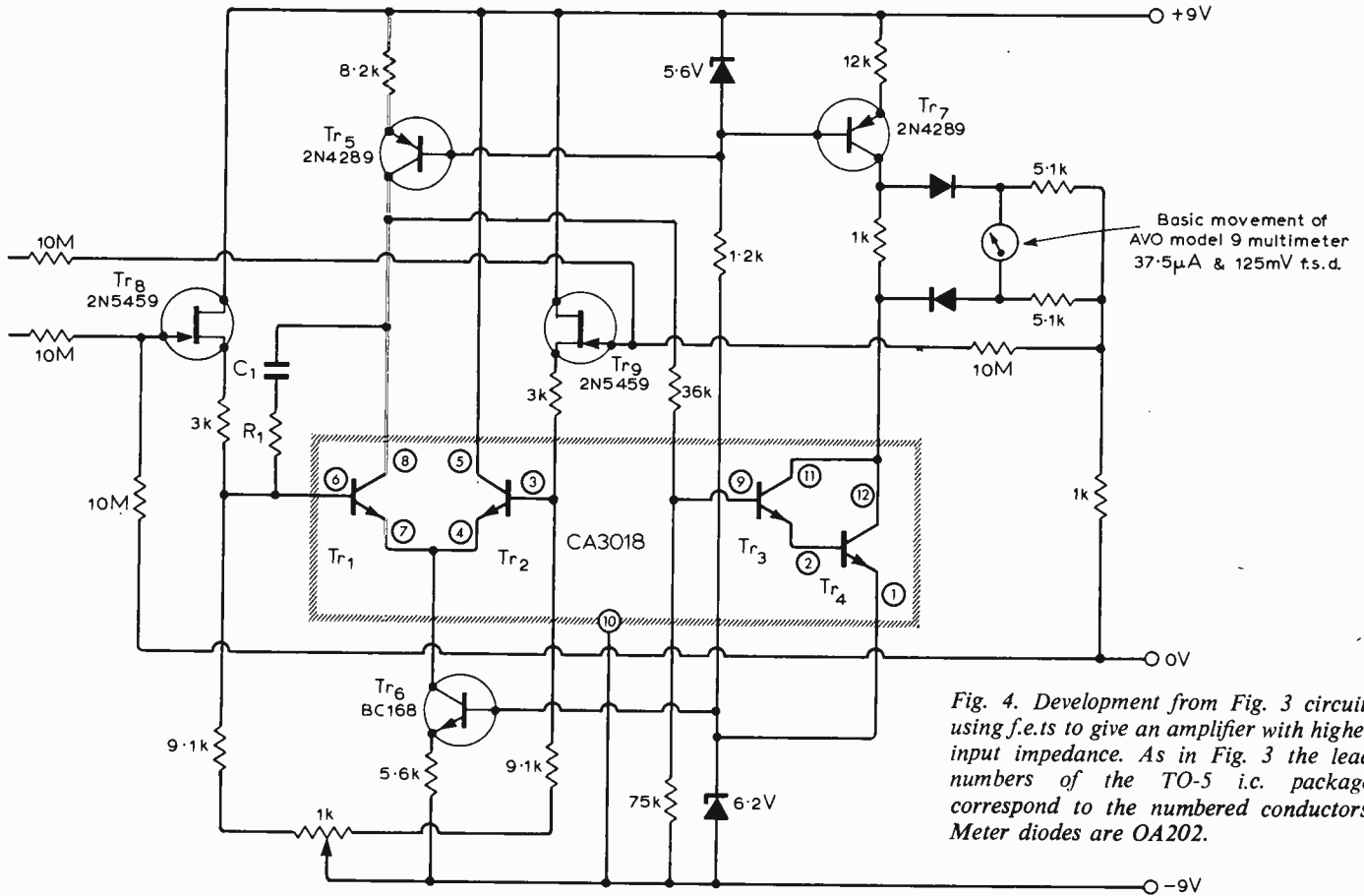


Fig. 4. Development from Fig. 3 circuit, using f.e.t.s to give an amplifier with higher input impedance. As in Fig. 3 the lead numbers of the TO-5 i.c. package correspond to the numbered conductors. Meter diodes are OA202.

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 Tr₃ 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 Tr₁ 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 Tr₁ 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 Tr₄ 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 R₁ and R₂ (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 not more than 0.33µA and the values of R₃ and R₄ are 100kΩ, the short circuit input conditions will not be out of balance by more than ±10% of 100kΩ × 0.33µA = ±3.3mV.

With 10µA flowing through the collectors of Tr₁ and Tr₂, 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.

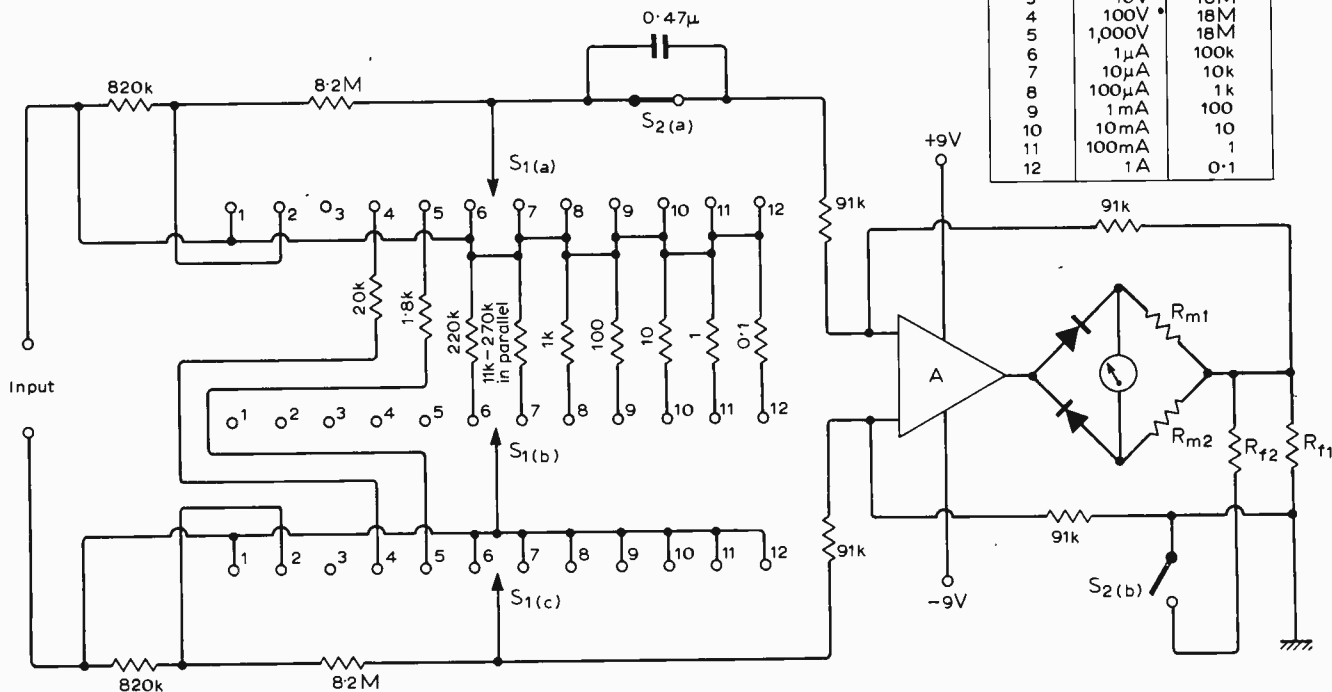
In order to correct the short circuit input balance, it is necessary to ensure that the voltages dropped across the resistors R₃ and R₄ (of Fig. 2) are identical. This can, of course, be done by altering the value of either R₃ or R₄, but as this would upset the gain of the amplifier an alternative method 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_e ≈ I_{c1} + I_{c2}. I_{c1} is a constant current and therefore so also is I_{c2}, equal approximately to I_e - I_{c1}. Thus, if I_{c1} 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 Tr₅ are designed to vary the collector current of Tr₅ 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 750k resistors, this means that the collector currents of Tr₁ and Tr₂ can be varied over a range of 9µA to 11µA. This gives an overall ±20% control over the base currents of Tr₁ and Tr₂. 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, R₃ and R₄. 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

Fig. 5. Multimeter constructed using amplifier circuit of Fig. 3, shown as amplifier 'A', with an attenuator. Note: $R_{f2} = 9 \times R_{f1}$. Meter diodes are OA202.



Position S ₁	Range (f.s.d.)	Input impedance
1	100mV	180k
2	1V	1.8M
3	10V	18M
4	100V	18M
5	1,000V	18M
6	1μA	100k
7	10μA	10k
8	100μA	1k
9	1mA	100
10	10mA	10
11	100mA	1
12	1A	0.1

V_{BE2} of the differential pair with temperature is typically $10\mu\text{V}/^\circ\text{C}$ for the CA3018A. However, the stability of the circuit in general will depend upon the variation of the base-emitter voltages, with temperature, of the transistors Tr_3 to Tr_7 and the temperature coefficients of the zener diodes. A typical variation of V_{BE} with temperature, for a silicon transistor, is $-2\text{mV}/^\circ\text{C}$. A variation of this order to V_{BE} of transistor Tr_6 would result in a change in the constant current to the emitters of the differential pair of approximately $7\text{nA}/^\circ\text{C}$ (i.e. $2\text{mV}/270\text{k}\Omega$). If it can be assumed that the collector current of Tr_1 remains constant, then this will result in a change to the collector current of Tr_2 of approximately 7nA and hence a change to its input base bias current of $7\text{nA}/30 = 200\text{pA}$. This would cause a change in the short-circuit balance conditions of approx. $200\text{pA} \times 100\text{k}$, i.e. approx. 20V and, if all of the 250ohms of the potentiometer is in the emitter circuit of Tr_2 , a change to V_{BE2} of $7\text{nA} \times 250$, i.e. $2\mu\text{V}$. Changes in the values of the base-emitter voltages of the other transistors, Tr_3 and Tr_7 , with temperature all have similar minimal effects and, because they are unlikely to occur in isolation from one another, will tend to have a cancelling out effect rather than an additive one. The temperature coefficient of the two base-emitter voltages of the Darlington pair is typically $4.4\text{mV}/^\circ\text{C}$. This could result in a change in the potential divider current of approximately $4.4\text{mV}/750\text{k}\Omega = 6\text{nA}/^\circ\text{C}$ and hence a 6nA change in the collector current of Tr_1 , resulting in a maximum change in V_{BE1} of $250\Omega \times 6\text{nA} = 1.5\mu\text{V}/^\circ\text{C}$ and a maximum change to the short-circuit input balance condition of $100\text{k}\Omega \times 6\text{nA}/30 = 20\text{V}/^\circ\text{C}$.

Provided the temperature coefficients of

the zener diodes are not very much more than $-2\text{mV}/^\circ\text{C}$, their effect will be similar in magnitude and unlikely to deteriorate the performance of the circuit. The circuit is thus extremely stable to variations of a few degrees centigrade around the normal ambient conditions and the voltage sensitivity could be made less than 100mV provided an accurate balance could be obtained. Increasing the sensitivity to 10mV resulted in great difficulty in obtaining an accurate balance, necessitating the use of additional, finer, open- and short-circuit zero controls. The linearity of the meter scale at the top end of the frequency scale was also not quite so good. Thus, the writer settled on a minimum, practical sensitivity of 100mV .

Having successfully developed the circuit of Fig. 3 the writer felt that the circuit would benefit by a still yet increased input impedance. Consequently the circuit of Fig. 3 was developed to that of Fig. 4. It will be seen that in front of each of the transistors of the differential pair has been placed an f.e.t. in a source-follower configuration. The bases of the transistors of the differential pair are connected to tapings of the source resistors of the f.e.t.s because the gate-to-source voltage of the types used is typically -3volts at the design value of the drain current (1mA) and the writer felt it desirable to retain the positive supply voltage of 9volts rather than increase it to 12volts . The slight reduction in open-loop gain is more than compensated for by the increase in open-loop gain due to operating the differential pair at collector currents of 500A instead of $10\mu\text{A}$. Apart from the fact that the input impedance has been increased by a factor of 100 , an additional advantage is obtained by being able to reduce the number of set-zero controls to one.

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 $1\text{k}\Omega$ 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 100Ω may be placed in series with one of the $9.1\text{k}\Omega$ 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_4 . 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.

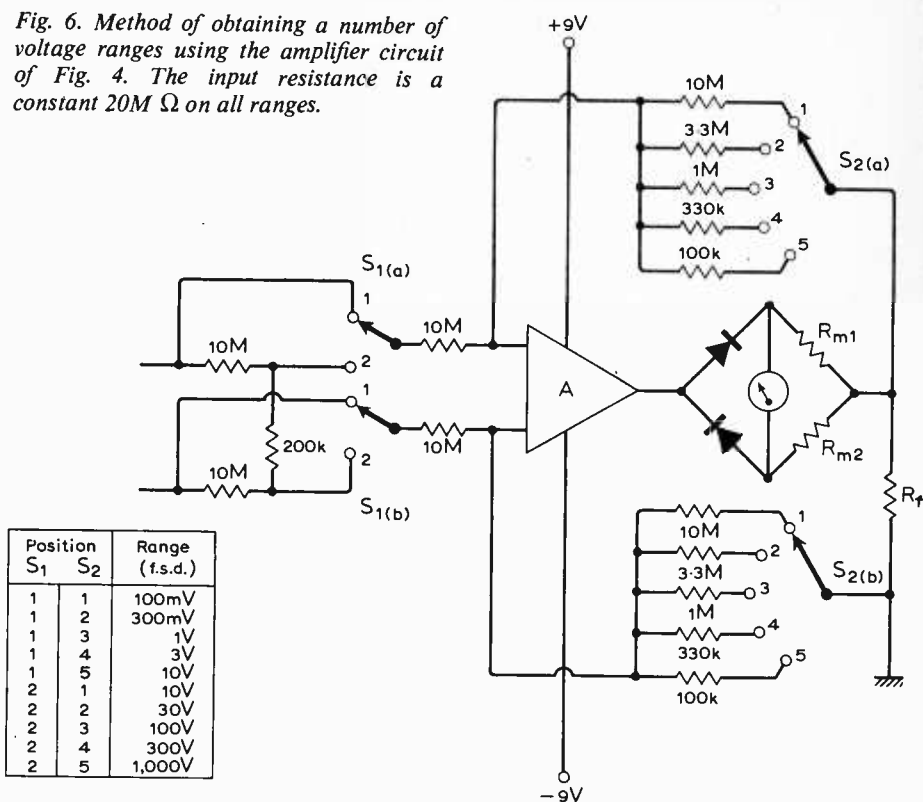
It was found necessary with the circuit of Fig. 4 to connect a capacitor and resistor, R_1 and C_1 of Fig. 4, in series between the collector and base of transistor Tr_1 , in order to prevent high-frequency instability. By trial and error, practical values for the resistor and capacitor of $5.6\text{k}\Omega$ and 470pF 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 $\pm 2\%$, 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 10M Ω 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 100 μ A meter of the Japanese 'SEW' type. The multimeter was constructed in an aluminium die-cast box of dimensions, 4 $\frac{1}{2}$ in wide \times 7 $\frac{1}{2}$ in long \times 3 $\frac{1}{4}$ in deep. The ranges provided by the attenuator are, 100mV, 1V, 10V, 100V, 1000V, 1 μ A, 10 μ A, 100 μ A, 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 R_{m1} and R_{m2} , and R_4 and R_5 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'

Fig. 6. Method of obtaining a number of voltage ranges using the amplifier circuit of Fig. 4. The input resistance is a constant 20M Ω on all ranges.



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 set-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.t.s 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

1. 'Linear Scale Millivoltmeter' by A. J. Ewins. *Wireless World*, December 1970.

Television Sound Adaptors

Methods requiring no electrical connection to the set

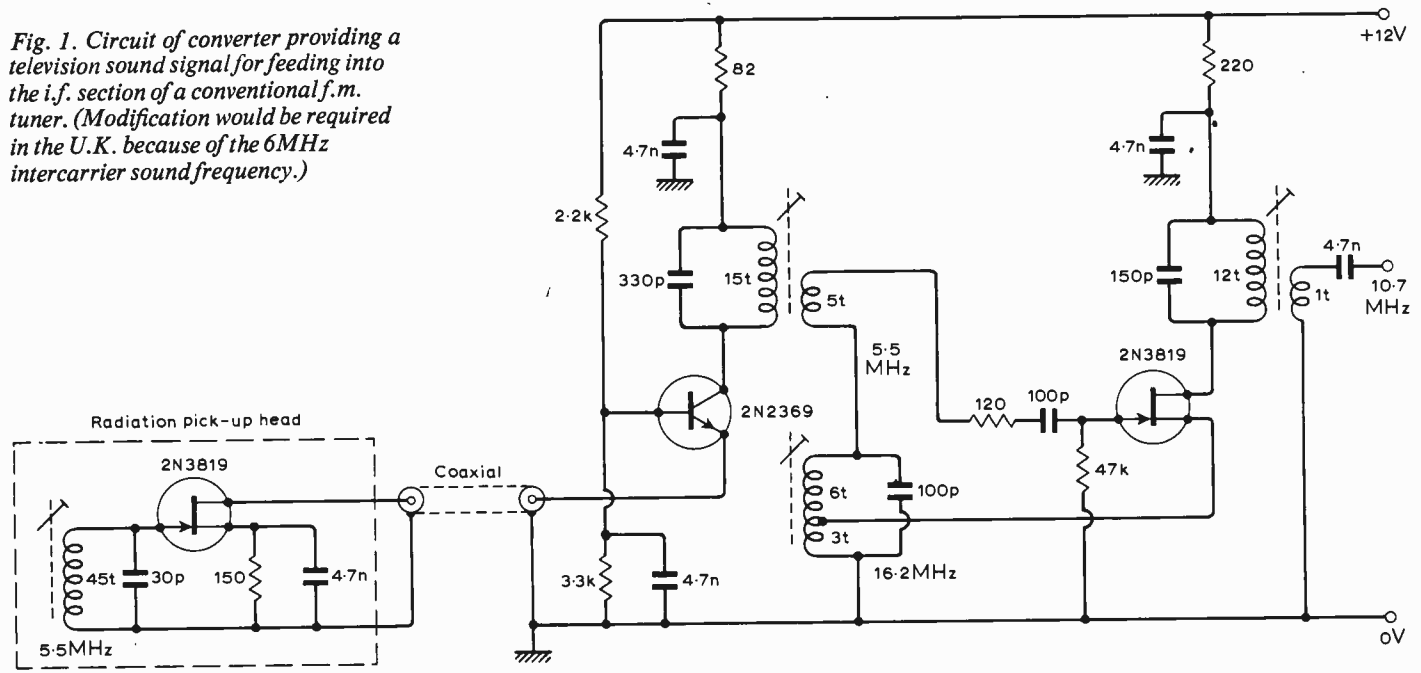
Since publication of the leader in the October 1971 issue, some letters to the Editor in the November and December issues and an article by J. C. G. Gilbert in the January issue, all on television receiver sound quality, we have heard of several methods of extracting the sound signal from the television 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

Fig. 1. Circuit of converter providing a television sound signal for feeding into the i.f. section of a conventional f.m. tuner. (Modification would be required in the U.K. because of the 6MHz intercarrier sound frequency.)



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 cascode input stage. The upper half of the cascode 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 cascode stage to give a difference frequency of $16.2 - 5.5 = 10.7\text{MHz}$, 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 switch. First the 6MHz intercarrier i.f. sound signal is amplified in a single-transistor stage, then the required band of frequencies centred 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



Commercial unit in fig. 2, showing encapsulated pick-up coil.

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 information 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 *BSI 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.

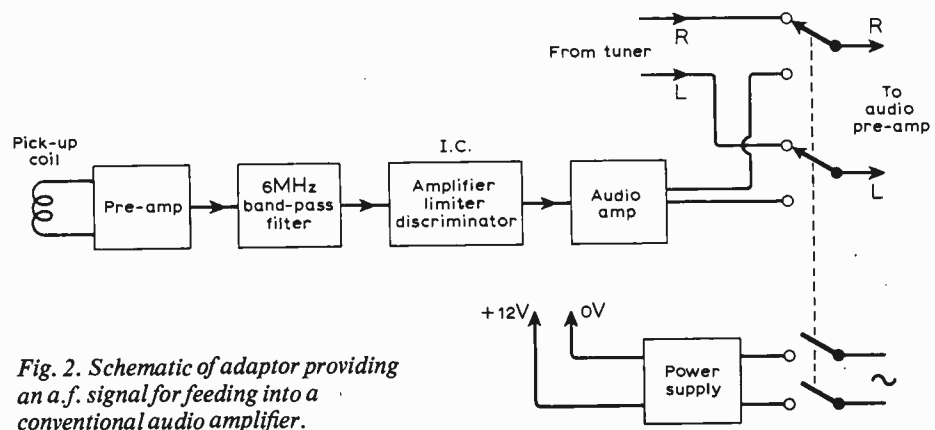


Fig. 2. Schematic of adaptor providing an a.f. signal for feeding into a conventional audio amplifier.

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 MC1466¹, 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, floats on the positive output terminal — this ensures that the only circuit component to have a high voltage across it is the series element Tr_{1,2}.

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

$$V_{out} = V_{ref} R_2/R_1$$

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_i, is compared with the drop across part of the potential divider chain formed by R₃ and R₄. If we assume that IC₂ is a perfect differential amplifier,

$$I_{lim} = (R_3 \cdot V_n)/(R_4 R_i)$$

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 R₂. (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 Tr₃. 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 Tr₃ for this transistor to be saturated, the output of IC₁ is effectively connected to the base of the series element Tr_{1,2}, 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 Tr₃ 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 Tr₃ is now an emitter follower IC₂ 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

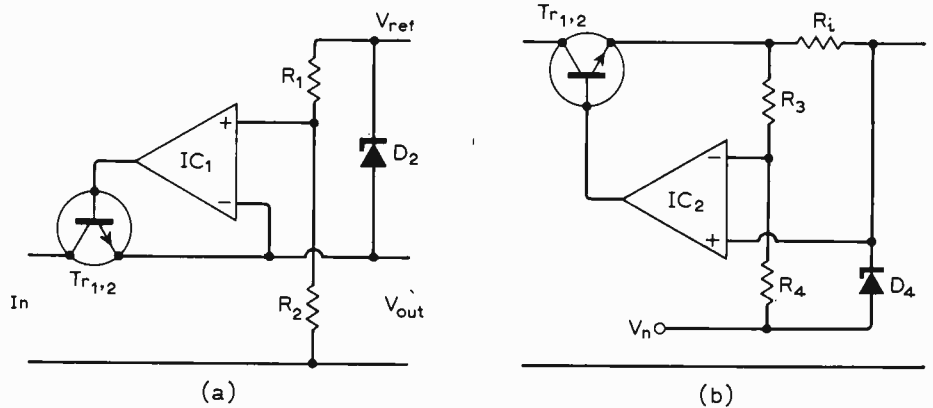
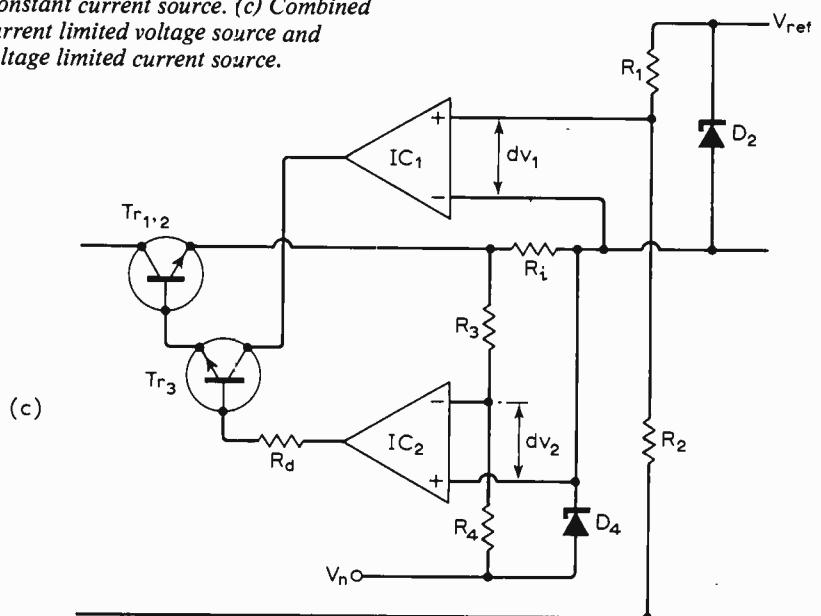


Fig. 1(a) Constant voltage source. (b) Constant current source. (c) Combined current limited voltage source and voltage limited current source.



* 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 56W. To reduce this re-entrant foldback protection has been incorporated by including Tr_4 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_7 flowing through R_5 causes Tr_4 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

TABLE 1. Performance

Output voltage	Fully variable, two ranges	0-25V, 25-50V
Output current	Fully variable, two ranges	0-100mA, 0-1.0A
Load regulation*	$\Delta I = 1.0A$	<0.03%
Current regulation	Load current = 1.0A	<0.1%
Line regulation*	10% mains variation	<0.03%
Ripple and noise	Constant voltage mode, $I = 1.0A$	<2mv
	Constant current mode	<0.03%
Temperature coefficient	Voltage	0.01%/deg C (typ.)
	Current	0.06%/deg C (typ.)

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

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_8 to parallel the current setting potentiometer, R_{20} . Thus the short-circuit current is defined by this resistor combination. It is important that R_6 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_1 .

Fig. 3 shows all the internal supplies. The mains transformer supplies 30V or 48V a.c., selected by the range switch S_{3b} , which is rectified and smoothed by D_6 and C_6 respectively. The positive and negative supplies, which float with the

TABLE 2.

Symbols used	Typical value	
A	Voltage gain of differential amplifier	10 ³
dV	Voltage between the input terminals of the differential amplifier	
I	Load current	
I_{lim}	Load current at limiting	
L	Load regulation (%)	
V_n	Negative reference voltage	8.2V
V_o	Output voltage with no load current	
V_{out}	Output voltage with load current I	
V_{ref}	Positive reference voltage	5.1V
R	Output resistance of differential amplifier	75Ω
R_D	Base resistance of Tr_3 (R_1)	15kΩ
R_1	Current sensing resistor	1 or 10Ω
β_2	Current gain of $Tr_{1,2}$	6000
β_3	Current gain of $Tr_{1,2,3}$	600,000
ΔV	Change in output voltage when load current changes by I amps	

positive output of the circuit, are given by an additional secondary. To obtain a ripple-free reference voltage D_8 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

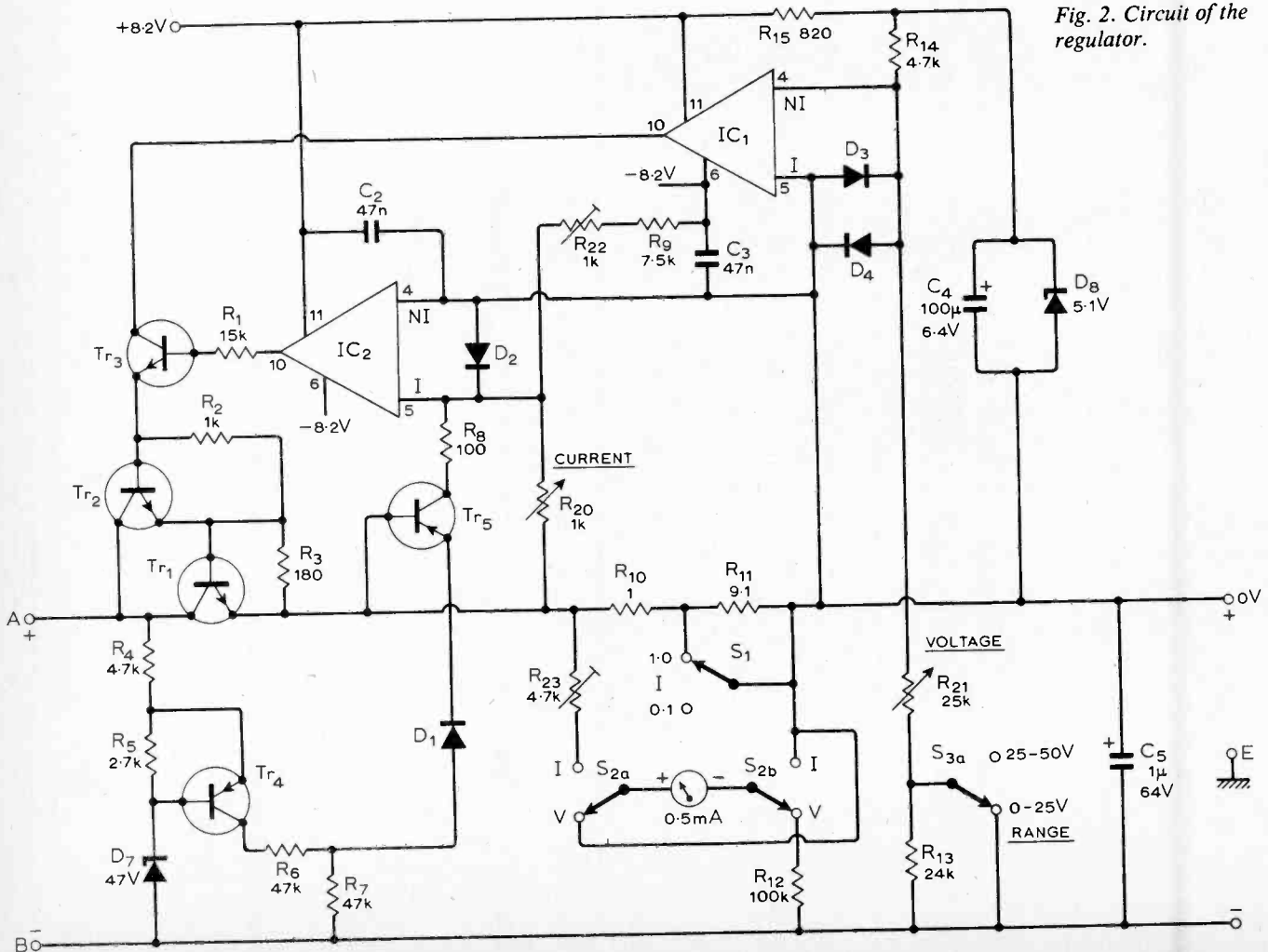


Fig. 2. Circuit of the regulator.

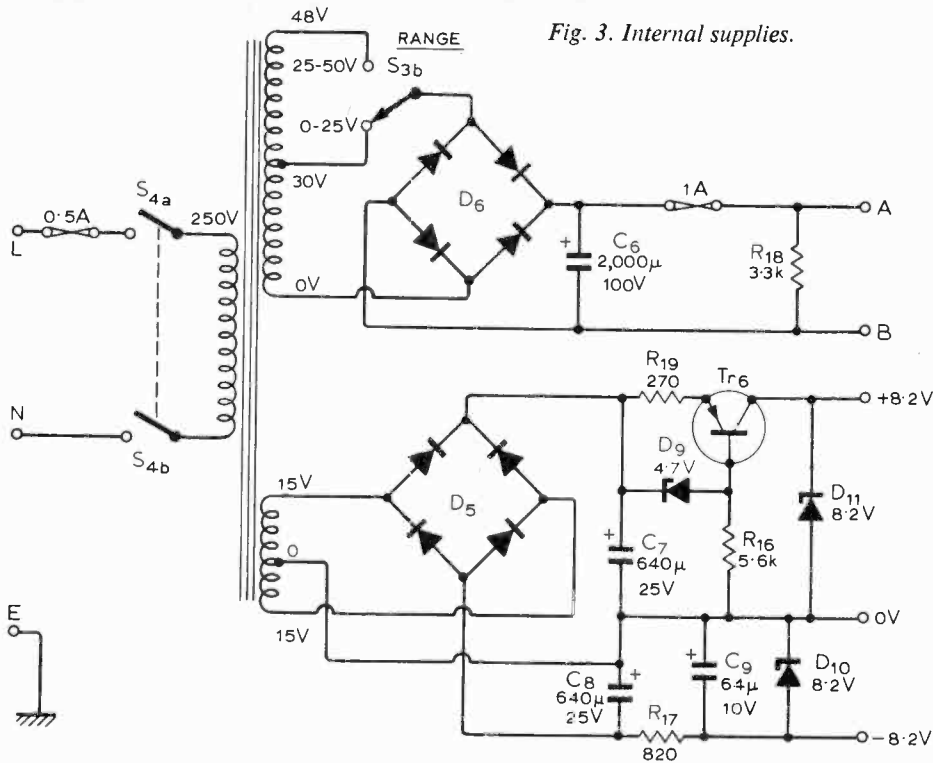


Fig. 3. Internal supplies.

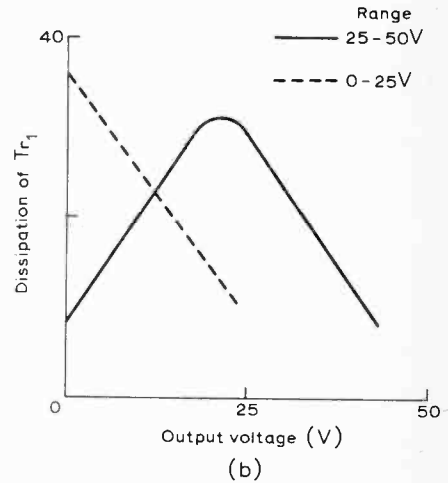
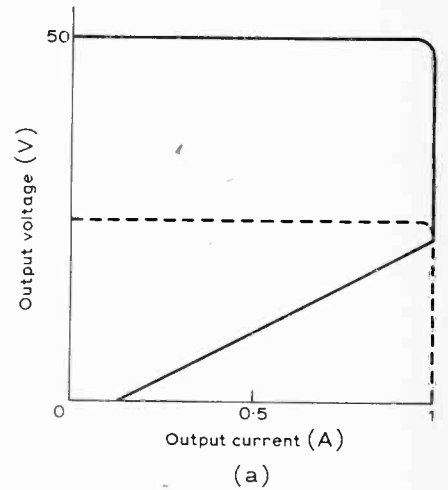
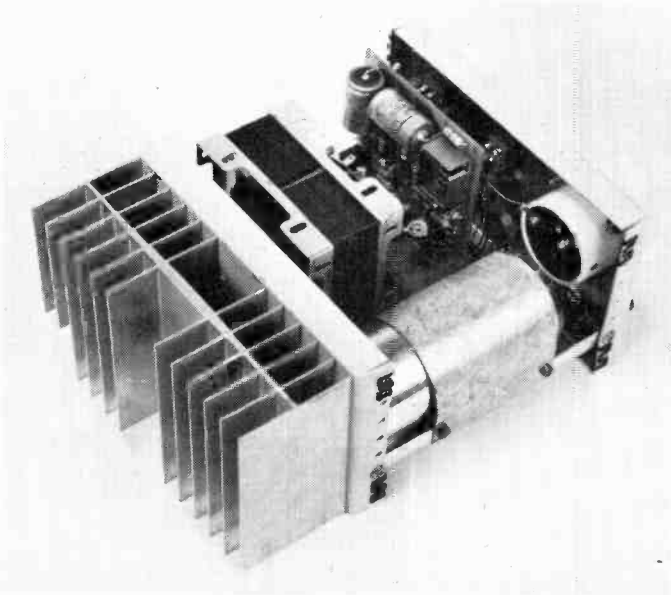


Fig. 4. (a) Output characteristic. (b) Tr_1 dissipation (W).

Completed prototype, showing the layout of controls.



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

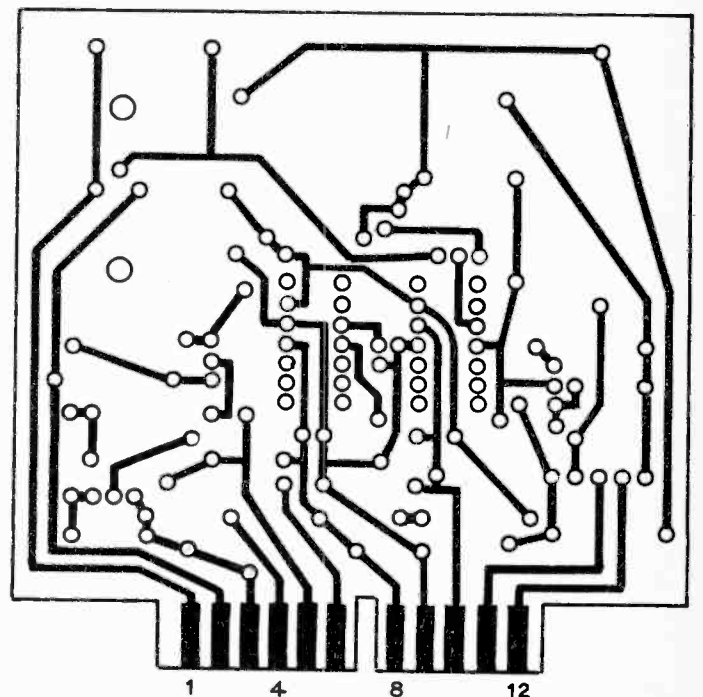


Fig. 5. P.c. board, copper side.

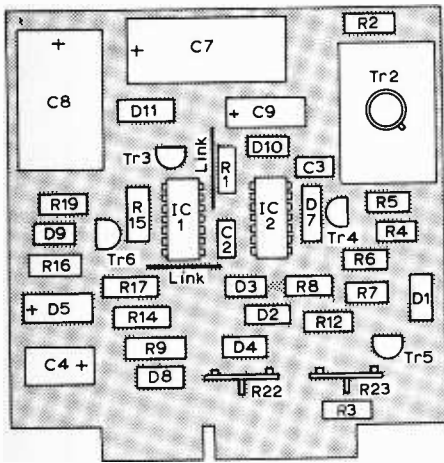


Fig. 6. Layout of p.s. board.

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 (0.06% per °C against 0.007% per °C). If this is unsatisfactory, D_9 could be replaced by one similar to D_8 , giving a common mode voltage of about -2V to IC_1 and IC_2 . This would probably not affect circuit operation.

To give the higher voltage range S_{3a} switches an extra resistor (R_{13}) into the voltage sensing chain and the two current ranges are implemented by altering the current sensing resistor from 1Ω to 10Ω by means of S_1 . The last switch, S_2 , 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_{20} varies the limiting current. Adjustment of R_{21} 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

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 $Tr_{1,2}$ is about 72V, so the ubiquitous 2N3055 with its V_{CER} of 65V would not be a good choice. The RCA 40636 is an improved version with a V_{CER} of 95V. Transistor Tr_2 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 ($\beta=30$ at 1.0A gives about 1.6W).

Diodes D_2 , D_3 and D_4 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 directly. Resistor R_{13} should be returned carefully to the negative output terminal, and pin 9 of the edge conductor, together with the current sensing resistor R_{11} , 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 note², 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.
3. *Review of Scientific Instruments*, Vol. 42, page 797 (1971).

Component list

Resistors

1	15k	11	9.1†
2	1k	12	100k.
3	180	13*	24k
4	4.7k	14	4.7k
5	2.7k	15	680
6	47k	16	5.6k
7	47k	17	820
8	100	18	3.3k (3W)
9	7.5k	19	270
10*	1 (3W/5%)		
all	$\frac{1}{2}W$, 5% unless otherwise shown		
20*	1k lin.	21*	25k lin.
22	1k lin. preset vertical mounting		
23	4.7k lin. preset for p.c.b.		
	† for 9Ω parallel with 820Ω		

Capacitors

2	0.047μ/20V ceramic	6*	2000μ/100V (Hunts)
3	0.047μ/20V ceramic		
4	64μ/10V (Mullard)	7	640μ/25V (Mullard)
5	1μ/64V (Mullard)	8	640μ/25V (Mullard)

C_1 has not 'disappeared' but was removed as a design improvement made by the author.

Integrated circuits

1 and 2 SN72741N (the dual 741s, SN72741D or L147B1 (Fairchild) could be used)

Transistors

1*	40636 (RCA)	3	BC183L
2	40409 (RCA)	4, 5, 6	BC212L

Diodes

1, 3, 4	1S921	8	HS7051
2	1N914		(Emihus)
5	BS05 (IR)	9	1S2047A
6*	IB08T20	10, 11	HS7082
7	1S3047A		(Emihus)

all semiconductors are Texas unless otherwise shown

Transformer* 0-30-48V/1A Douglas MT 126 is suitable

Wind on another secondary 15-0-15 by adding 2 × 75 turns of 38 s.w.g. wire.

Switches*

1	s.p.d.t. WEL TS106D recommended
2, 3, 4	d.p.d.t. WEL TS206N recommended.

S_{3a} must be able to withstand the charging surge of C_7 .

Meter* 0.5mA (re-scale 0-50V, 0-100mA, 0-1.0A).

Edge connector. 12 way, 0.15in. contact pitch.

Heatsink* Marston 50DN0400 A100 recommended. This has a thermal resistance of about 1 deg C/W but any other may be used if it has a thermal resistance of less than 1.75 deg/W. The usual 4 × 4 × 1in. sink is not suitable unless the maximum current is limited to less than 0.8A (maximum ambient and junction temperatures of 40°C and 175°C respectively).

* Not mounted on 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.

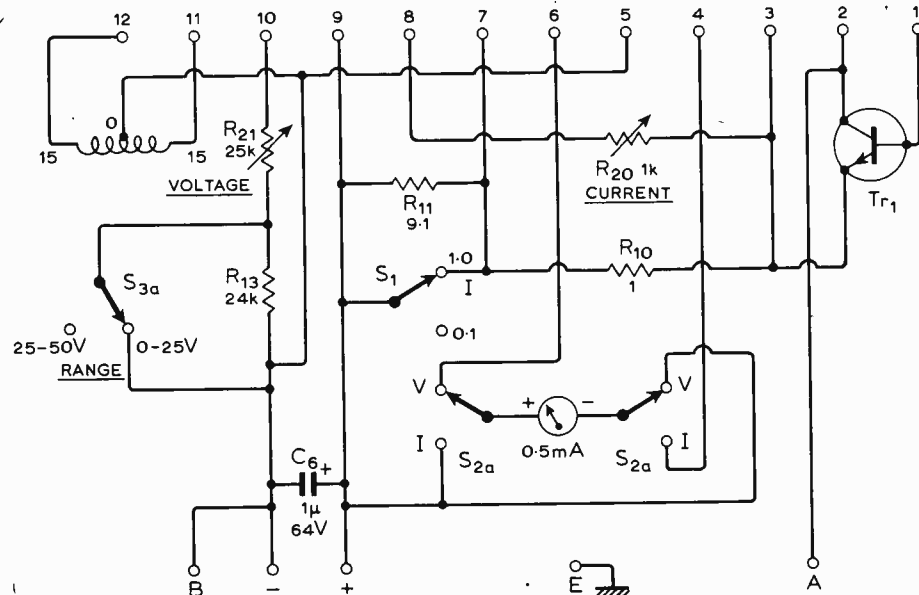


Fig. 7. Connections to p.c. board.

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 $Tr_{1,2,3}$ are ignored.

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

With no load current the output voltage is,

$$V_o = V_{ref} \cdot R_2/R_1$$

If the current is drawn from the supply an error voltage is developed across the input of IC_1 to compensate for the drop over R_1 and R so,

$$V_{out} = ((V_{ref} - dV_1) \cdot R_2/R_1) - dV$$

where

$$dV_1 = \frac{I(R_1 + R/\beta_2)}{A}$$

Thus the change in output voltage is

$$\Delta V_1 = V_o - V_{out} = \frac{I(1 + R_2/R_1)(R_1 + R/\beta_2)}{A}$$

The load regulation, L is defined as

$$L = \frac{\Delta V_1}{V_o} \times 100\% = \frac{I(R_1 + R/\beta_2)(1 + R_2/R_1)}{AV_{ref}} \times 100\%$$

and the output resistance is

$$R_{out} = \frac{\Delta V_1}{I} = (R_1 + R/\beta_2)(1 + R_2/R_1)/A$$

We can calculate the performance using the typical parameters:

Current Range	0.1		1.0		
Output voltage	5	50	5	50	V
L	0.0004	0.0002	0.0005	0.0003	%
R_{out}	0.20	1.1	0.022	0.12	mΩ
ΔV_1	0.02	0.11	0.022	0.12	mV

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,

$$\text{now } dV_2 = I_{lim}(R_1 + R_D/\beta_3)/A$$

and at limiting the drops across R_1 and R_3 differ by dV_2

$$\text{so } I_{lim}R_1 = (I_{lim}R_1 + V_n)R_3/(R_3 + R_4) + dV_2$$

after some re-arrangement

$$I_{lim} = \frac{R_3 V_n}{R_4 R_1 \left(1 - \frac{R_3 + R_4}{AR_4} (1 + R_D/\beta_3) \right)}$$

For the simple case $A = \beta_3 = \infty$

$$I_{lim} = R_3 V_n / R_4 R_1$$

so the actual current is higher than that given above by

$$(1 + R_D/\beta_3)(R_3 + R_4)/AR_4 \times 100\%$$

Since $R_4 = 8R_3$ ($I_{lim} = 100\text{mA}$ 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 out-

put 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,

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

Since the current and voltage are linearly dependent on the values of R_{20} and R_{21} these potentiometers could be replaced by helipot or decimally switched resistors.

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

Nelson Tansley Ltd, who specialize in telecommunications systems for railways, are now marketing these systems internationally under the registered trade name ENTEL.

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 as exclusive agent and distributor throughout France.

Guest International Ltd., of Thornton Heath, Surrey, have acquired the U.K. agency for the Lintronix 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 Servomex 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 die-stamped foil strain gauges and temperature sensors manufactured by Dentronics Inc., of New Jersey, U.S.A.

Data Recognition's ranges of Dataterm 3 and Dataspec 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 £15M contract for final development, supply and installation of large electronic exchanges of the TXE4 design with Standard Telephones and Cables Ltd.

Pye TVT Ltd have received the first u.h.f. colour television transmitter order, valued at £30,000, 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 Ilderton Road, London SE15 1NS. (Tel. 01-639 7844.)

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 "Remotocopies" 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.

Emihus 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 Mostek 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.S.A. 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-Worms International Ltd, Hodford House, 17-27 High Street, Hounslow, Middx, have recently been appointed U.K. distributor for the range of electronic components supplied by Thomson-CSF United Kingdom Ltd.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

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 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

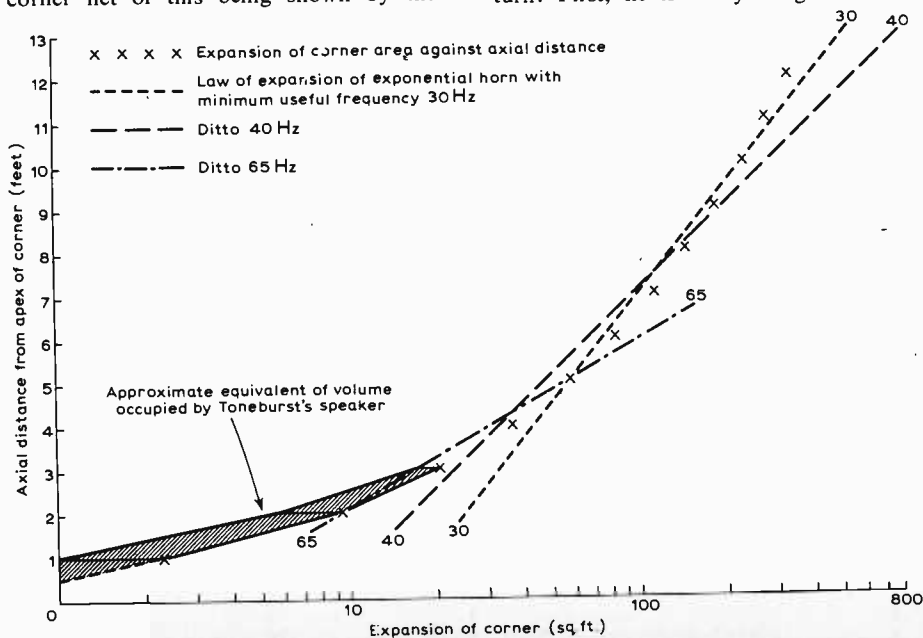
believes plastic foam has no acoustic function! For a start there are many types of foam rubber or foam plastic — latex, polyurethane and polystyrene being only three of the better known ones. Depending on the process of manufacture these are again subdivided into many types, and each type has its own 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/felt 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 10^8 - 10^9 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.

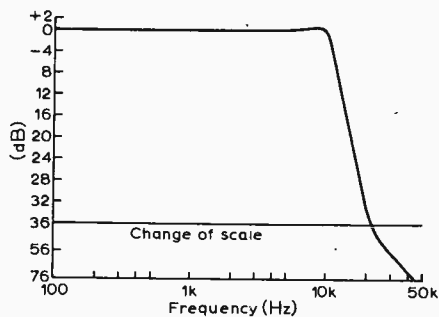
In conclusion may I thank Mr. Halliday for his interest in the article and the many points he has brought up.
PHILIP D. HARVEY,
 Chelmsford, Essex.

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 Ω input element.
R. CURLING-HOPE,
 Johannesburg,
 South Africa.

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 backlash (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.

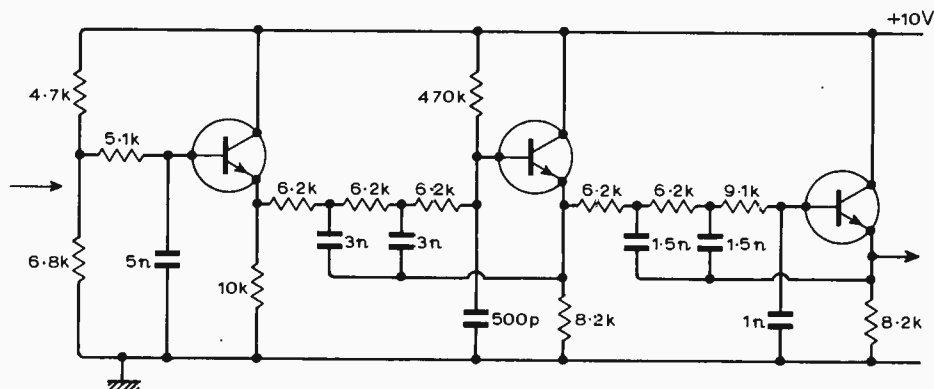
1. Cole, H. A. 'A single-channel pulse-height analyser with 100nsec resolution', *Nuclear Instruments & Methods* 84 (1970), pp93-101.
2. Cole, H. A. 'The use of integrated-circuit amplifiers to provide variable back bias in single-channel pulse-height analysers'. *Nuclear Instruments & Methods* 79 (1970), pp356-358.

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. WILLOUGHBY,
 Grafton,
 Balaclava 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 shown 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

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

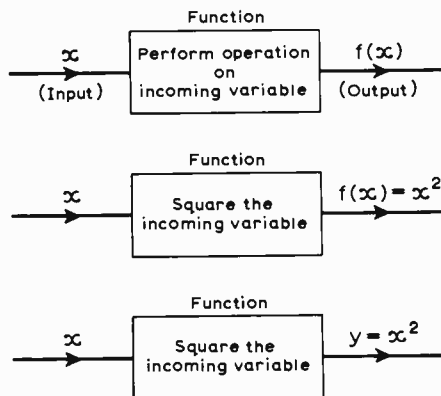


Fig. 1 Schematic representation of a function generator. The output variable is a function of the input variable.

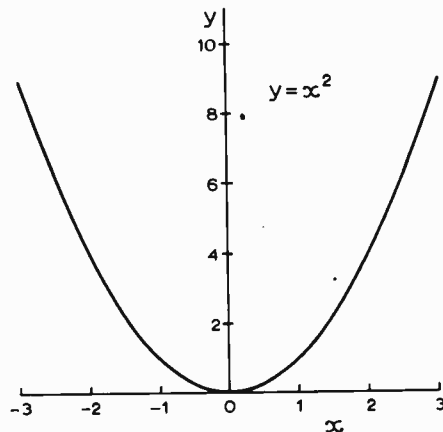


Fig. 2. Graphical representation, or 'model', of the equation $y = x^2$, as given by the function generator in Fig. 1.

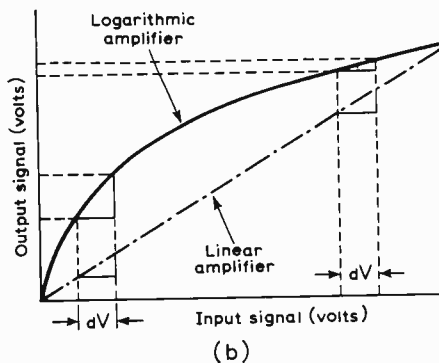
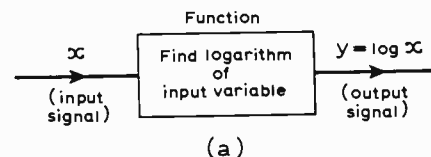


Fig. 3. Function generator with a logarithmic characteristic (a); and output/input characteristic of a logarithmic amplifier compared with that of a linear amplifier (b). A given change of input signal, dV , results in different changes of output signal, depending on where it occurs on the log. curve.

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.

†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.

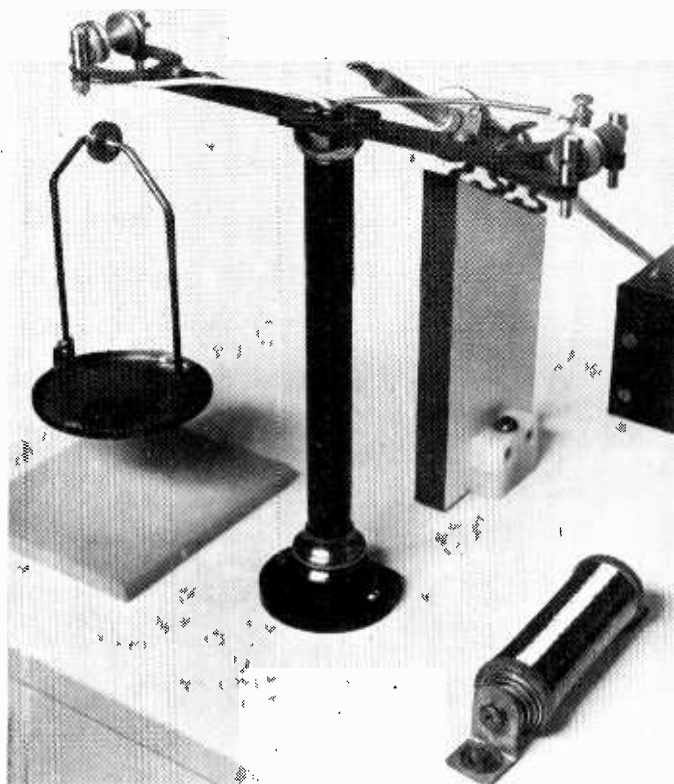
Previous work on the rate of solution of metals in solder has been based on the rate of wear of a soldering iron bit under simulated practical conditions¹, or on the reduction of diameter of wires immersed in solder^{2,3}. The former method is entirely empirical and the results are applicable only to 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

Fig. 1. General view of test equipment.



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 $272 \pm 17^\circ\text{C}$, $318 \pm 9^\circ\text{C}$, $370 \pm 7^\circ\text{C}$, and $430 \pm 10^\circ\text{C}$. The wires tested were soft-drawn o.f.h.c.+ copper of diameter 0.067

+Oxygen-free high-conductivity.

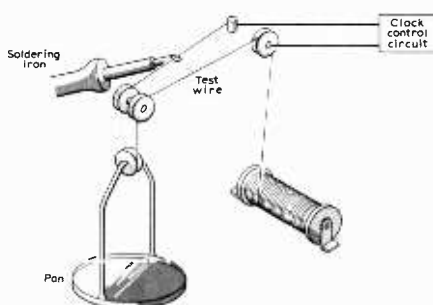


Fig. 2. Diagram of test equipment.

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

* Senior works chemist, Multicore Solders Ltd.

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 Beal³, 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; $\sqrt{T} = kd$

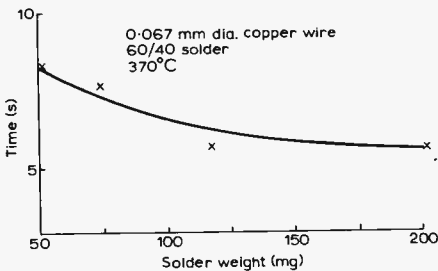


Fig. 3. Influence of solder weight on time to breakage.

In the present method, timing stops when the wire can no longer support the 50g load of the pan. The diameter d_0 at this moment can be estimated by measuring the load w required to break the wire at the same temperature without solder on the bit; then

$$d_0 = d\sqrt{50/w}$$

This was measured rather roughly by running water into a beaker on the pan, stopping when the wire broke. Using this value of d_0 and two different values of the initial diameter d gives the three points required to define the straight line, which should have a negative intercept on the \sqrt{T} axis since timing stops before the wire

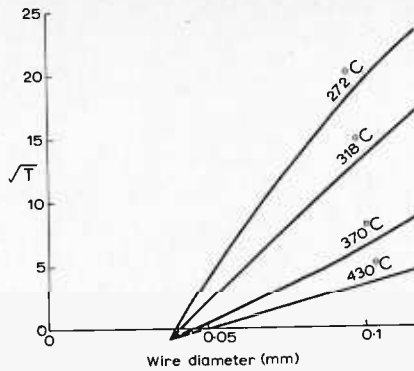


Fig. 4. Relation of breakage time to wire diameter in 60/40 solder.

diameter is reduced to zero. Fig. 4 shows that this condition is approximately realised despite the considerable variability in individual results. However, the equation

$$\sqrt{T} = kd - C$$

implies that the lines for the four different temperatures should make the same intercept on the \sqrt{T} axis. Fig. 4 shows that in practice, the lines meet at a vertex to the right of this axis. This probably means that some reduction of diameter occurs not by solution but by stretching of the wire under the 50g load. Though this prevents accurate calculation of the absolute rate of solution and its dependence on temperature, it does not invalidate comparisons between the very large differences in breakage time found for different solders.

Comparison of solders

Table 1 gives the composition of solders used. The test results are shown in Fig. 5 as a function of temperature, and show that the rate of attack is strongly dependent on the tin content. Low-tin solders have been recommended for the soldering of fine copper wires⁴ but it should be remembered that reduced tin content gives not only a reduction in rate of attack, but also a reduction in wetting power and an increase in the temperature required to make the joint — choice of solder must not be made on the basis of one property only. By contrast, Savbit solder has a wetting power similar to 60/40 and a liquidus temperature only 25°C higher, but a fine copper wire will withstand it for between 50 and 100 times longer depending on the temperature.

According to the equilibrium diagram for tin-copper⁵, the liquidus for 97 tin 3 copper is 318°C; in practice, partly due to the presence of lead (with which copper is immiscible) and partly because equilibrium is not reached or maintained (the solid Cu_6Sn_5 compound being slow to redissolve

and the temperature not being constant) complete re-solution of the Cu_6Sn_5 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

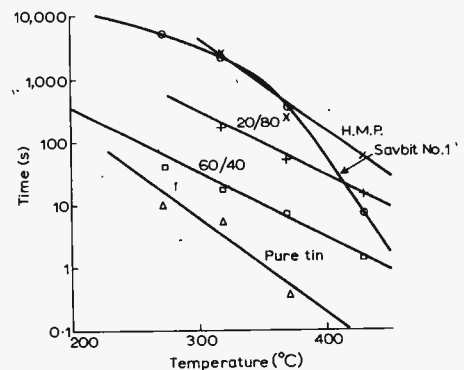


Fig. 5. Breakage time of 0.067mm copper wire in various solders as a function of temperature.

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.

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2. Bader, W. G. — 'Dissolution of Au, Ag, Pt, Cu and Ni in molten tin-lead solder' *Welding Journal* research supplement Dec. 1969, p.551.
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Table 1: Characteristics of solders tested.

solder	approval or standard	melting range. °C	% tin	% copper	% silver
Pure tin	BS 3252, Type T2	232	99.8	0.01	—
60/40	BS 219, Grade K	183 - 188	59.1	0.005	—
20/80	BS 219, Grade V	183 - 277	19.2	0.005	—
H.M.P.	BS 219, Grade 5S	296 - 301	4.8	n.d.	1.63
Savbit No. 1*	DTD. 900/4535	183 - 215†	50.3	1.50	—

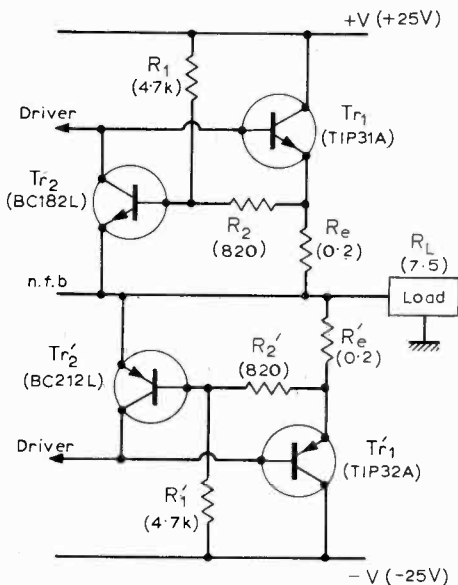
*Registered trade mark. Multicore Solders Ltd.

†As indicated by cooling curve. The temperature at which the dispersed Cu_6Sn_5 dissolves is much higher, see text.

Circuit Ideas

Output transistor protection in class B amplifiers

The frequently used method of protecting output transistors Tr_1, Tr_1' by using the voltage developed by the output current across sensing resistors R_e, R_e' to turn on base bypass switches Tr_2, Tr_2' , can be improved by the addition of resistors R_1, R_1' from the bases of Tr_2, Tr_2' to the power lines. The result is that the peak output current available is reduced as the impedance R_L 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_L in most cases. If the load were purely resistive then near optimum results would be obtained by resistors satisfying the equations

$$R_e = 0.6/I \text{ and } R_1/R_2 = R_L/R_e,$$

and R_2 could be chosen to be around 100Ω . 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_L 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.

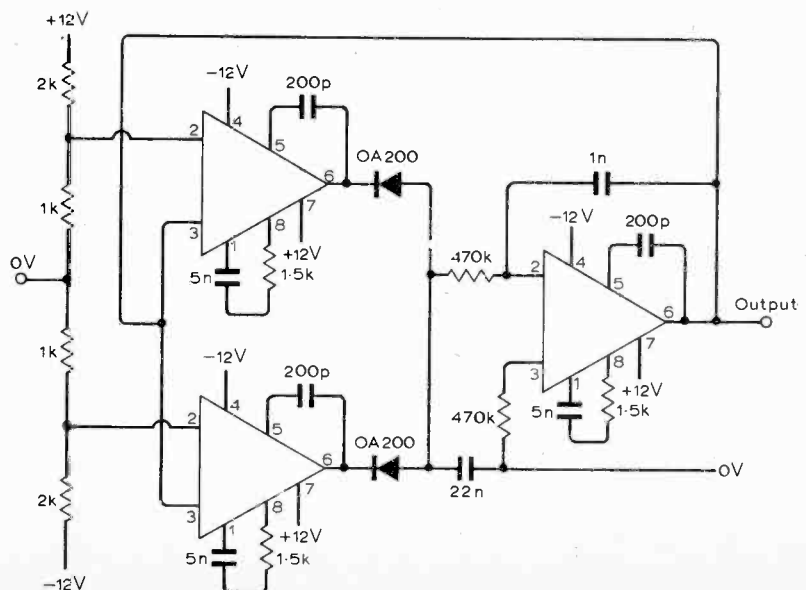
The circuit is equally suited to dual polarity or single polarity power supply amplifiers; in the latter case replace V in the calculation of R_e by half the supply voltage. The circuit cannot be used with class A output stages as it causes the output transistors to be cut-off for most of alternate half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

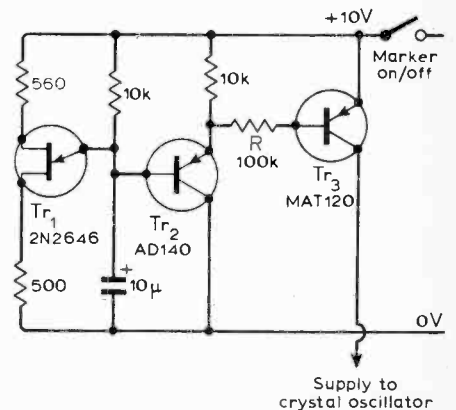
This circuit uses three MC709 amplifiers to generate a triangular waveform, the amplitude of which may be independently varied either side of zero by adjusting the



input voltages to the first two amplifiers. The frequency is determined by the $470k\Omega$ 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.
H. MACDONALD,
London W.3.

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 Tr_3 and earth

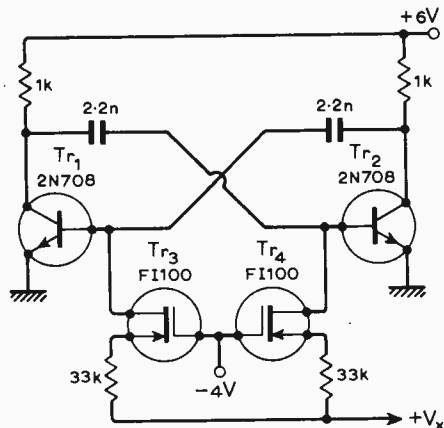
should show a very small fluctuation about a value approximately 1V below supply voltage. This arrangement has the advantage of enhancing marker recognition in the 18 to 30MHz band where high order harmonics of the crystal are comparatively weak.

GWILYM DANN,
Chipstead,
Surrey.

Stable voltage-to-frequency converter

When a conventional free-running multi-vibrator 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 $Tr_{3,4}$ 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 FI100 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.

Circuits specified in the text	Percentage departure of frequencies	
	at 60°C	at 100°C
a	0.9	2.0
b	2.1	4.5
c	0.2	0.6

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.

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 1k Ω load is 15V peak, and the tone controls have an impedance of about 10k Ω in the flat position, so that the maximum additional load is limited to about 1k Ω for 3V r.m.s. at clipping.

The circuit is equivalent to a one-transistor stage but distortion is reduced by operating Tr_2 in common-base mode and Tr_3 is operated at an almost constant collector voltage, hence reducing 2nd harmonic distortion. Transistor Tr_1 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 10k Ω input impedance.

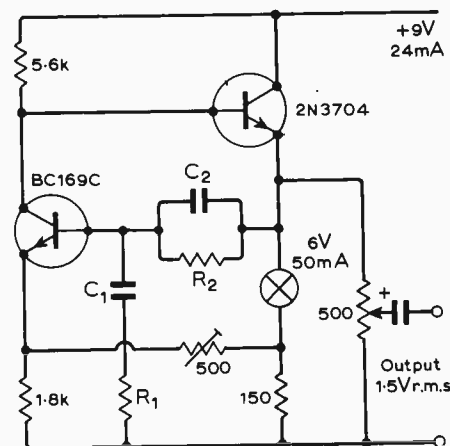
The presence of C_1 may cause some l.f. instability and the value may need adjustment to allow the collector voltage of Tr_2 to rise in a critically damped manner, especially if the supply line has a high impedance for some reason. Omission of C_1 and combining the 51k Ω resistors to one of 100k reduces the gain to \approx 14dB and also reduced the dynamic tone control range. H.F. overshoot can be

eliminated by either a 1nF capacitor across the 6.8k Ω resistor or an 820 Ω plus 2.2nF from Tr_2 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.

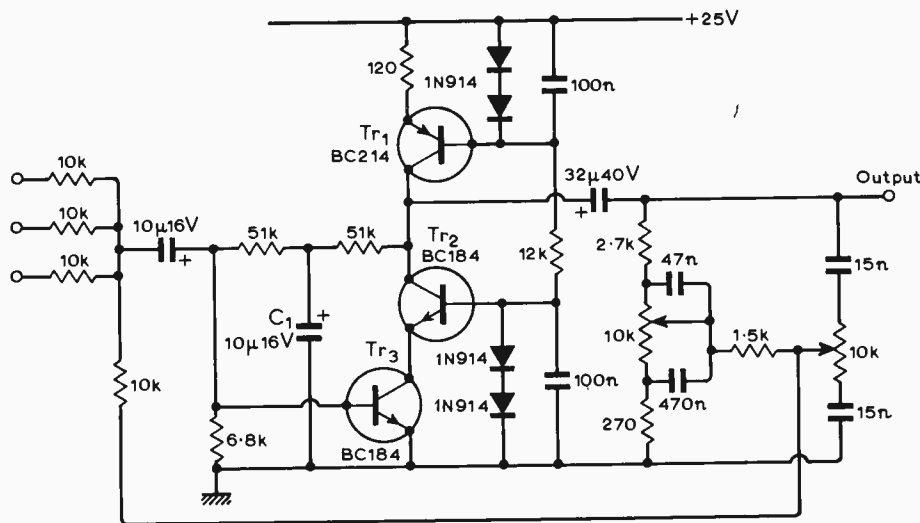
Inverted Wien bridge oscillator

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

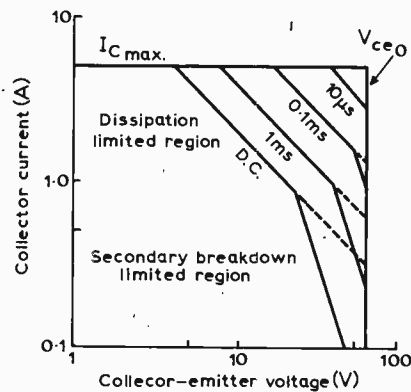


Fig. 1. Areas of safe operation for a small power transistor.

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.¹ 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

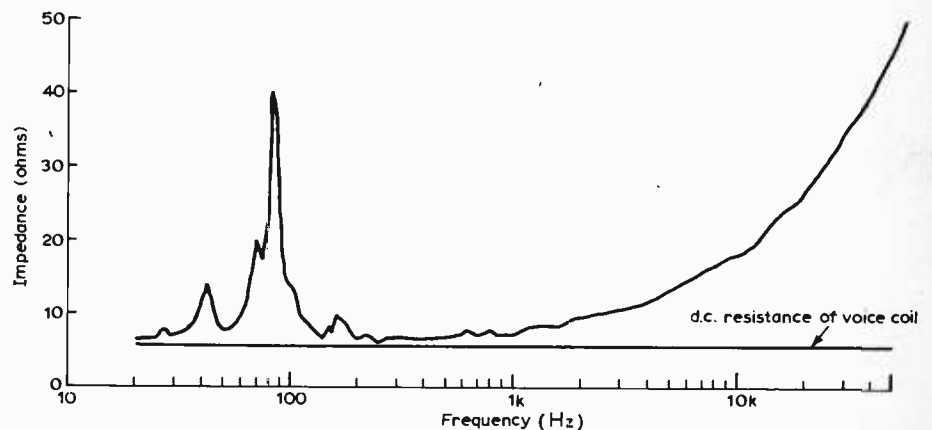


Fig. 2. Impedance/frequency relationship for a typical 8Ω speaker.

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_{cc} 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 6.5 Ω .

Taking the worst case condition of 5.5 Ω 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

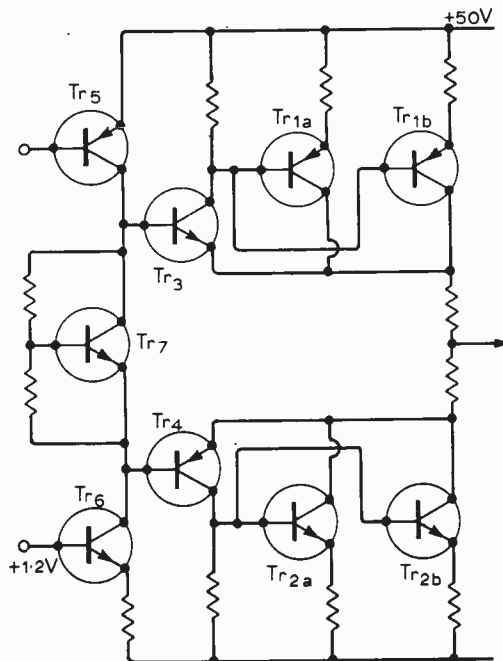


Fig. 4. Basic output stage used in 100W amplifier.

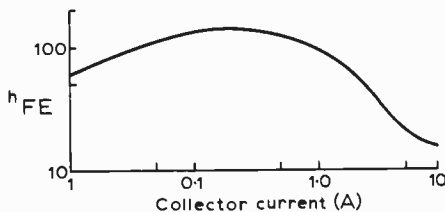
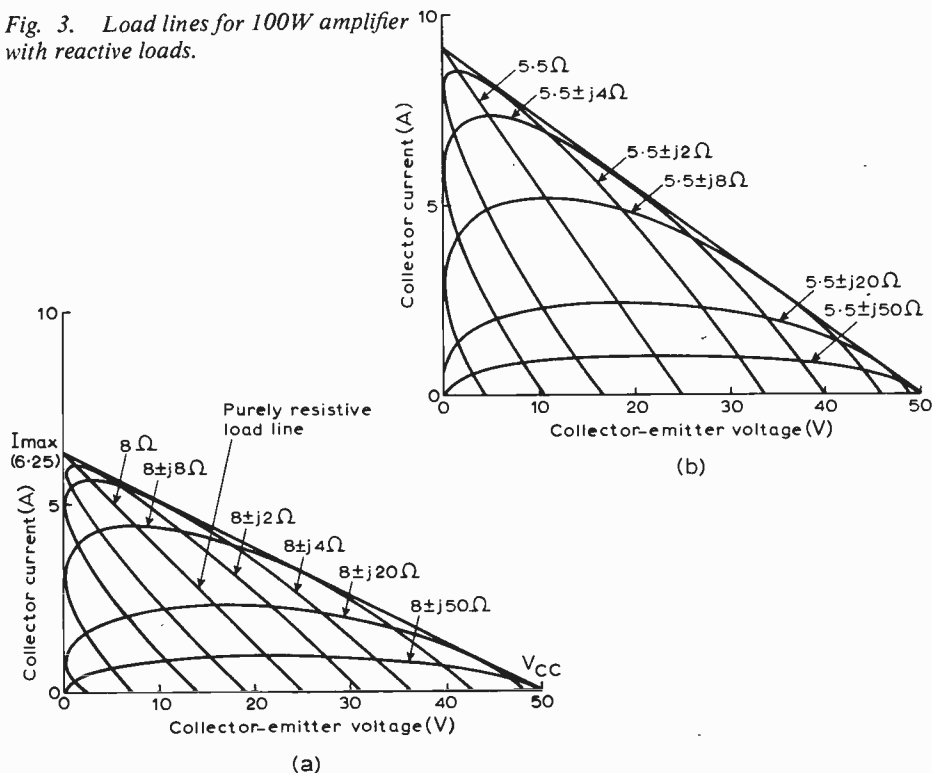


Fig. 5. Typical static forward current transfer ratio for TIP33A.

Fig. 3. Load lines for 100W amplifier with reactive loads.



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 $Tr_{3,4}$, in Fig. 4 we need to know the gain of the output devices $Tr_{1a,b}$, $Tr_{2a,b}$ at different currents which is shown in Fig. 5. Transistors Tr_1 and Tr_3 have approximately the same collector-emitter voltage, while the collector current of Tr_3 is approximately equal to the base current of $Tr_{1a,b}$. Thus the peak dissipation in Tr_3 will be equal to the dissipation in Tr_1 divided by the h_{FE} of Tr_1 . 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_3 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_3 for peak signals, i.e. 7.3A load current as will occur when there is 80V p-p being applied to a 5.5 Ω 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_1 is 13 at 7.3A load current. As a result 590mA is being demanded from Tr_3 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_6 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_5 , 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

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: Tr_1 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. Bailey². In this circuit R_1 produces a voltage proportional to the emitter current of Tr_1 . When this is great enough, the voltage across R_5 is sufficient to turn on Tr_2 removing further drive current from the base of Tr_1 . Also fed to the base of Tr_2 is a current from R_4 proportional to the voltage between the supply line and the output line (a close approximation to V_{ce} of Tr_1). In this way the emitter current of Tr_1 required to turn on Tr_2 , is reduced for increasing values of V_{ce} of Tr_1 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_3 is required to protect each protection transistor from the otherwise large currents which would flow down it from the turned-on Tr_2 of Fig. 4, down to the shorted output line. Unfortunately, R_3 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_3 . This drop will also limit the voltage swing of the amplifier, when maximum power is being delivered, making undesirable a high valued R_3 . Looking at the bottom half of the output stage the problem is already solved for us by Tr_6 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 Tr_5 . Transistor Tr_6 always draws 25mA from Tr_5 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

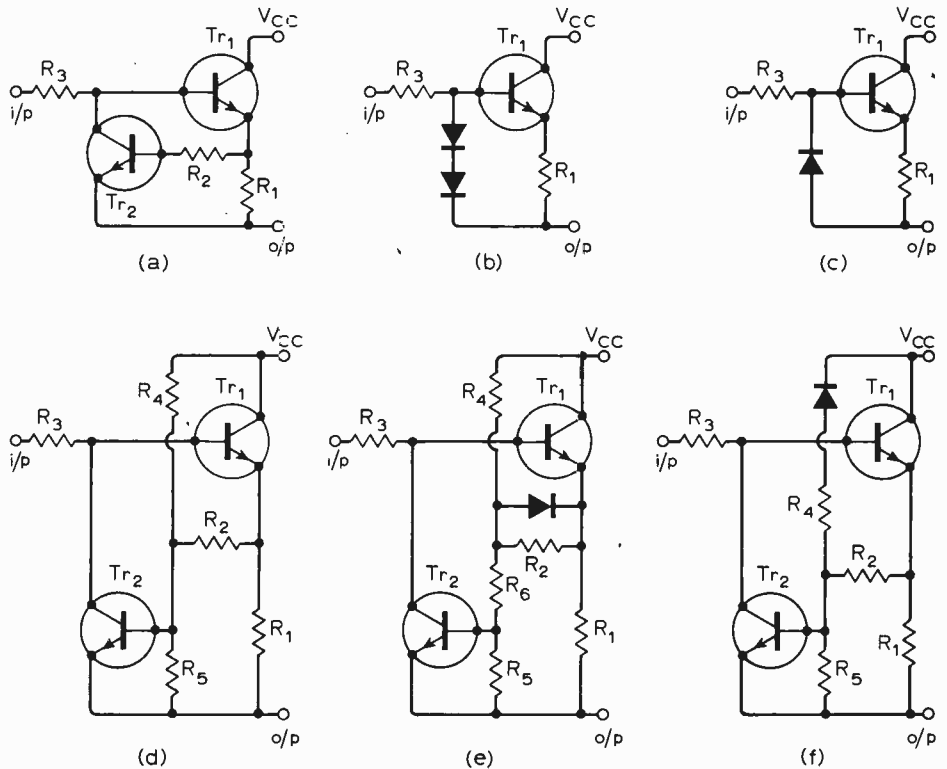


Fig. 6. Various transistor protection arrangements.

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 Tr_5 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 Tr_5 is sufficient to turn on the diodes preventing any further increase in current. Because of their well defined forward voltage drop 1S 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 pro-

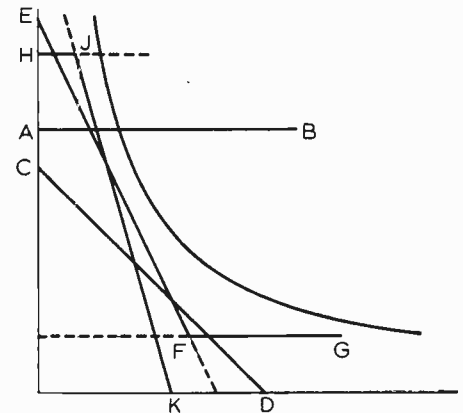


Fig. 7. Boundaries of transistor operation defined by circuits in Fig. 6.

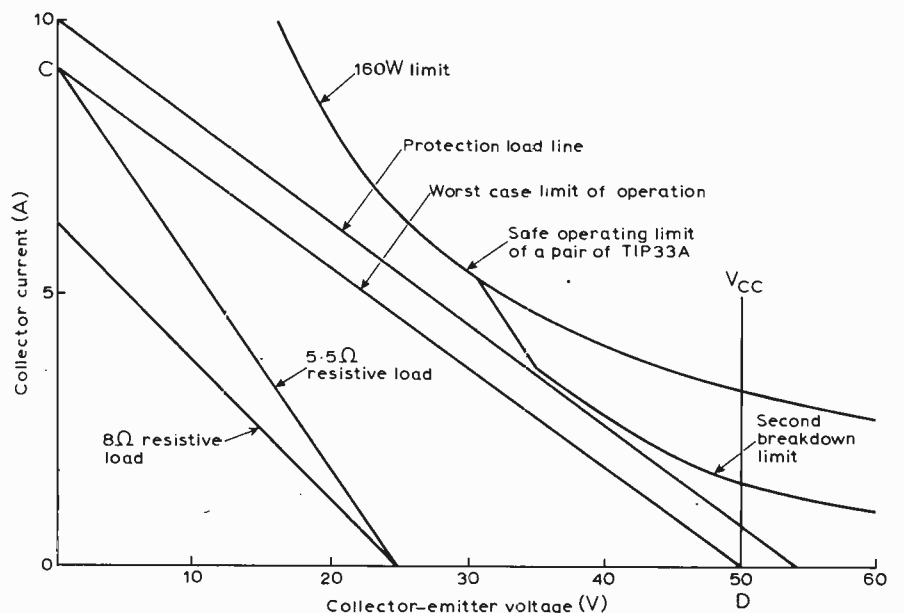


Fig. 8. Safe operating limits for a pair of TIP 33A transistors.

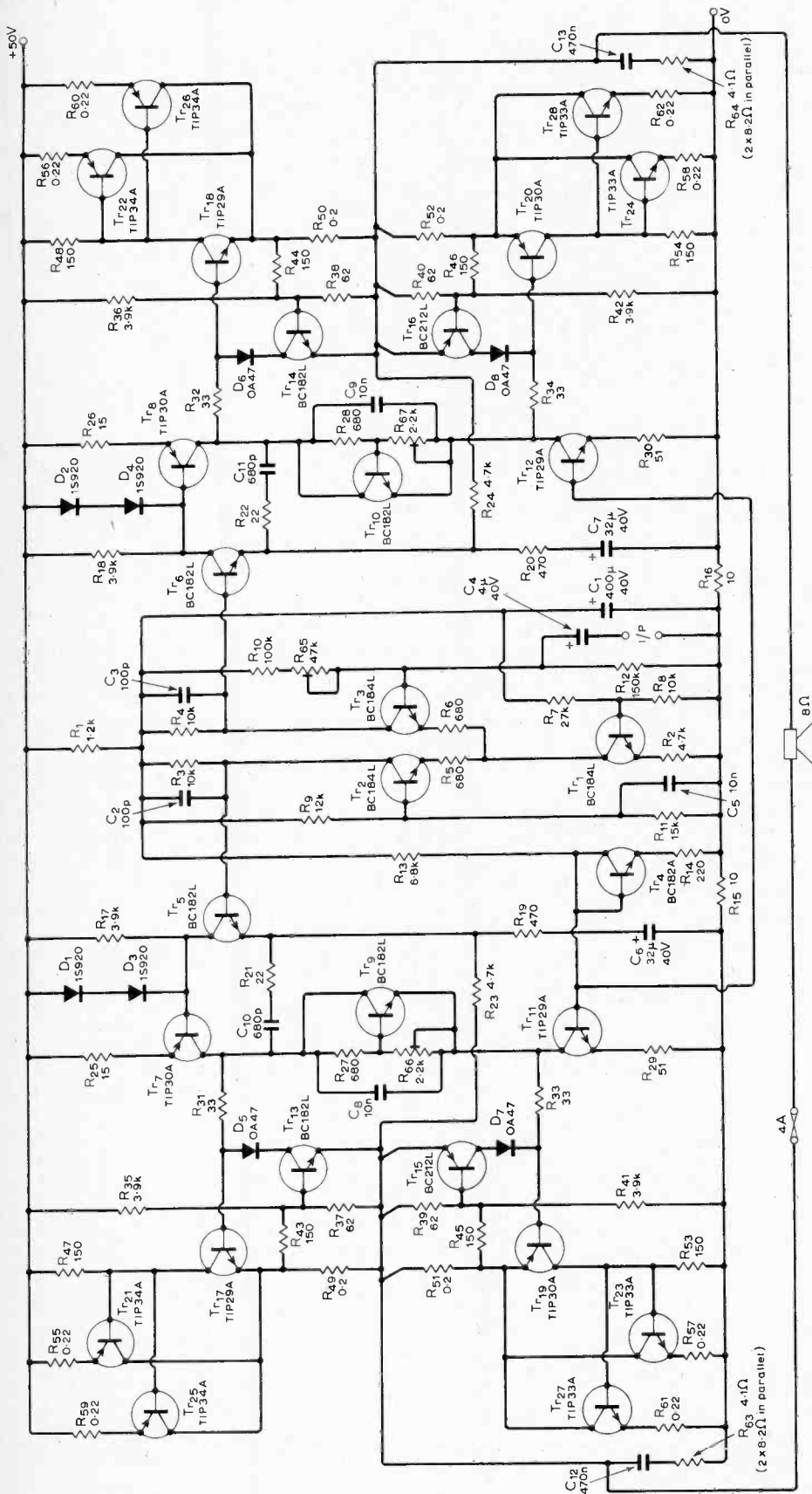


Fig. 9. Circuit of 100W amplifier with comprehensive overload protection.

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_4 , R_5 , 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)mA per °C temperature difference between the junctions of the driver transistor Tr_3 and the 'amplified diode', Tr_7 . 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 0.2Ω 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 at its base. If we allow 10mA to flow down R_2 and R_5 their values are 140Ω and 60Ω respectively. At point D there is 600mV across the parallel combination of R_2 , R_5 and the current down these comes via R_4 which has 53.4V across it hence the value of R_4 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_{21} and C_{11} , R_{22} 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 Zobel networks C_{12} , R_{63} and C_{13} , 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 $Tr_{9,10}$ are a bypass for high frequencies where the conductance of the transistor is reduced. Potentiometer R_{65} 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

for Tr_9 and Tr_{10} 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 Tr_9 or Tr_{10} to ensure constancy of current in the driver stages, with variations in the junction temperature of the current sink transistors. Also the $10k\Omega$ 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_{15} , R_{16} , 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

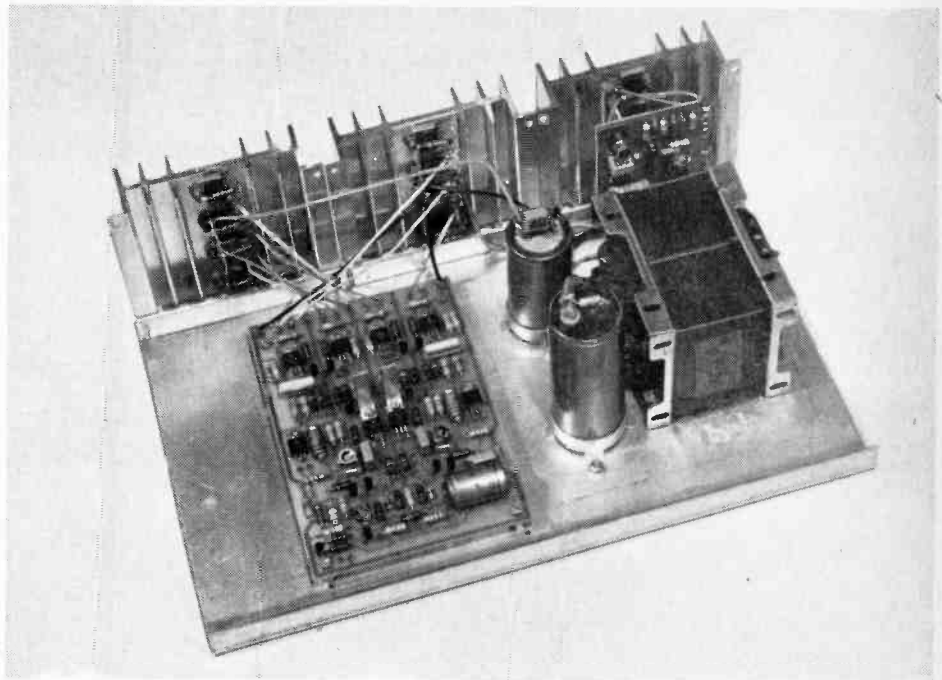


Fig. 11. Layout of prototype amplifier.

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 Tr_7 and Tr_8 and also on Tr_{11} and Tr_{12} 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 $Tr_{9,10}$ should be maintained at similar temperatures as can be achieved by glueing the flat of Tr_9 to the cap of Tr_{17} and likewise for $Tr_{10,18}$.

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 Th_1 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

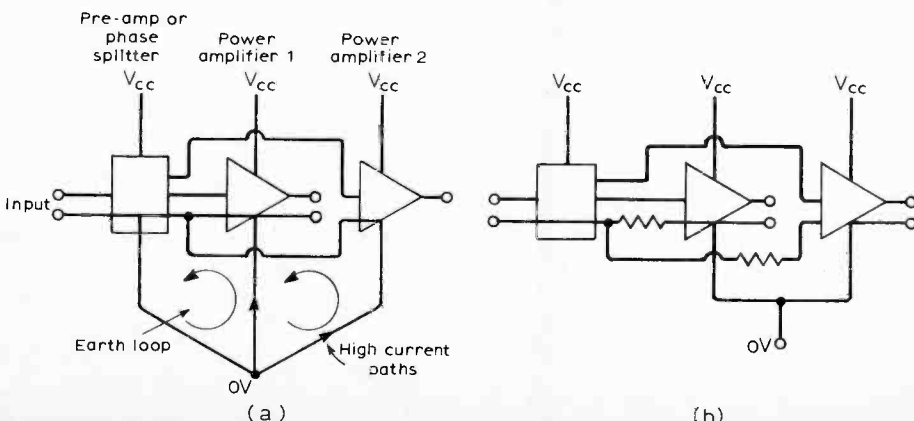


Fig. 10. Illustration of the common-line problem in amplifiers.

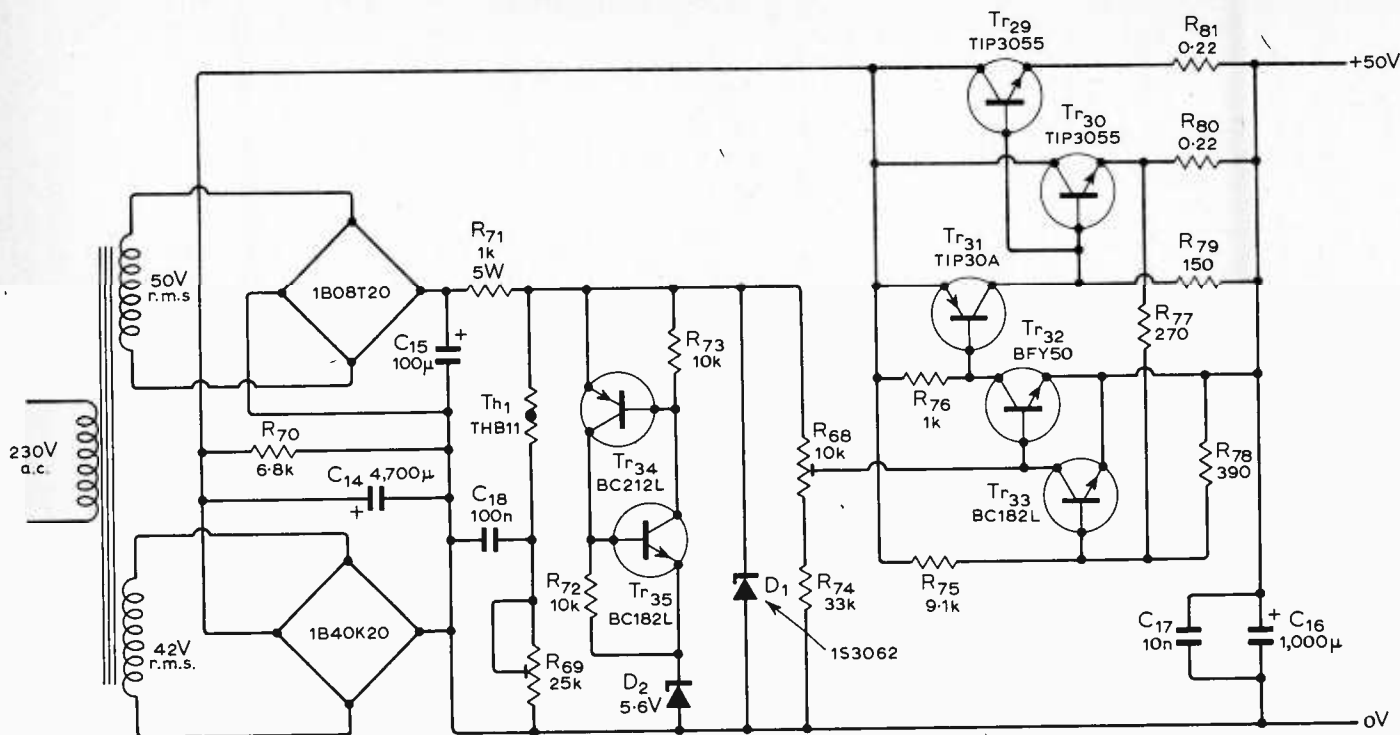


Fig. 12. Power supply incorporating short circuit protection and a thermal trip.

D_2 breaks down and allows enough current to flow through it to forward bias the base-emitter junction of transistor Tr_{35} . Transistors Tr_{34} and Tr_{35} 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_{67} before the output of the supply is connected to the amplifier. To set up the thermal cut-out temperature, R_{68} 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_{68} 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_{25} , R_{26} , R_{29} , R_{30} , and R_{47} through to R_{54} . To cater for the increased base current called for by the output devices it is advantageous to substitute TIP31A/32A for Tr_{17} through to Tr_{20} . Quiescent 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.

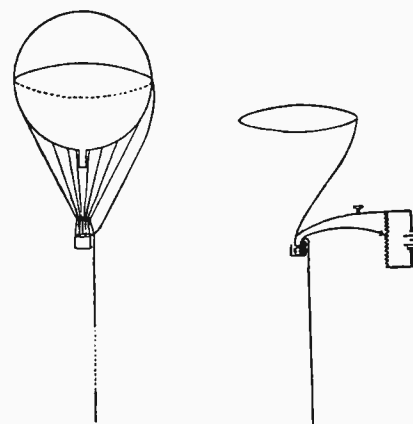
Acknowledgement. I express my appreciation of the assistance given by Ian Hardcastle and Basil Lane in the preparation of the article.

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1. Hardcastle and Lane, 'High Power Amplifier' *Wireless World*, October, 1970.
2. Bailey, 'Output Transistor Protection in A.F. Amplifiers' *Wireless World*, June, 1968.

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 Schloemilch 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.

Schloemilch 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.t.s 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.t.s 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.

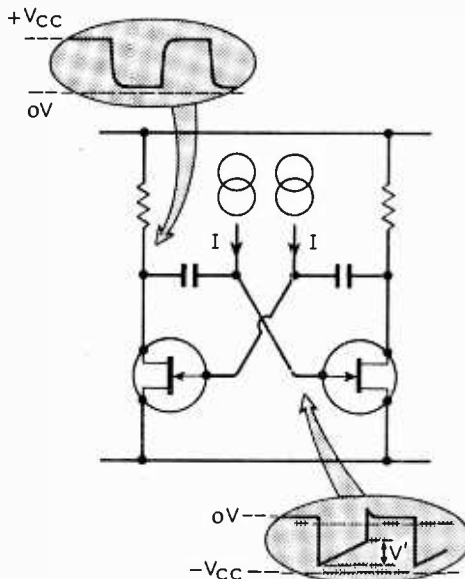


Fig. 1. Use of constant-current sources as shown allows linear discharge of capacitors.

Fig. 3. Relation of oscillation frequency with input voltage is linear over about seven decades.

Fig. 4. High-frequency limit can be increased by driving capacitors from a low-impedance source (response shown in Fig. 3).

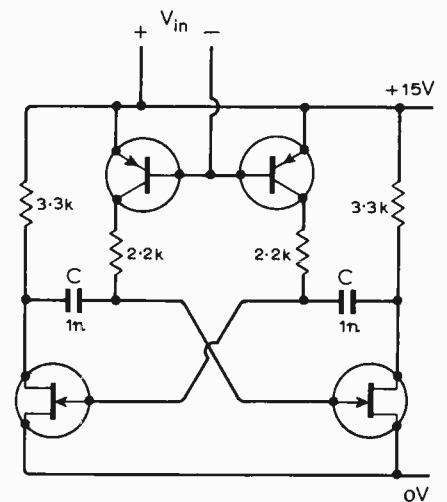
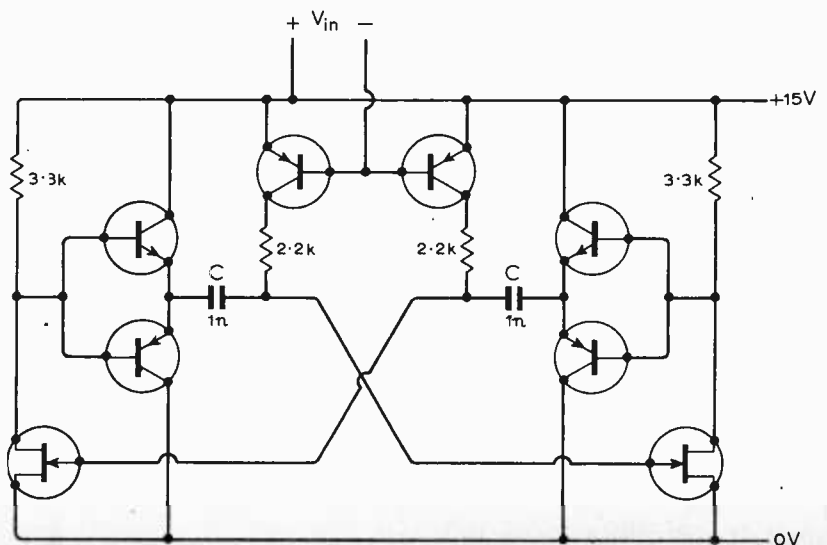
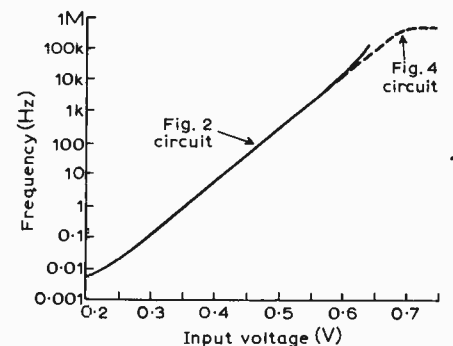


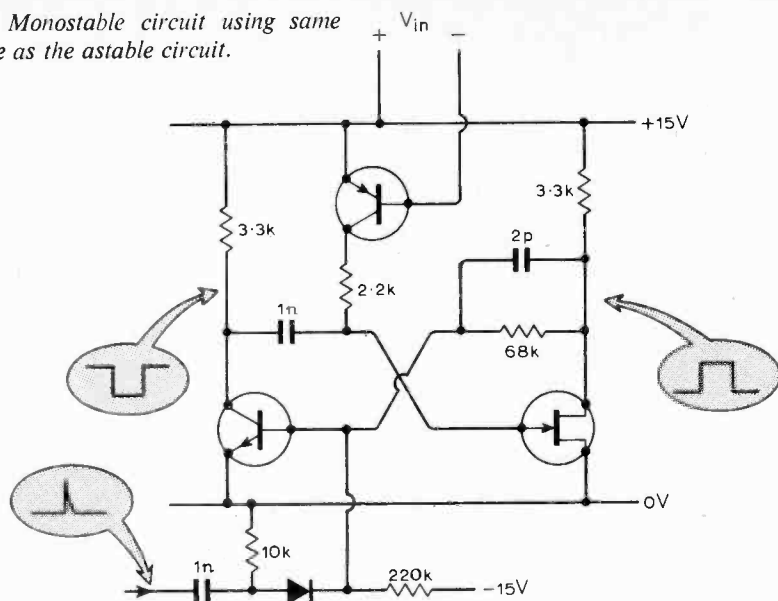
Fig. 2. Practical circuit based on Fig. 1 with bipolar transistors used as controlled-current sources.



*Clarendon Laboratory, Oxford.

†Bowes, R. C. 'Improved cross-coupled multivibrator controllable in frequency over a wide range' *Electronics Letters*, vol. 7, 1971 pp. 180-2.

Fig. 5. Monostable circuit using same technique as the astable circuit.



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 2.2k (s) safety resistors.

Monostable circuits

The monostable circuit follows from the astable circuit and uses the same technique to obtain a wide range of period.

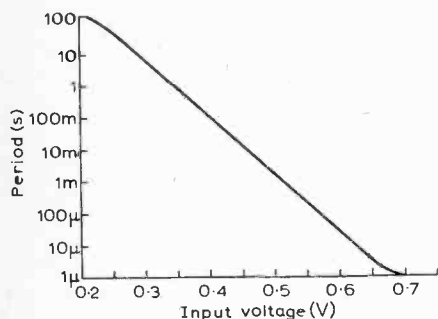


Fig. 6. Relation between monostable period and input voltage showing minimum range of period is seven decades.

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.

Books 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, convergence, chrominance signals (in detail), synchronous demodulators, decoding, chrominance amplifiers, colour difference and luminance amplifiers, and sub-carrier 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.

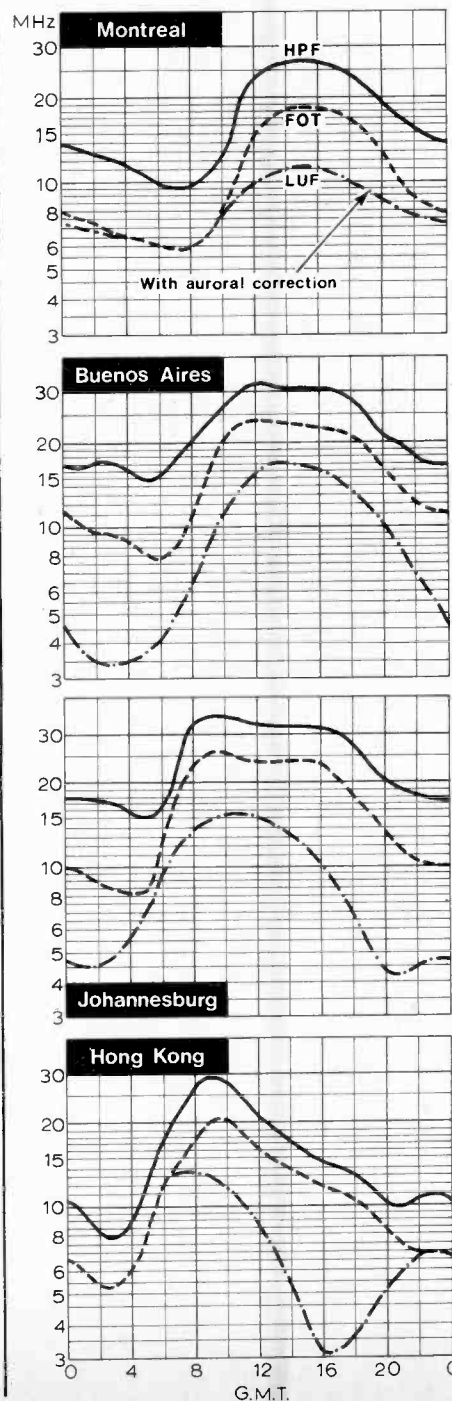
Signals and Information by C. C. Goodyear. The main emphasis is on signals and the ways in which they may be processed for transmission. The book is aimed at final year undergraduate level, specializing in telecommunications and other fields, such as control engineering where a knowledge of processing 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.

H.F. Predictions February 1972

As mentioned last month two of the parameters presented in these charts (produced by Cable & Wireless) have been changed. (1) MUF is replaced by HPF (highest probable frequency) which is the MUF (median usable frequency) exceeded on 10% of the days. (2) FOT (optimum traffic frequency) has the same definition as previously but is no longer a constant 85% of MUF.

HPF and FOT are derived from MUF by applying factors, having diurnal, seasonal, sunspot and geographic variations, which describe the distribution of daily values of standard MUF about their monthly median.

MUF can be approximated from the curves as being mid-distance (not mid-frequency) between HPF and FOT.



Educational Television Network

Broadcast system with talkback serving colleges and industry

by Aubrey Harris

Stanford University, in Palo Alto, California, operates a scheme of post-graduate 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, government, 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

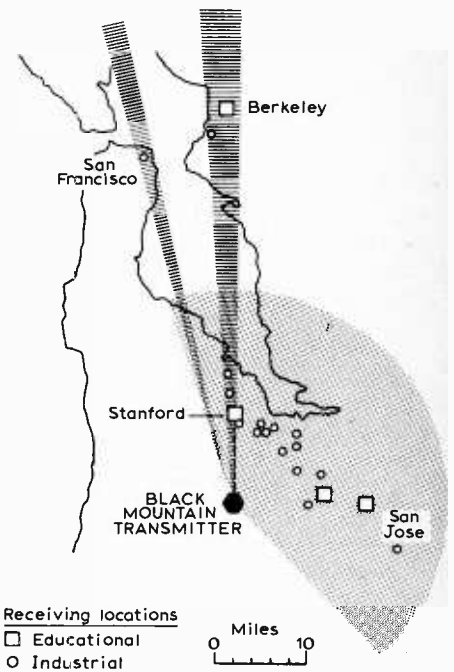


Fig. 2. Approximate coverage of directional and semi-omnidirectional aerials.

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).

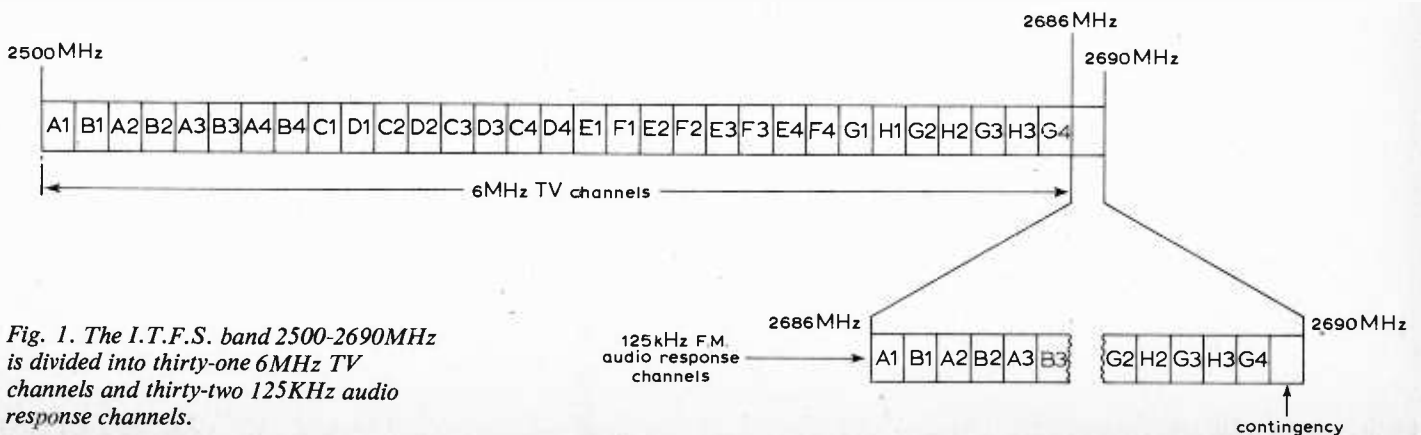


Fig. 1. The I.T.F.S. band 2500-2690MHz is divided into thirty-one 6MHz TV channels and thirty-two 125KHz audio response channels.

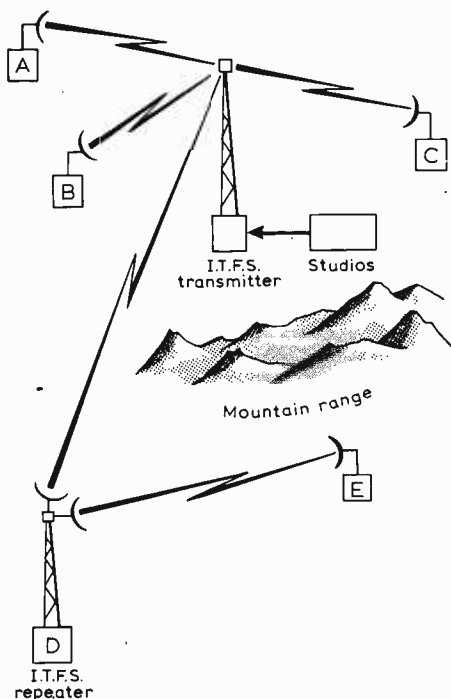


Fig. 3. I.T.F.S. transmitter may be used for direct transmission to schools and colleges (A, B, C) or via repeater (D) to location (E) not directly in view of transmitter due to natural obstruction.

These transmissions are classified as Instructional Television, Fixed Service (I.T.F.S.) and are available to educational institutions for transmission of instructional material to remote receiving classrooms. The area covered by the I.T.F.S. transmissions is shown in the map (Fig. 2). This indicates the receiving locations in each area and the siting 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 loca-

tions 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 broadcast, 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

* 525 lines, 60 fields/sec, 6 MHz channel width, 4.2 MHz vision bandwidth, negative modulation, sound carrier +4.5 MHz from vision carrier.

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 (£16,000-£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 installed above, another 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

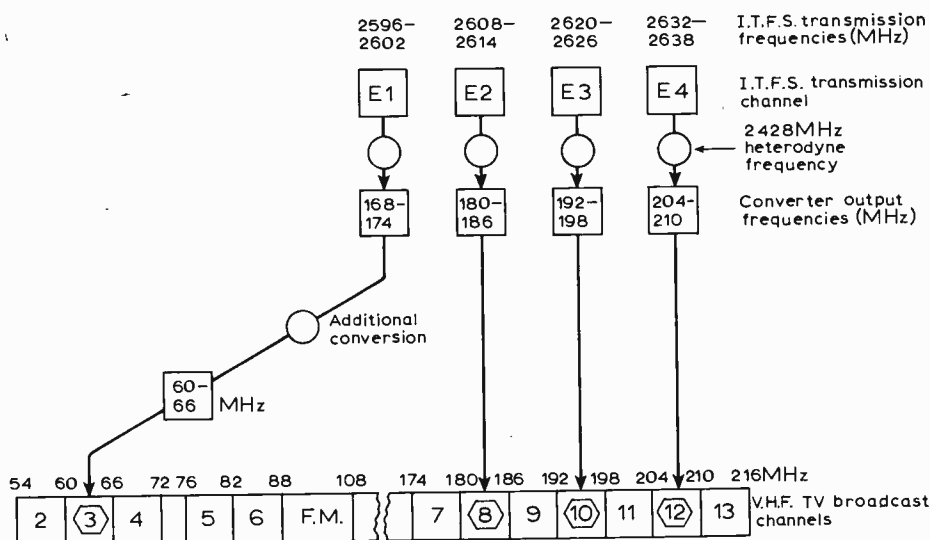


Fig. 4. 'Down conversion' of the I.T.F.S. channels brings the signals into the v.h.f. TV broadcast band with same spacing of bandwidth. E1 would drop out of normal TV channel; additional conversion places it in v.h.f. low-band.

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 paraboloid 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.

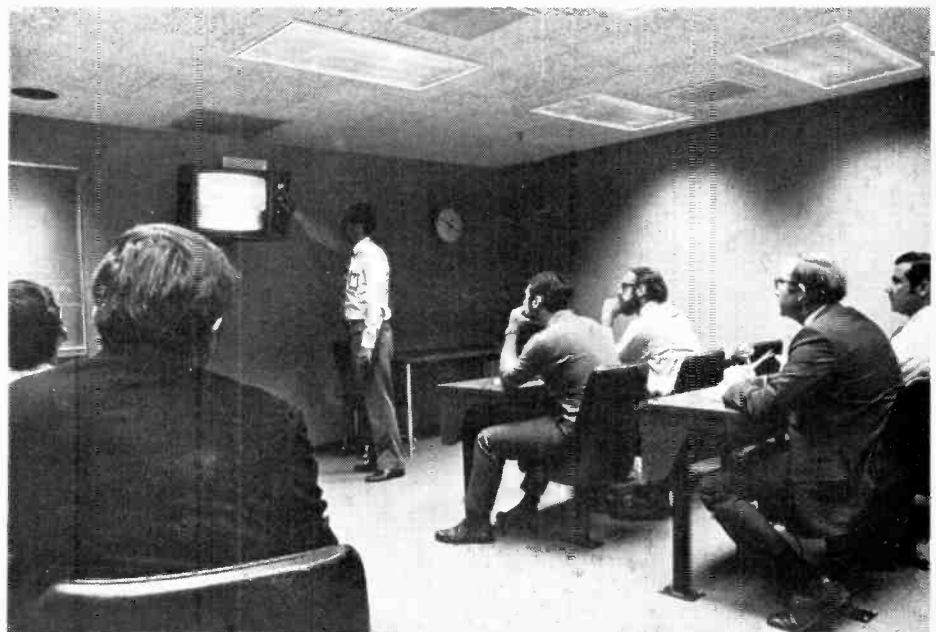


Fig. 5. A view from the front of the TV studio classroom. At the front can be seen the lecturer's monitor and microphone; similar units are positioned between each two students. The control room window is at the rear, to the left of which is the rear-mounted camera.



Fig. 6. A receiving location aerial tower. The lower dish is used for TV pick-up and is connected to the down-converter bolted to the tower. The upper dish is used for the audio talk-back transmission.

Fig. 7. A typical TV classroom at a remote receiving location.



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 $1k\Omega$ 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 to use 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 uniselector 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 uniselector.

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 uniselector returns to zero and the exchange switches itself off.

The switch contacts on the uniselector, 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 also a contact breaker 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 uniselector 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 uniselector 1. The uniselector 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

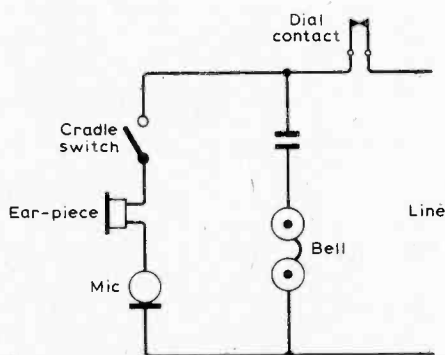


Fig. 1. Simplified telephone circuit.

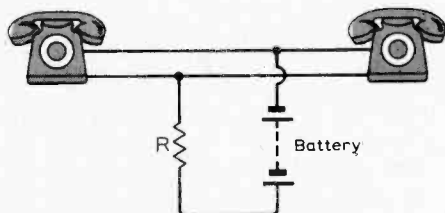


Fig. 2. Method of connecting two telephones. R should be selected to suit the battery employed as follows: for 50V use $1k\Omega$, for 67.5V use $1.5k\Omega$ and for 90V use $1.8k\Omega$.

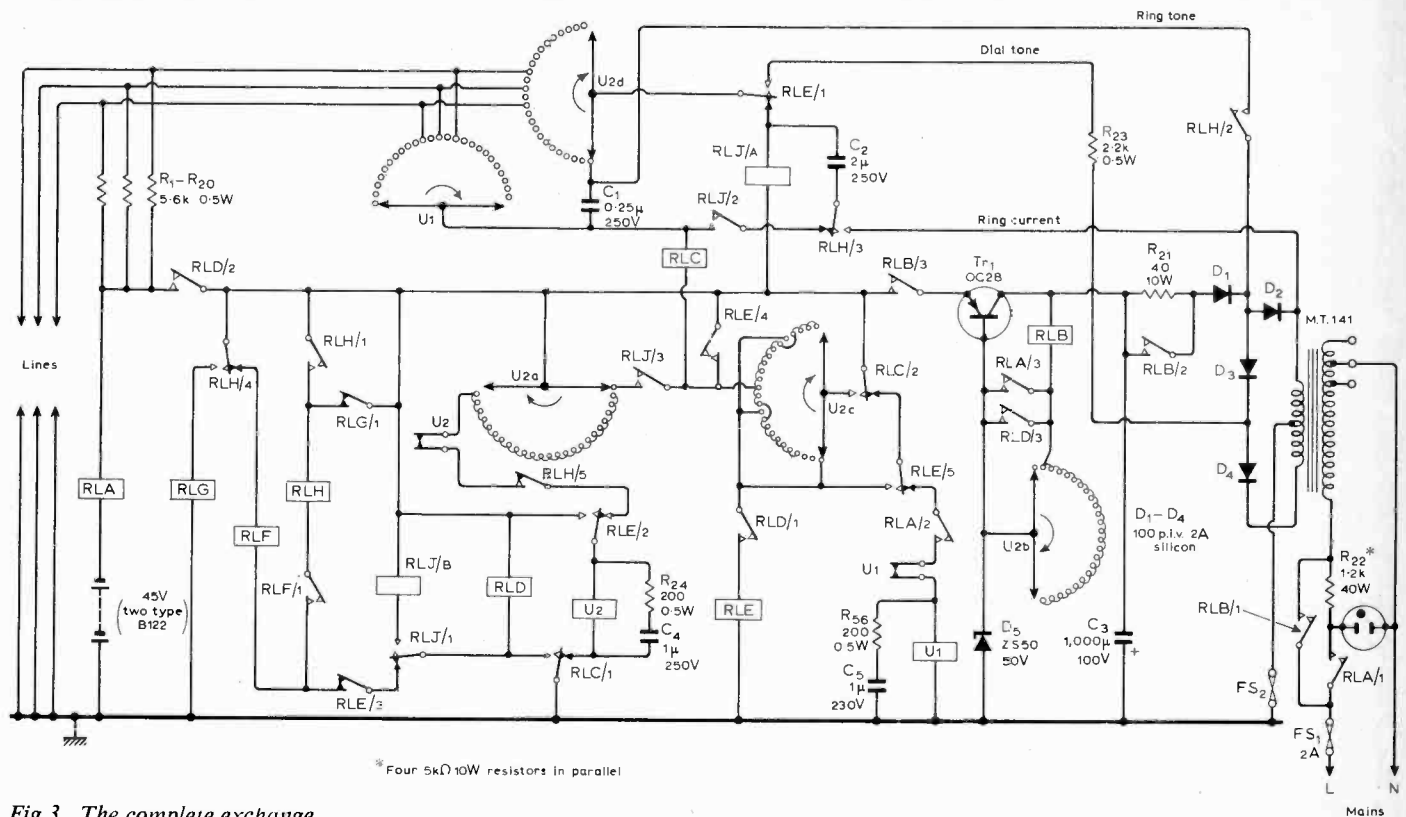


Fig.3. The complete exchange.

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 get its supply via bank c of U2, 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 U2c before RLE has a chance to turn off.

When RLE de-energizes RLE/3 energizes RLH via RLG/1, RLF/1 and RLC/1. RLH/3 switches the ringing current to the telephone being called and RLH/2 feeds the ringing tone to the calling telephone. RLF and RLG are both very slow release relays, and together with RLH 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 RLJ using winding A, and it holds itself on by energizing winding B via RLJ/1. Its action is to stop the ringing signals and connect

the called telephone to the calling telephone with RLJ/2. The purpose of RLJ 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 speech 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 R22. The rectified output is smoothed by C3 via R21, and stabilized by D3 and Tr1. The coil of RLB forms the bias resistor of Tr1, and the relay operates only when C3 has charged up to 50V, i.e. when the bias current starts rising. When RLB/1 closes R21, and R22 are short-circuited (RLE/1 and RLB/2) and RLB/3 connects the power to the rest of the circuit.

RLD/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 RLD and RLJ. A supply passes through U2a and the coil of U2, and this returns the wiper of U2 to its homing position by the action of the contact breaker. Bank b of U2 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
Post Office type 3000 relays

No.	Coil resistance	Contacts			Slug length in inches
		make	break	change over	
RLA	5000Ω	3	—	—	—
RLB	800Ω	3	—	—	1 or 1.5
RLD	400Ω-5kΩ	3	—	—	1
RLE	400Ω-5kΩ	1	1	3	1
RLF	400Ω-5kΩ	1	—	—	1.5
RLG	400Ω-5kΩ	—	1	—	1.5
RLH	400Ω-5kΩ	2	1	2	—
RLJ/A	3000Ω	—	—	—	—
RLJ/B	400Ω-5kΩ	1	1	1	1

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_3 charges, RLB will also operate. U_1 should rotate, then stop when RLC energizes, RLD and RLE 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 RLE should break and RLF, RLG and RLH should turn on. 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.

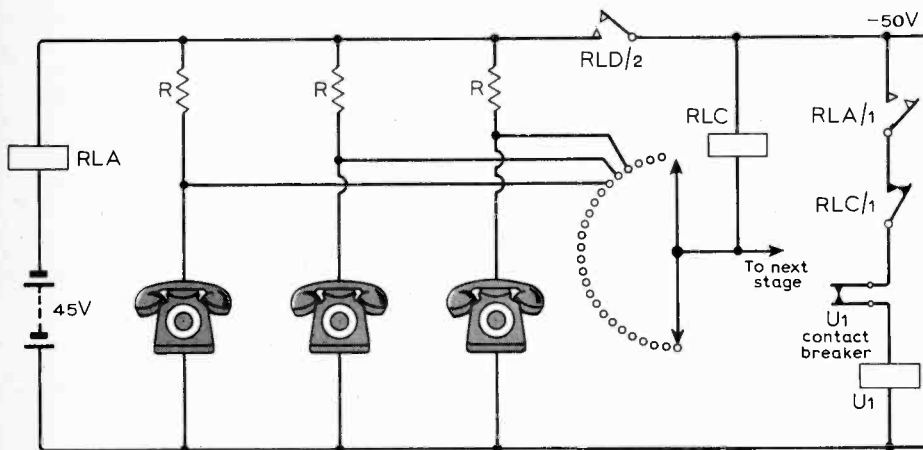


Fig.4. Simplified circuit of part of exchange.

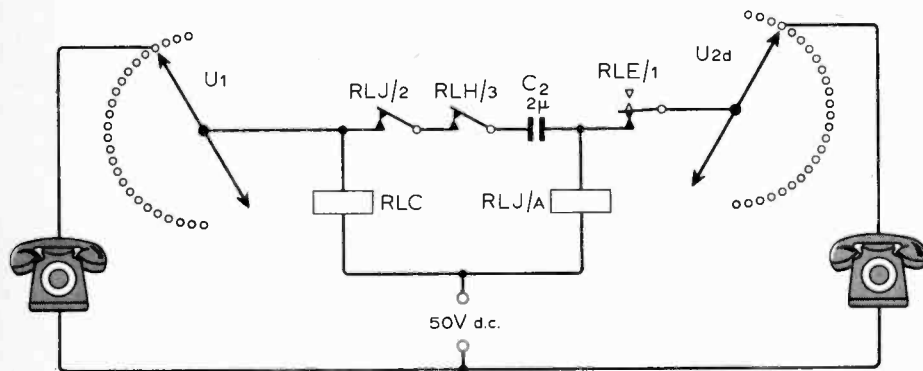


Fig.5. The final speech path through the exchange.

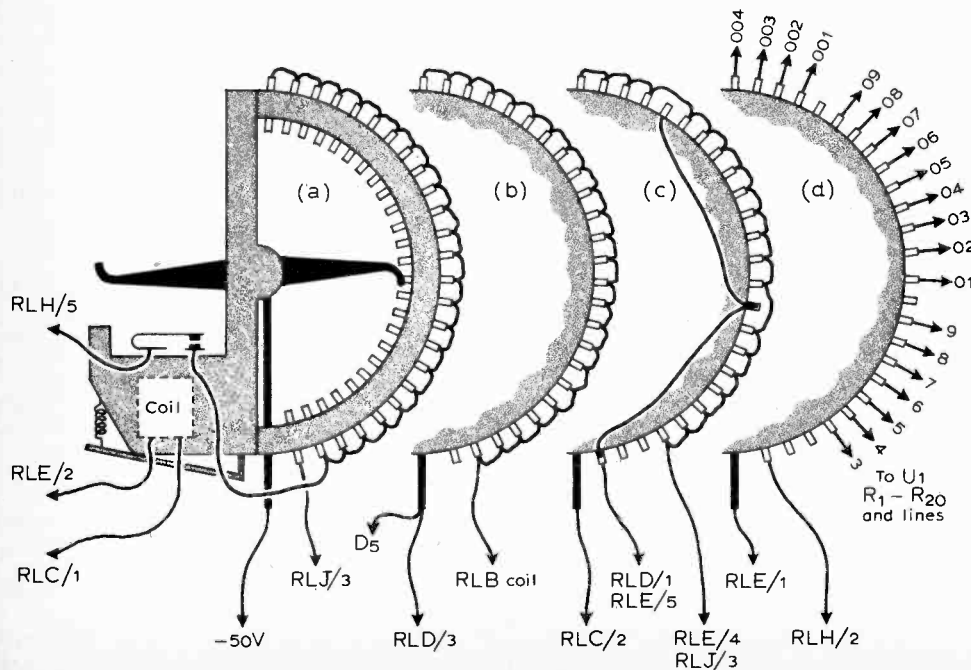


Fig.6. Connecting uniselector U_2

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. Haines**, formerly of the Plessey Company, who has resigned from the industry. Dr. O'Kane, who is 59 and a graduate of Liverpool University, joined in 1935 the Research Laboratories of the General Electric Company where he served until 1941 when he was transferred on loan to the Telecommunications Research Establishment at Malvern. He rejoined the G.E.C. Radio Department in 1945. During 1947 Dr. O'Kane moved to International Aeradio 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 fluid dynamics; 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 1943. After 18 months with the Admiralty Signal Establishment at Witley, Surrey, he transferred to Bath to supervise the work of the Admiralty Computing Service. The duties and staff of this group were transferred to the National Physical Laboratory in 1945, and in 1951 Dr. Goodwin became superintendent of the Division of Numerical Analysis and Computing.

Captain Roger L. Clode, 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 latter 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 commanded the Naval Air Electrical School at Lee-on-Solent. Capt. Clode subsequently held the posts of Director of Naval Electrical Training and then the Director of Radio Equipment at the Admiralty.

I. G. Phillipps, 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. Phillipps' 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 City & 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.E.-I.E.R.E. Proceedings, India*, and **K. B. Kapasi**, M.Sc., has joined the I.E.R.E.'s 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 in the U.K. and on the Continent. 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 electroacoustics where he obtained his Dipl.-Ingénieur. 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. A physics graduate of London University, Mr. Singer joined N.T. 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.

S. E. Goodall, M.Sc., F.I.E.E., president, International Electrotechnical Commission.

O.B.E.

A. E. W. Austen, Ph.D., B.Sc., F.I.E.E., technical director, C.A.V. Ltd.

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.R.E., director of research, Plessey Telecommunications Co. Ltd.

R. M. Trim, F.I.E.R.E., F.I.E.E., technical director, Cossor Electronics Ltd.

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

M.B.E.

A. E. M. Wase, M.I.E.E., senior executive engineer, International & Maritime Telecommunications Region, Post Office.

C. H. Wolff, B.Sc., M.I.E.E., head of data handling division, Plessey 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 U.S.A. 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 SHAEF 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 1957 as chief development engineer — switches.

World of Amateur Radio

More light on transequatorial propagation

Recent observations by the Australian Ionospheric Prediction Service (see *Nature Physical Science*, November 22, 1971) on paths between Japan/Okinawa and Queensland, Australia serve strongly to support the view, frequently expressed in the past by amateurs, that transequatorial 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 transequatorial openings, but it can be observed occasionally, particularly on 28MHz, in the second form.

Among the useful beacons for t.e.p. work are GB3SX at Crowborough, Sussex, now operating regularly on 28.185MHz and on 70.669MHz (on the latter frequency the aerial points north to detect auroral openings); ZS6VHF near Johannesburg on 50.1MHz from Friday evenings to Monday mornings; and, from about February, a new beacon ZC4VHF on Cyprus which plans to radiate 20 watts on 50.5MHz 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? Could not local clubs and societies 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 1 News*. The total number of people now licensed as amateur operators has reached 256,655 although only about 54% of these actually hold call signs 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 14MHz) 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.L. without formal examinations. Two bands are severely restricted in Japan: 160-metres comprises only 1907.5 to 1912.5kHz and 80-metres is limited to 3500 to 3575kHz.

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 switch in an aerial attenuator at the receiver and note whether the unwanted products persist or disappear. But modulation problems can give rise to controversy, as has happened as a result of the 1971 V.H.F. National Field Day when the 144MHz entry of the Radio Society of Harrow was disqualified and so lost Harrow the chance of winning the contest. Local amateurs admit the disqualification followed complaints by other stations of excessive bandwidth, but claim that without objective measurements it is virtually impossible to say with certainty whether the fault was in the receivers or the transmitter.

In brief

The Bangor & District Amateur Radio Society has recently re-activated the GB3GI 145.990MHz beacon from a new location near Saintfield in County Down (XO 4IJ) 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, GI2FHN, 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.

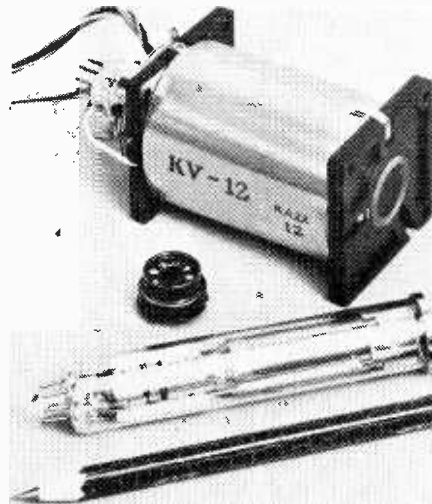
PAT HAWKER, G3VA

New Products

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 ± 1 dB (model NS110S) and ± 3 dB (model NS110G) over the range 60-300Hz, ± 5 dB and ± 10 dB respectively over 20Hz-3MHz. Asymmetry is better than 15%. The output amplifier presents an output impedance of 600 Ω . The module measures approximately 2 x 1 x 0.62in deep (51 x 29 x 16mm). 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

Miniature vidicon tube

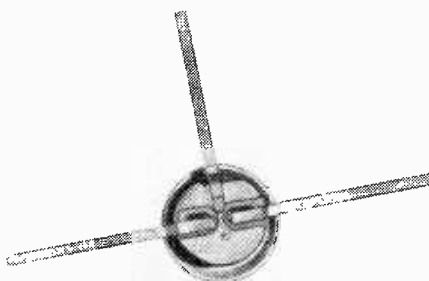
A miniature vidicon television camera tube type 20PE13, announced by Mullard, has a resolution of 550 lines. Electromagnetic focusing and scanning are used. When subject to an illumination of 10 lux from a lamp with a colour temperature of 2854°K the tube provides a signal current of 150nA. Its dark current is 20nA. External measure-

ments are 17.70 x 108mm. The tube can be supplied complete with yoke assembly and socket for £26.25, or the individual components can be ordered separately. Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

WW331 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 insertion 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, as 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.

WW 337 for further details (transistor)

WW 338 for further details (Gunn oscillator)

WW 339 for further details (Gunn diode)

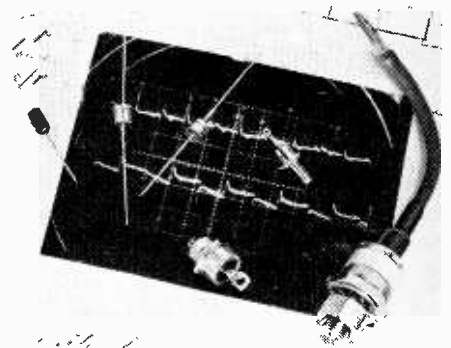
5W regulator diode

An axial-lead 5W sub-miniature plastic-encapsulated voltage regulator diode has been introduced by I.R. The new series, designated 5ZS (1N5333A to 1N5378A JEDEC code), is available in selected voltage grades between 3.3 and 100V in both $\pm 5\%$ or $\pm 10\%$ tolerances. Operation is possible in the range -65 to +200°C. Size: 9mm long, 4mm diameter. International Rectifier, Hurst Green, Oxted, Surrey.

WW 335 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 1 μ s. The extreme stand-off voltage values for each range are shown in the table below; intermediate values correspond to the E24 series.

Type	Stand-off voltage range	Dissipation rating (for 1ms)
BZW70	5.6 to 62V	400W
BZW86	7.5 to 62V	12kW
BZW91	5.6 to 62V	5kW
BZW93	5.6 to 62V	450W
BZW95	7.5 to 62V	450W
BZW96	3.9 to 7.5V	90W

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

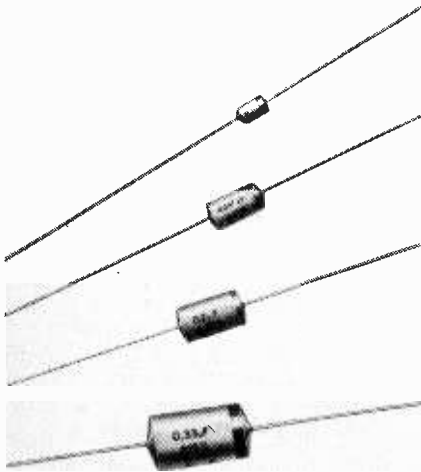
sion with transient suppression diodes' (TP1280), it outlines the principles of surge suppression and gives information on the design of practical circuits. Instrumentation and Control Electronics Division, Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

WW333 for further details (devices)

WW334 for further details (reprint)

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.
WW 336 for further details

Instant edge connectors

A range of Circuit-Stik pressure sensitive connector sub-elements is available from Bourns (Trimpot). They are drilled to 0.100in, 0.125in or 0.156in pin centre spacing and nickel gold plated ready for

soldering and for use with 15- and 22-pin edge connectors. Circuit side-matching connector strips with staggered feed-through holes are designed for dual row connectors allowing 30 or 44 connections in a compact area. The strip is trimmed to the shape of the p.c. board (with a sharp knife), the adhesive backing removed, holes aligned and pressed by hand into place. Connections can be immediately soldered to the strip and a connector fitted. Bourns (Trimpot) Ltd, Hodford House, 17/27 High Street, Hounslow, Middx.
WW305 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

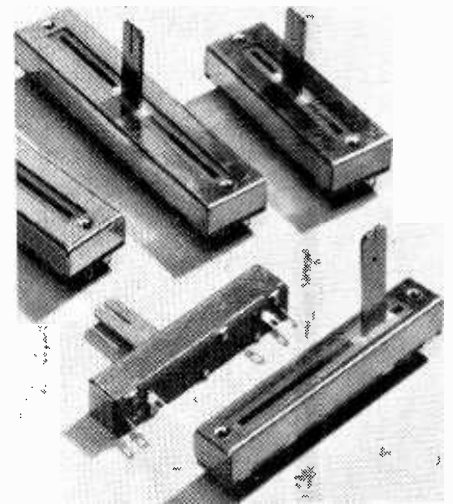
The Tektronix type 211 solid state miniature oscilloscope has double insulation, and is designed for use in severe environments. It will operate from a.c. mains or for up to 5 hours on internal batteries. Bandwidth is 500kHz, and deflection factors are 1mV/div to 50V/div. Sweep rates extend to 1μs/div. 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 × 133 × 229in. 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

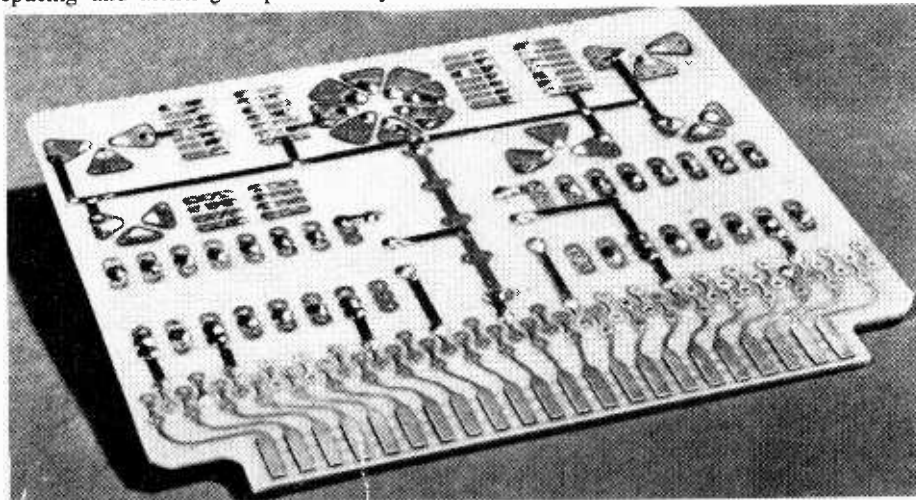
Fast recovery silicon diodes

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

Microwave 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,



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.p.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, $Z_o = 1k\Omega$
- 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

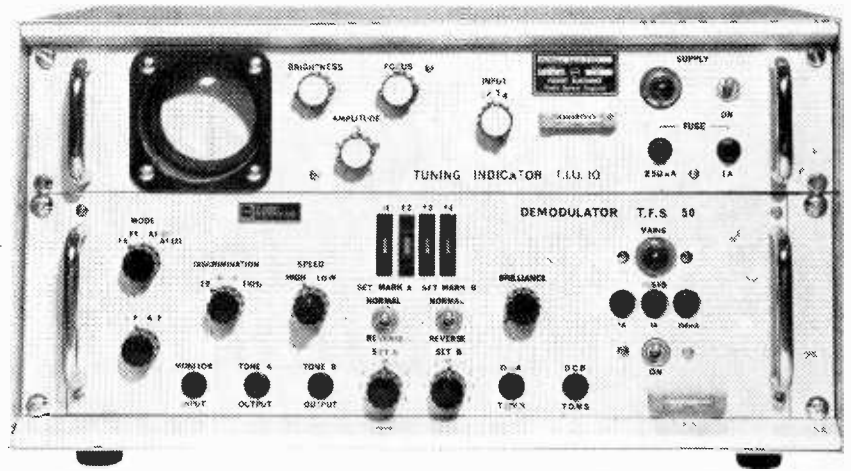
A solid-state variable-power supply controls the temperature of the bit of the Litesold ETC/1 lightweight soldering iron. The operating temperature can be set within the range 150-400°C. The plug-in soldering



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 d.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 frequency 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

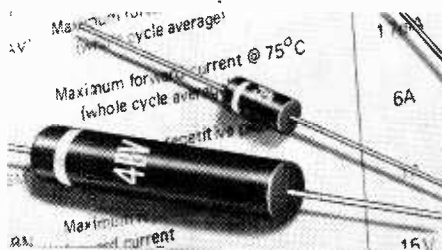
Dual in-line package sockets

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

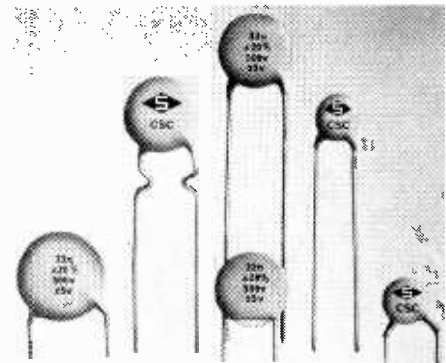


13kV (V_{RRM}) in a larger package size, 40 × 8mm. This series is designated 4BV and is rated at 40mA at 25°C. Both sizes of cartridges are capable of operating at up to 20kHz into a resistive inductive load with an operating temperature range of -40 to +75°C. International Rectifier, Hurst Green, Oxted, Surrey.

WW309 for further details

Ceramic disc capacitors

A range of ceramic disc capacitors, type CSC, from Seatronics (UK) range between 1000pF and 50000pF (E6 series) with ±20% 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

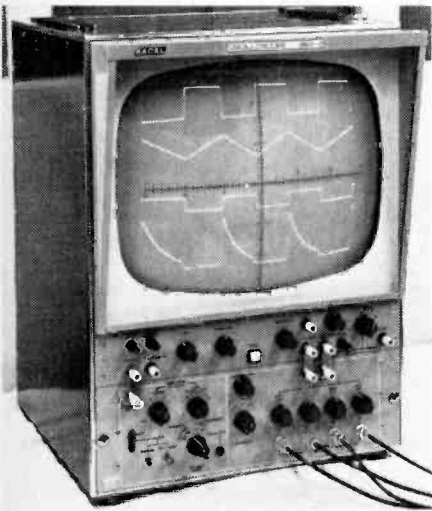
Small power operational amplifier

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 20M Ω 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. WW329 for further details

Large-screen oscilloscope

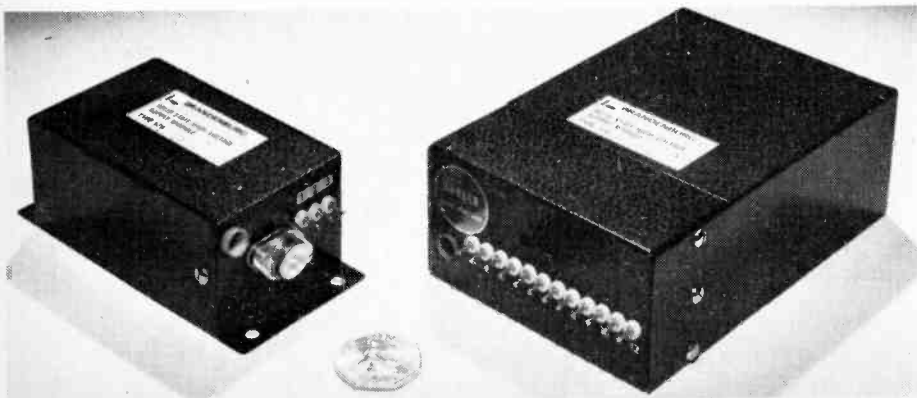
An automatic brightness control, which enables the timebase range to be changed from 1ms to 30s without adjustment to the brightness control, is provided on the large screen display oscilloscope type 383-2 from Racal Instruments. Up to four orange traces can be displayed



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. WW324 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 500 μ A. 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.

WW318 for further details

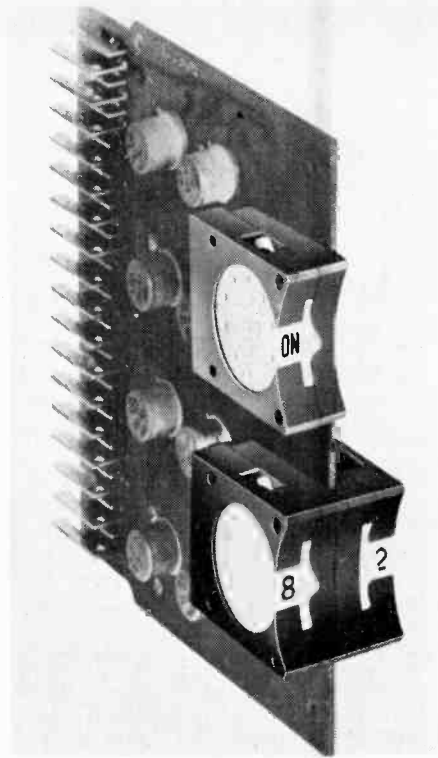
Logic power-driver i.c.

Motorola Semiconductors have introduced type MCH 2890 integrated circuit dual logic output power driver. Its function is to boost the outputs of d.t.l. and t.t.l. 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.l. AND gate and a pair of Darlington power transistors. With normal d.t.l. and t.t.l. 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 6 $^{\circ}$ C/W. Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middx.

WW316 for further details

Thumbwheel switch

The EEC0 8000 series thumbwheel switch from Plessey is the result of a licensing arrangement with EEC0 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 $\frac{1}{2}$ in 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.

WW325 for further details (switch)

WW326 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 \pounds 215. Marconi Instruments Ltd, St. Albans, Herts.

WW315 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

the design of the TA940 linear amplifier. The MA1720 exciter unit used in the transmitter gives coverage from 1.6 to 30MHz in 100Hz 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 transmitter terminal, including amplifier, exciter and feeder matching unit, is under 153cm. Rascal Communications Ltd, Western Road, Bracknell, Berks. RG12 1RG.

WW314 for further details

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.

WW317 for further details

Plug-in 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 ± 25 V. Each card is fully protected against excess current and all have provision for the Roband EPO over-voltage protection module. Except for the 10A and twin units, the size of the Rokards (105 x 121 x 41mm) is no greater than existing cards having much lower power capabilities. The p.c.b. track is gold plated. Prices range from £8 for 1A to £15 for 10A. Twin units are £16.25 (25W) and £21 (60W). Roband Electronics Ltd, Charlwood Works, Charlwood, Horley, Surrey.

WW311 for further details

High-stability voltage reference units

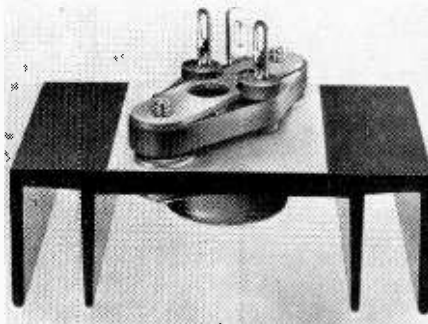
Semitron have introduced a range of units designed to provide a very stable voltage reference from unregulated inputs over extended periods of time and a wide range of temperatures. The reference units have their own built-in power supply which is temperature compensated. The result is stability of $\pm 0.005\%$ and temperature coefficients as low as $\pm 0.0005\%$ 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 x 25.5 x 13mm. Semitron Ltd, Cricklade, Wilts.

WW312 for further details

Transistor/thyristor insulating caps and sockets

The MS 501 and MS 502 series of insulating caps from Waycom have been designed 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 temperature of 120°C. Also introduced is the MS 510 socket which facilitates ease of insertion and replacement of TO-3 transistors or thyristors. Waycom Semiconductors Ltd, Wokingham Road, Bracknell, Berks.

WW327 for further details

Precision voltage reference diodes

Stable voltage reference diodes from Mullard 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 coefficients ranging from ± 0.01 to $\pm 0.005\%$ per °C; a typical value for the temperature coefficient 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.

WW320 for further details

Fourteen-track tape recorder

An f.m.-analogue 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 four switch selected tape speeds — $\frac{1}{16}$, $3\frac{3}{4}$, 15 and 30in/s. The recorder is portable and may be battery operated. It uses reusable cartridges containing standard 8-inch reels of $\frac{1}{2}$ -inch tape, and is also pro-

vided with a voice channel as standard. Pye TVT Ltd, Coldhams Lane, Cambridge. WW313 for further details

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 opposite in polarity. The GSV series is rated for 125A peak-pulse current for 1ms. Breakdown voltage is typically 0.35V to 0.50V at 10 μ A and 0.74V to 0.85V at 100mA. Units are contained in an axial lead package 0.200 x 0.370in. Auriema Ltd, 442 Bath Road, Slough SL1 6BB. WW328 for further details

S.S.B. radiotelephone

The Trident type T20 single-sideband radiotelephone is solid-state and has a frequency range of 1.6 to 4.2MHz. It provides 20W of peak envelope power. The six receive and transmit channels are independently selected and crystal controlled, and a two-tone alarm generator can be supplied as an optional extra for use on the marine distress frequency (2.182 MHz). Operation is from a d.c. supply of 12 or 24V. Trident Electronics Ltd, 3 Towerfield Close, Towerfield Estate, Shoeburyness, Essex.

WW330 for further details

Polyester capacitors

A range of capacitors with polyester film dielectric and aluminium foil electrodes is available from Jermyn. The capacitors have a low dissipation factor, high insulation resistance, and low self inductance. Capacitance and voltage ranges are:

0.01 μ F to 0.47 μ F	160V d.c.
1000pF to 0.47 μ F	400V d.c.
1000pF to 0.1 μ F	1000V d.c.

Standard capacitance tolerance is $\pm 10\%$ but $\pm 5\%$ or $\pm 20\%$ are available to order. Jermyn Industries, Vestry Estate, Sevenoaks, Kent.

WW322 for further details

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 harmonically 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.

WW319 for further details

February Meetings

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

LONDON

1st. Soc. Relay Eng. — "British Relay multi-channel colour television relay system" by K. A. Russell, A. Burke and D. Bashford at 14.30 at the I.E.E., Savoy Pl., W.C.2.

1st. IEE — "Ceramic dielectrics for microwave devices" by R. C. Kell 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. I.Mech.E./IEE — Discussion on "The design of gyros and accelerometers for inertial navigation" at 18.00 at 1 Birdcage Walk, Westminster, S.W.1.

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.

7th. IEE/IERE — Computer architecture colloquium at 14.30 at Savoy Pl., W.C.2.

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. IERE — "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.

14th. IEETE — "The training of technician engineers and technicians" by R. E. Stevenson at 18.00 at the IEE, Savoy Pl., W.C.2.

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

16th. IERE — "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 speech synthesis" by A. J. Fourcin at 18.30 at the Engineering Lecture Theatre, University College.

23rd. R.Ae. S./IEE — 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 auger spectroscopy symposium at 11.00 at Imperial College.

23rd. IEE/I. Meas. Control — "Optimal input signals for non-linear system identification" by G. C. Goodwin and "The design of digital controllers for randomly disturbed systems" by D. W. Clarke at 17.30 at Savoy Pl., W.C.2.

23rd. IERE — "Solid-state transmitters" by B. D. Bradshaw at 18.00 at Engineering Lecture Theatre, University College London, Torrington Pl., Gower St, W.C.1.

24th. I.Phys./B.Acoustical Soc. — "Concert hall acoustics with special reference to the Royal Festival Hall" by J. Bradley, P. H. Parkin and B. Tunbridge at 15.15 at the Royal Festival Hall.

ABERDEEN

23rd. IERE — "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. IERE/IEE — "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. IEETE — "Principles and applications of lasers" by Dr. L. Allen at 19.30 at Royal Albion Hotel.

29th. IERE — "The electronic teleprinter" by J. V. Evans at 18.30 at the Technical College.

BRISTOL

16th. IERE — "Digital techniques applied to the measurement of low-frequency noise" by Prof. H. Sutcliffe at 19.00 at the University, School of Chemistry, Lecture Theatre No.4.

CARDIFF

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

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

CHATHAM

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

CHELMSFORD

24th. IERE — "Radio astronomy telescopes" by R. Levente at 19.30 at The Hoffman Lecture Theatre.

CONNAH'S QUAY, Flint.

3rd. BCS — "The Common Market — a challenge to the computer industry" by C. J. C. McOustra at 19.30 at Half-Way House.

EDINBURGH

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

23rd. BCS — "Machine intelligence" by Prof. D. Michie at 18.00 at Mountbatten Building, Heriot-Watt University.

EVESHAM

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

EXETER

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

FARNBOROUGH, Hants

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

GLASGOW

7th. IEE/IERE — "Electronics in crime detection" by A. T. Torlesse at 18.00 at The Institution of Engineers and Shipbuilders, Rankine House, 183 Bath St.

GUILDFORD

16th. IEE — "Telecommunications in the supply industry" by J. W. Dillow at 19.30 at the C.E.G.B., Burymead House.

LIVERPOOL

9th. IERE — "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.

LOUGHBOROUGH

15th. IERE/IEE — "High-fidelity reproduction" by R. L. West at 18.45 at the University.

MANCHESTER

17th. IERE — "Automatic landing — the instrument landing system" by J. W. B. McIlwraith 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 Charles Trevelyan Technical College, Maple Terrace.

9th. IERE — "Sensitive 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.As" by L. Briggs at 19.30 at Norwich City College, Ipswich Rd.

16th. IERE/IEE — "A sound in sync system for television distribution networks" by D. Allanson at 19.00 at The Assembly House.

PLYMOUTH

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

READING

16th. IEE/IERE — "Thyristors and their applications in control systems" by P. Atkinson at 19.30 at the J. J. Thomson Laboratory, University of Reading, Whiteknights Park.

STAFFORD

28th. IERE/IEE/IPOEE — "The applications of electronics in large electrical machines" by Dr. R. D. M. Whitelaw at 19.00 at North Staffordshire College of Technology, Beauside.

SWINDON

8th. IERE/IEE — "The computer in management: master, parasite or servant?" by D. Firnberg at 18.15 at The College.

WOLVERHAMPTON

22nd. IERE — "Teaching electronics at sixth form level" by Mr. Foxcroft at 19.15 at the Technical Teachers' College.

Literature Received

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

ACTIVE DEVICES

A data sheet from M.C.P. Electronics Ltd, Alperton, Wembley, Middx, HAO 4PE, gives information on the PT5741 r.f. power transistor from TRW Semiconductors. The PT5741 is an epitaxial 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 controlled rectifiers — 40RCS. International Rectifier, Hurst Green, Oxted, Surrey WW402

A 32in x 23in wallchart detailing Motorola's plastic encapsulated small-signal transistors is available from GDS (Sales) Ltd, Michaelmas House, Salt Hill, Bath Rd, Slough, Bucks. WW403

Data sheets on five linear circuits are available from National Semiconductor GmbH, Lechstrasse 255, 891 Landsberg, West Germany.

LM112 (micropower operational amplifier) WW404

LM113 (temperature compensated low-voltage reference diode) WW405

LM114 and LM115 (transistor pairs with high 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 Celdis, 37/39 Loverock Rd, Reading, Berks, RG3 1ED, lists all types of components (active and passive) available from Celdis 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

Rastra Electronics Ltd, 275 King St, Hammersmith, London W.6, 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., Newsbureau, 17 Rue des Pierres-du-Niton, 1207 Geneva, Switzerland. The d.i.l. package features seven thick film resistors of equal value available from the Helipot 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 ± 3 ns over greater than 100MHz bandwidth) is described on a sheet from Wavecom Inc, 9036 Winnetka Avenue, Northridge, California 91324 WW413

APPLICATIONS

Data Recognition Ltd, Loverock 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' (Ref. TP1281) has been sent to us by Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD. WW415

TRW Semiconductors have a publication describing the design of a linear 100W solid-state amplifier intended for s.s.b. transmitter applications. MCP Electronics Ltd, Alperton, Wembley, Middx 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 Pandicon 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 all relevant parameters which are discussed in terms of two Philbrick Testers (Models 5104 and 5107) WW419

EQUIPMENT

New data sheets from Ferranti, Electronic and Display Equipment Division include data on the 06A/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, Chadderton, 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, Audobon Rd, Wakefield, Massachusetts 01880, U.S.A. WW421

Information on the 'M' range fractional horsepower induction motor is contained in a leaflet from Comtex. The 'M' range motor is slightly larger than the standard range but will develop 50% more power. Comtex Ltd, Senate House, Tyssen St, London E8 2ND WW422

'Hospital Headsets for Patients' is the title of a leaflet describing and illustrating the headsets and also the audio drive units, panel adaptors and a 'pillowphone' produced by Danavox (Gt. Britain) Ltd, 'Broadlands', Bagshot Rd, Sunninghill, Ascot, Berks, SL5 9JW 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 investigation by direct viewing of the effect upon spontaneous and induced nystagmus when fixation is removed by total darkness. Levell Developments, 7-10 Hildreth Mews, London SW12 9RZ 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, LS2Z 4D8, have sent us a leaflet describing several of their products including signal generators, power supplies, printer, digital controller and a 1.5GHz millivoltmeter 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 KGM Electronics. Specifications and description of the M107B and M107D Maxitrons are given. KGM Electronics Ltd, Clock Tower Rd, Isleworth, Middlesex WW429

Advance Electronics Ltd, Raynham Rd, Bishop's Stortford, Herts, have sent us information on their 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 Coutant Electronics Ltd, 3 Trafford Rd, Reading, Berks WW433

We have received a brochure describing the 58 different Kepco type 'PRM' voltage stabilizers covering 4.5 to 240V d.c., with ratings from 60 to 300W. Also included is a guide to the ferroresonant technique of voltage stabilization. Auriema Ltd (UK representatives), 442 Bath Rd, Slough, Bucks, SL1 6BB WW436

Available equipment and services provided by Aveley Electric Ltd, Arisdale Ave, South Ockendon, Essex, RM15 5SR, are given in a publication WW438

We have received leaflets on variable speech control which enables speeding up or slowing down recorded speech with any gramophone or tape recording machine without any change in pitch. Cambridge Research and Development Group, 21 Bridge Square, Westport, Conn. 06880 WW440

Descriptive literature of a Kontakt product 'Positiv 20' — used for the production of individual or small series printed circuits — and other Kontakt products are available from Specialist Products Distributors Ltd, Piccadilly, London W1V 0HL WW441

GENERAL INFORMATION

A bulletin on short advanced 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 environmental 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 Central Services Ltd, Publicity Division, P.O. Box No. 451, Carlton House, Lower Regent St, London SW1Y 4LS WW444

I.T.A.'s annual report and accounts 1970-1971 can be obtained from Independent Television Authority, 70 Brompton Rd, London S.W.3. Price 85p.

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

A list of new films from the Central Film Library (Industrial section) can be obtained from Government Building, Bromyard Avenue, Acton, 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.

A leaflet on a range of glass-to-metal seals has been issued by Cathodeon Ltd, Nuffield Rd, Cambridge CB4 1TF. These seals can be used for taking electrical leads into hermetically sealed or continuously evacuated enclosures WW450

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

'Dimensions of semiconductor devices' (Addendum No.2 (1971) to B.S. 3934: 1965) Price £2

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

Real and Imaginary

by "Vector"

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 pitying 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 — from 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 did try 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 post-graduate 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 perpetuated in a note in 'News of the Month' in the January issue. The abbreviated title of the report referred to in the centre column of p.410 should have been Jofro (Joint Fire Research Organization of the Dept. of the Environment and the Fire Offices Committee) and not Jitco.