WW DIGITAL COMPUTER DESIGN

AUGUST 1967
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Wireless World

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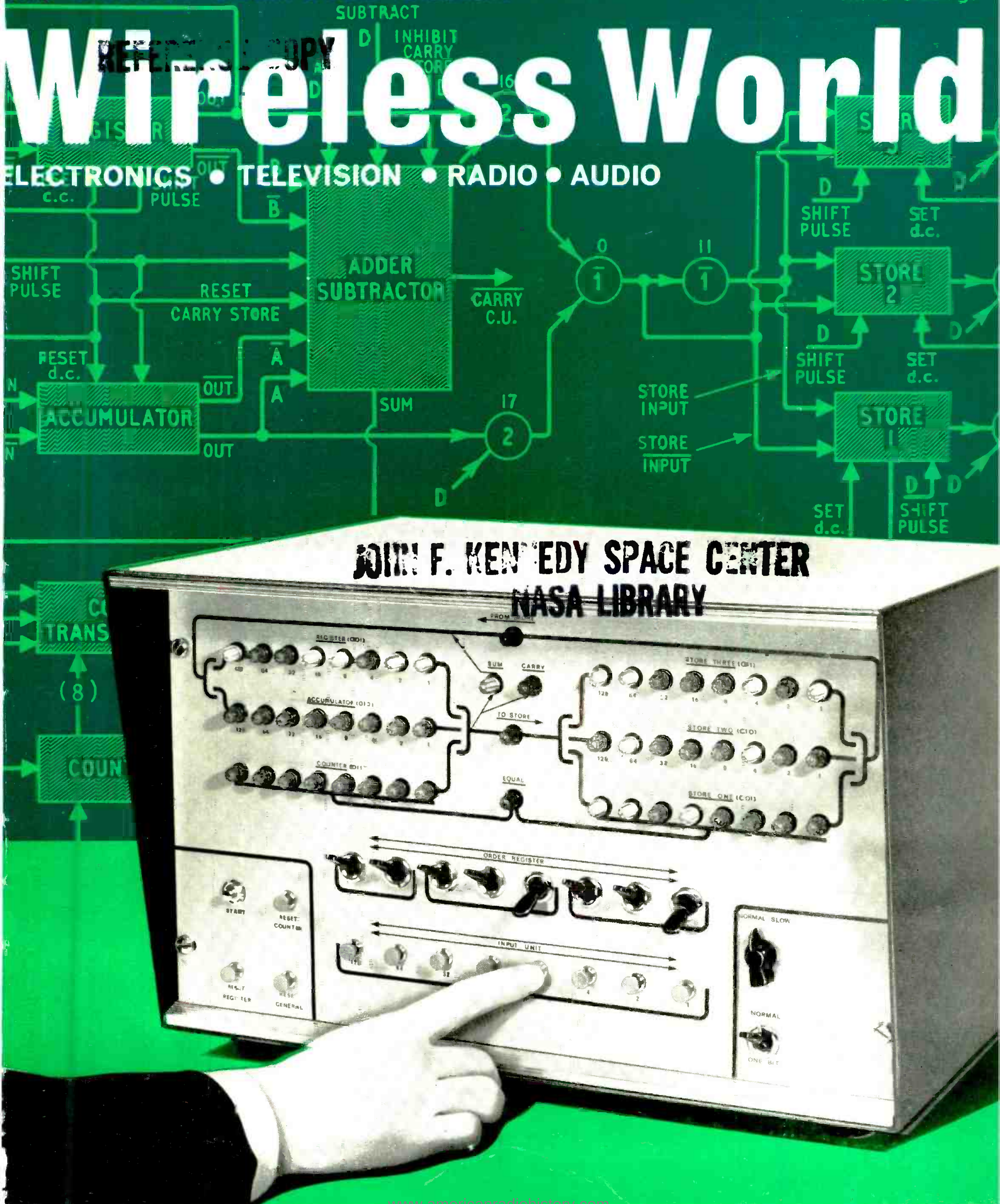

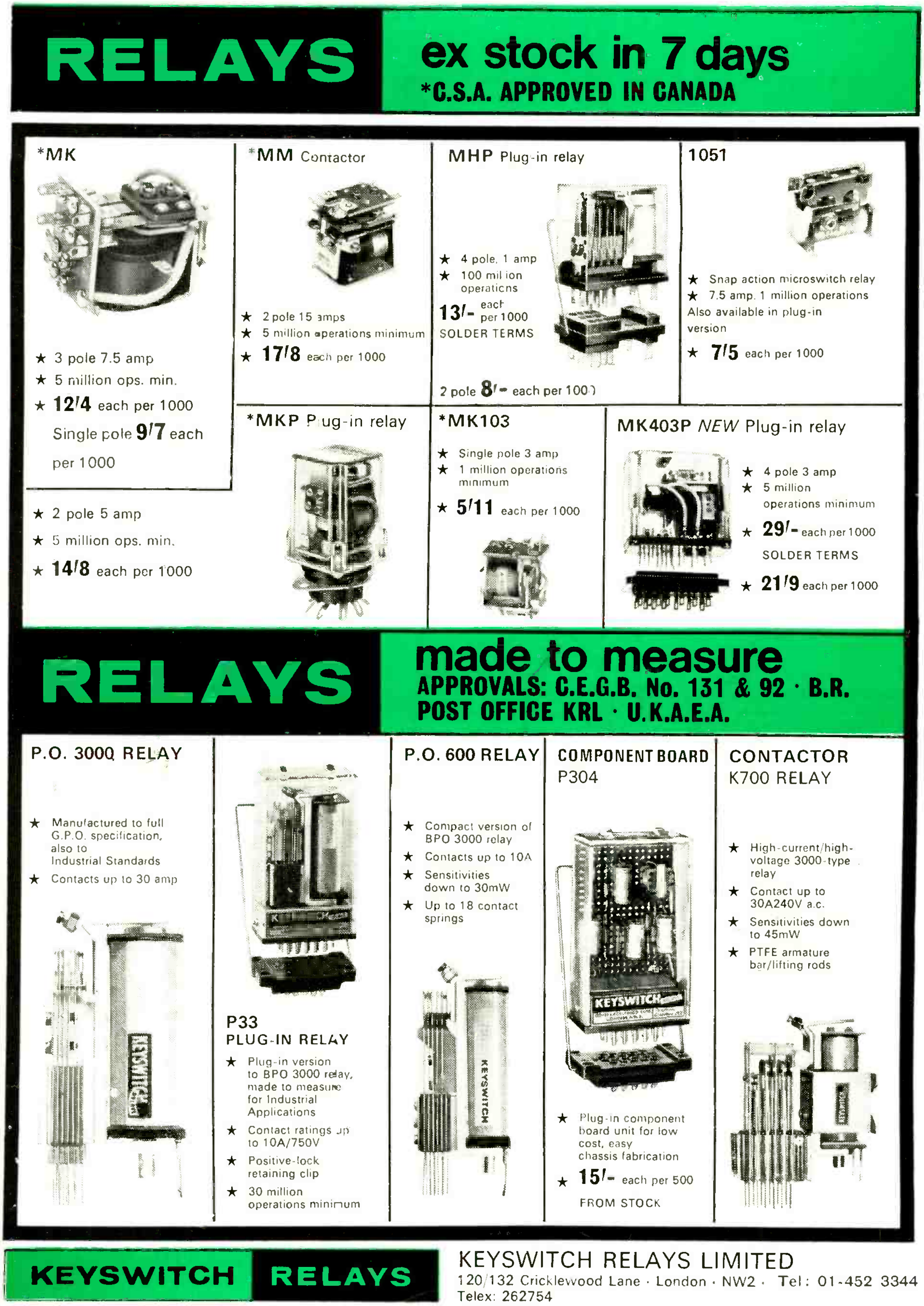

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Wireless World

ELECTRONICS, TELEVISION, RADIO, AUDIO

Colour Television is Launched

ALTHOUGH the daily transmissions of colour television, which the B.B.C. introduced on July 1st, are described as "a colour launching period" they will undoubtedly be considered historically as opening this country's colour service—the first in Europe. There have, of course, been criticisms of the decision to start public transmissions several months ahead of the officially announced opening in December. These have come mainly from manufacturers who feel that by so doing the B.B.C. has put them in the unenviable position of being unable to meet the demand for receivers. This may well be so but the "launching period" will undoubtedly give a tremendous fillip to the colour service. At the moment, of course, the service is available to about 65% of the population but this will have increased to some 70% by December. As announced elsewhere in this issue the Postmaster General has now authorized the building of a further 16 u.h.f. stations for BBC-2 which when completed in about 1974 will bring the total number of high-power stations to 44 (out of the scheduled 64) and the coverage will then be 90% of the population. There will, of course, also have to be a very large number of fill-in stations.

What is the demand for colour receivers? Manufacturers have said that they have orders for several thousands but many of these are undoubtedly for installation in dealers' shops for demonstration purposes. However, enquiries among dealers in several London boroughs have shown that there are many viewers ready to buy or rent receivers as soon as they become available. The relay companies too have received so many requests for colour installations that they have had to start a "waiting list."

What of the prices of receivers; are they prohibitive for the large majority of the population? The short answer is, "yes, for the *large majority*" but the industry will be satisfied for some time to come to meet the needs of the minority, which it is estimated will total some 1M in four years. Many people are asking will the prices of receivers, now at around £300, be reduced substantially in the not-too-distant future. The industry says no, but when one looks at the American scene and sees "end of the range" colour receivers at knock down prices who is to say that, with the abolition of retail price maintenance, it will not happen here.

On the technical side there can be nothing but praise for the B.B.C. and the industry (both the transmission and receiver side) for the quality of picture received. Some of us have been privileged to see the laboratory development of colour over many years and maybe tend to take it for granted. We must not, however, lose sight of the tremendous amount of effort and skill that has been devoted to the development of equipment in this country. It was therefore, to say the least, a little disappointing to find that the cameras used at Wimbledon were not of British manufacture.

Few who have seen the transmissions will now question the choice of the PAL system. One outstanding point in its favour by comparison with N.T.S.C. is the fact that ghosting is so much less objectionable.

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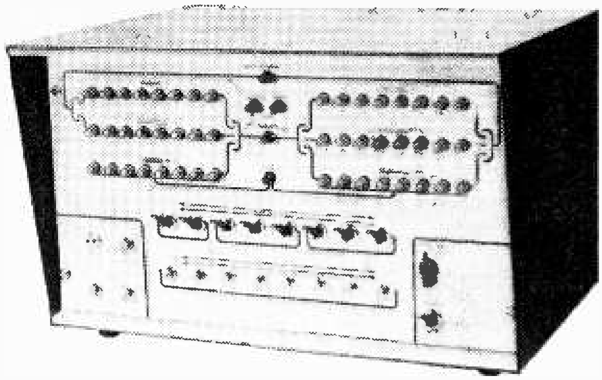
WIRELESS WORLD DIGITAL COMPUTER



1—Outline description: basic theory: circuit elements

Low-cost desk-top binary machine for small-scale calculations and for use in schools as a teaching aid, designed by B. Crank of " Wirless World " staff. Numbers are fed in manually and results of calculations are read from indicator lamps. Instructions, entered in binary coded form by a set of switches, are interpreted and carried out automatically by the machine.

THE *Wireless World* Digital Computer has been designed as a low cost system capable of demonstrating basic computer methods and various operations in the binary number system. It will add, subtract, multiply and divide eight-bit binary numbers, which are entered manually by means of press switches. It will also complement a binary number, and this feature makes possible arithmetic operations with mixed positive and negative numbers and subtraction using the 1s and 2s complement methods. The machine can be programmed to convert natural binary numbers into natural binary coded decimal form, making the job of interpreting results easier. The largest number that can be accommodated by the computer is the maximum obtainable with eight binary digits, which is 255.

Results of calculations and the states of all major circuits are indicated on the front panel by small neon lamps. This means that each computer operation can be analysed in detail and fault diagnosis is made easier. Instructions to the computer to perform required operations are entered, in numerical code form, by means of a set of eight switches, and the machine interprets the code and carries out the instructions automatically.

A choice of three operating speeds is provided. These are: " one bit," in which the start button is used to initiate separately each operation at each successive binary position; " slow," in which pressing the start button causes a complete arithmetical operation (e.g. adding two 8-digit numbers) to be performed at the rate of about 2 binary positions per second; and " normal " which is similar in principle to " slow " but at the higher speed of 2,500 positions per second. In its present form the computer will carry out one instruction—a complete arithmetical operation—at a time. With the addition of a few extra parts a whole sequence of instructions could be carried out automatically, enabling basic programming to be taught.

A simplified block schematic diagram of the computer is shown in Fig. 1. Numerical data are fed straight into the arithmetic unit by the data input unit and are operated on by the computer in a manner determined by the order register, at one of three speeds (mentioned above) selected by the demonstration switching. The order register is the means by which binary coded instructions to the computer to perform a particular operation are fed in and held. The order decoder translates the instruction presented to it by the order register into a form that the computer can " understand," and causes it to be carried out by routing pulses, generated in the control unit, to the correct sections of the computer. The exact number of pulses generated by the control unit will depend on what the decoder " tells " it to do and on the internal condition of the arithmetic unit. Data can be transferred from the arithmetic unit to the store for later use or from the store to the arithmetic unit. The condition of all circuits in the arithmetic unit and store are continuously monitored on the front panel by the readout section so that if desired any particular operation can be analysed in detail.

The computer operates in the " serial " mode, which means that the binary information being transferred along the routes shown in Fig. 1 is represented by time sequences of electrical states. Thus when a number is being handled the digits in the successive binary positions are dealt with one after the other, starting with the least significant digit and working upwards.

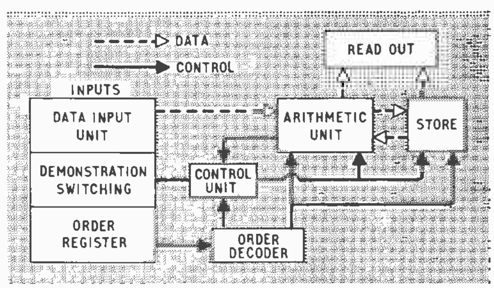


Fig. 1. Simplified schematic showing principal units of the computer.

An important factor in any project is cost and everything possible has been done to keep this within reasonable bounds. The prototype cost something in the region of £50–60, not including the cabinet. The transistors used are, for the most part, reject germanium types available for under 1s each, the remainder being 60 V silicon types that can be bought for about 2s each. An attempt has been made to use resistors of the same value wherever possible so that price reductions can be obtained by quantity buying (say 2d each). The large quantity of diodes used cost about 4½d each. The main difficulty is in obtaining cheap capacitors, as prices range from about 9d to 1s per item. However, if a systematic approach is made to this problem much more favourable prices can be obtained. The method adopted in the prototype was to ask various retailers for quotations for the quantities involved, in this manner a price of 4d per capacitor was achieved.

Before construction of the computer is contemplated it is essential that the intending builder be thoroughly conversant with the theory involved (see " reminder " sections following). Accuracy in construction is equally important, for while trouble-shooting on the correctly built computer is not too difficult, locating faults when wiring errors are involved can be very trying.

Stored-programme facility.—Development work is being done on a sub-routine store for the computer. Early results are encouraging and should the store prove to be reliable it will be described in detail later. Basically it provides a further 64 words of storage (512 bits) that can hold either control words or data. Each word is stored by means of wired-in diodes or by diode " pegs " inserted into a matrix programming board. As the computer completes each operation the next instruction is automatically fed to the computer from the sub-routine store and is executed and this process continues until a " stop " instruction is received by the computer. In this way complete sequences of up to 64 separate instructions can be carried out automatically and basic programming can be taught and demonstrated.

Continued on page 369

REMINDER ON BINARY ARITHMETIC

Binary notation.—In the binary system, only two characters are required for counting, and we shall use the conventional 0 and 1. As in the decimal system the digits have positional as well as numerical value as shown in the table (right).

Decimal Number	Positional value 2^3 (8)	2^2 (4)	2^1 (2)	2^0 (1)	Decimal Number	Positional value 2^3 (8)	2^2 (4)	2^1 (2)	2^0 (1)
0	0	0	0	0	8	1	0	0	0
1	0	0	0	1	9	1	0	0	1
2	0	0	1	0	10	1	0	1	0
3	0	0	1	1	11	1	0	1	1
4	0	1	0	0	12	1	1	0	0
5	0	1	0	1	13	1	1	0	1
6	0	1	1	0	14	1	1	1	0
7	0	1	1	1	15	1	1	1	1

The values of digits in successive positions from right to left are increasing powers of two, 2^0, 2^1, 2^2, 2^3 (or 1, 2, 4, 8). Each binary digit is termed a " bit " and a complete binary number a " word." The weights of the digits in the eight-bit word used in our computer are therefore 2^0, 2^1, 2^2, 2^3, 2^4, 2^5, 2^6, 2^7 (or 1, 2, 4, 8, 16, 32, 64, 128).

To convert a binary number to a decimal number one can add the weights for each column in which a 1 appears. Consider the word 0110 (which from the table can be seen to equal decimal 6). The decimal number is obtained as follows:—
$0110 = (0 \times 8) + (1 \times 4) + (1 \times 2) + (0 \times 1) = 0 + 4 + 2 + 0 = 6.$

A decimal number can be converted into a binary word by successive division by the weights to give successive quotients of 1s and 0s as follows:

```
Convert decimal 163 to binary.
128)163   1
    128
    ---
 64) 35   0
 32) 35   1
     32
     --
 16)  3   0
  8)  3   0
  4)  3   0
  2)  3   1
      2
     --
  1)  1   1
      1
     --
      0
```

Reading the right hand column of quotients from top to bottom, this gives 163 = binary 10100011.

A number in the decimal system is based on powers of 10 and is said to have a radix of 10; similarly the binary system has a radix of 2. To indicate the radix being used, where necessary, the radix will be enclosed in brackets at the end of the number as is standard practice, i.e. $163_{(10)} = 10100011_{(2)}$

In addition to the pure or natural binary system discussed above, the natural binary coded decimal (n.b.c.d.) system is used in the computer. This uses four bits for each decimal place, these bits being the natural binary representation of each decimal digit, i.e.

$163_{(10)}$ in n.b.c.d. is 0001 0110 0011

The instructions to the computer are given in another number system with a radix of 8 known as the octal code. The method used to convert a pure binary number to octal is similar to that for converting n.b.c.d. to decimal. The number is divided into groups of three digits starting from the right, then the decimal equivalent of each group is written down, as follows:—

00/001/001
0 / 1 / 1 $\qquad 00001001_{(2)} = 011_{(8)}$

Another example:—

11/101/111
3 / 5 / 7 $\qquad 11101111_{(2)} = 357_{(8)}$

Because the computer uses only an eight-bit instruction word two bits appear in the left hand group, and therefore the maximum octal number that can appear in the left-hand place is 3.

Binary arithmetic.—This is best started by examining the following rules for adding two binary numbers:—

$0 + 0 = 0$
$0 + 1 = 1$
$1 + 0 = 1$
$1 + 1 = 0$ carry 1 to next most significant position.

Where the addition of two 1s takes place, as there is no symbol to represent any number greater than 1, a carry to the next most significant position occurs. This is the same as saying $2^0 + 2^0 = 2^1$ or $1 + 1 = 1\ 0$. Throughout this article, to indicate in a table that a carry has been generated the following symbolism will be used:

$$1 + 1 = 0 \rightarrow 1$$

The next step is to add two four-bit binary numbers. The working is as follows:—

$$\begin{array}{rl}
2^4 2^3 2^2 2^1 2^0 & \text{(positional values)} \\
1\ 1\ 1\ 1\ + & \\
0\ 1\ 1\ 0 & \\
\hline
1\ 0\ 1\ 0\ 1 & \text{(sum)} \\
\hline
1\ 1\ 1\quad\quad & \text{(carries)}
\end{array}$$

From this it is seen that in order to add two binary numbers it is necessary to be able to add three digits to take into account any carry that may be generated during the previous addition. Another addition table is obviously called for:

A		B		C			
0	+	0	+	0	=	0 → 0
1	+	0	+	0	=	1 → 0
0	+	1	+	0	=	1 → 0
0	+	0	+	1	=	1 → 0
1	+	1	+	0	=	0 → 1
1	+	0	+	1	=	0 → 1
0	+	1	+	1	=	0 → 1
1	+	1	+	1	=	1 → 1

Binary subtraction can be explained by a similar table. However, here a " borrow " can occur instead of the carry in addition. In the following table a borrow is indicated in the same way as a carry i.e. → 1.

A — B		C		
0 — 0	with an	0	borrow	= 0 → 0
1 — 0	,, ,,	0	,,	= 1 → 0
0 — 1	,, ,,	0	,,	= 1 → 1
1 — 1	,, ,,	0	,,	= 0 → 0
0 — 0	,, ,,	1	,,	= 1 → 1
0 — 1	,, ,,	1	,,	= 0 → 1
1 — 1	,, ,,	1	,,	= 1 → 1
1 — 0	,, ,,	1	,,	= 0 → 0

As an example we will subtract 0 1 from 1 0, as follows:—

$$\begin{array}{ll}
2^1\ 2^0 & \text{(positional values)} \\
1\ \ 0 & \text{(minuend)} \\
0\ \ 1 & \text{(subtrahend)} \\
\hline
0\ \ 1 & \text{(difference)} \\
1 & \text{(borrow)}
\end{array}$$

Starting with the right hand column $0 - 1 = 1 \rightarrow 1$. In the left-hand column we have already borrowed a 1, so this has to be returned and we get:—$1 - 0$ with a 1 borrow $= 0 \rightarrow 0$.

Henceforward a borrow will be called a carry and an unnecessary term dispensed with.

Multiplication can be performed by repeated addition (e.g. $8 \times 4 \equiv 8 + 8 + 8 + 8 = 32$) and division by repeated subtraction, e.g. $32 \div 8 \equiv [(\{(32-8)-8\}-8)-8] = 0$. The quotient being obtained by counting the number of times subtraction was necessary to reduce 32 to 0, that is 4.

Subtraction can be performed by use of the addition process, although our computer does not normally operate in this mode. The computer, however, will demonstrate the process, so an explanation is called for here. Two methods can be used, known as the 1s complement method and the 2s complement method.

First the 1s complement method. Consider 1 0 1 1 — 0 1 0 1. First, it is necessary to form the 1s complement of the subtrahend 0 1 0 1. This is done by changing all the 1s to 0s and all the 0s to 1s, to give 1 0 1 0. To complete the subtraction we now add the two numbers and perform the " end around carry " operation:—

$$\begin{array}{rl}
1\ 0\ 1\ 1 & \\
+\ \ 1\ 0\ 1\ 0 & \\
\hline
1\ 0\ 1\ 0\ 1 & \\
+ \longrightarrow 1 & \text{(end around carry)} \\
\hline
0\ 1\ 1\ 0 & \text{(sum, and result of subtraction)} \\
\hline
1\quad &
\end{array}$$

We have used a four bit word and a carry is generated that exceeds our word length. This carry is added to the result of the first addition—hence " end around carry ". Thus 1 0 1 1 — 0 1 0 1 = 0 1 1 0.

In the 2s complement method, first the 2s complement of the subtrahend is formed. This is equal to the 1s complement + 1. Using the previous example again (1 0 1 1 — 0 1 0 1), the 2s complement of 0 1 0 1 = 1 0 1 0 + 1 = 1 0 1 1. Then:

$$\begin{array}{ccccc}
2^4 & 2^3 & 2^2 & 2^1 & 2^0 \\
 & 1 & 0 & 1 & 1 \\
+ & 1 & 0 & 1 & 1 \\
\hline
1 & 0 & 1 & 1 & 0 \\
\hline
1 & & 1 & 1 &
\end{array}$$

but here the carry generated in the 2^4 position is ignored.

Our computer forms the 1s complement of a number by adding to it a series of 1s and ignoring any carry that may be generated.

Thus to form the 1s complement of 0 1 0 1:—

$$\begin{array}{rl}
0\ 1\ 0\ 1 & \\
+\ 1\ 1\ 1\ 1 & \\
\hline
1\ 0\ 1\ 0 & \\
\cancel{1}\quad\cancel{1}\quad & \text{(ignore carries)}
\end{array}$$

Other processes in binary arithmetic, such as operations with positive and negative numbers, will be discussed in the section dealing with programming the computer.

REMINDER ON BOOLEAN SYMBOLS AND LOGIC ELEMENTS

Boolean symbols do not represent quantities but logical states or conditions. For example, the symbols A and B could represent two switches in the " on " position, while $\overline{A}$ and $\overline{B}$ could represent the same two switches in the " off " position. If these switches were connected in series with a battery and lamp, when both switches were " on " this condition would be symbolized as " A and B ", and it would result in the lamp being lit. If any other condition existed, i.e. A and $\overline{B}$ (A on, B off), or $\overline{A}$ and B (A off, B on) or $\overline{A}$ and $\overline{B}$ (both off), the lamp would not light. This is a demonstration of the Boolean AND function—the lamp is lit only when the condition A AND B obtains. If the condition of the lamp being lit is represented by the symbol C then it can be said that

$$A \text{ and } B = C \qquad \text{or } A \cdot B = C \qquad \text{or } AB = C$$

These three statements are identical; the two letters with a dot between, or close together, are shorthand for the AND function. The condition $\overline{A}$ is read as NOT A. Proceeding further with our two switches and lamp we can therefore say that

$$AB = C \qquad A\overline{B} = \overline{C} \qquad \overline{A}\overline{B} = \overline{C}$$

where $\overline{C}$ (NOT C) indicates that the lamp is not lit.

If the switches were connected in parallel then either switch being " on " would result in the lamp being lit. Here it is true to say that the condition " A or B " would result in C. The Boolean symbol for this function is +, which is read as OR. With the switches connected in parallel the following equations are true:—

$$A + B = C \text{ (A OR B = C)}$$
$$\overline{A}\overline{B} = \overline{C} \text{ (NOT A and NOT B = NOT C)}$$

In the above example $\overline{A}$ (NOT A) was represented by the absence of something (absence of conduction path), but it is important to remember that NOT A really means the *presence* of something *other than* A—in fact its opposite state $\bar{A}$. Thus A and $\bar{A}$ can be represented by any pair of defined electrical states: +6V and +2V, 0V and −4V and so on.

Logic elements.—The simplest of the circuits performing the above Boolean functions is the NOT gate. With this, for example, state A can be negated into $\bar{A}$ or *vice versa*, and the graphical symbol for such an element is shown in Fig. 2. If the input of this element goes to earth the output will go to some voltage above earth. The $\bar{1}$ in the centre of the circle indicates that the element will negate one input, while the arrow shows the direction of information flow.

The time has come to introduce one piece of terminology that will be used throughout the series of articles. The fact that a voltage exists at a particular point in a circuit can be represented by the terms "true", "up", 1 (binary), −ve, +ve and so on, and the fact that a voltage does not exist can be represented by "false", "down", 0 (binary), 0V, etc. The use of 1s and 0s to define voltage levels has been rejected, as far as this series is concerned, for fear that they may become confused with binary 1s and 0s, i.e. numerical data. The term "up" will be used to indicate that a voltage is present and "down" to indicate that that line is at earth potential.

An extension of the NOT gate is the NOR gate. Electronically they are almost identical but the NOR gate has more inputs, as shown by the symbol in Fig. 2. In the logic element used in the computer any input going "up" will result in the output going "down" and for the output to be "up" *all* inputs must be "down". More or fewer inputs may be provided.

Next in Fig. 2 is shown the symbol for an element performing the previously discussed OR function. In the practical device any one of the inputs going "up" will result in the output going "up" as indicated by the "1" in the symbol, and the output will be "down" only when *all* the inputs are "down".

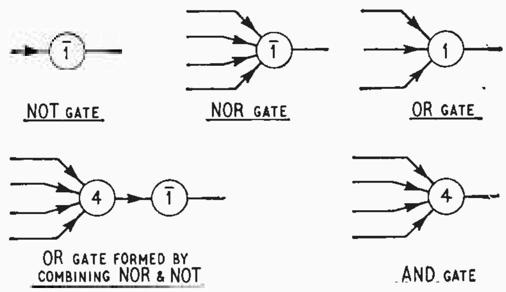


Fig. 2. Symbols for basic logic elements used in the computer.

An OR gate can be formed by the combination of a NOR and a NOT gate as shown in Fig. 2, and in fact this method is used in certain parts of the computer. Any "up" input to the NOR gate will cause its output to go "down", and the resulting "down" input applied to the NOT gate will cause its output to go "up".

The final symbol shown is for an AND gate, the "4" indicating that all four inputs have to be "up" simultaneously before the output will go "up", and (switches-in-series example in Boolean algebra section). More or fewer inputs can be provided, within limits, as required.

continued from page 364

THE BASIC CIRCUITS

In the circuit diagrams Figs. 3-12 an indication is given in the captions of the cost of each circuit block, based on the prices in existence at the time of writing for quantity buying. This, however, is only intended as a rough guide. Components only, and not mounting boards, wire, solder, etc., are taken into account.

The computer employs negative logic, that is, an "up" signal is represented by the presence of a negative voltage. It is very important to note, especially when bistables and flip-flops are discussed, that a change of signal level from "down" to "up" is a negative-going voltage change, and a change of signal level from "up" to "down" is a positive-going voltage change.

The circuit of the NOT gate is shown in Fig. 3. This is an exception to one of the rules explained in the "reminder" section, in that a "down" signal can be represented by an open circuit—a feature that is used to advantage in the computer's order decoder. When the input is earthed the base of Tr1 is held at a positive potential by virtue of the potential divider between the +V supply and earth (0V) formed by R_1 and R_2. Tr1 will therefore be switched off and its collector held at a negative potential. When the input goes "up," Tr1 base goes negative, and Tr1 switches on and connects anything coupled to the output effectively to earth.

The NOR circuit, shown in Fig. 4, can be seen to be almost identical to the NOT circuit except that more inputs are provided. The operation is exactly the same.

The circuit of an AND gate is shown in Fig. 5. When none of the inputs is "up" all the inputs are connected to earth (0V line), hence all the input diodes (D1-D4) are forward-biased by virtue of R_1. As a result the left-hand side of R_2 is connected to earth, and the right-hand side of R_2 and the base of Tr1 are held at a positive poten-

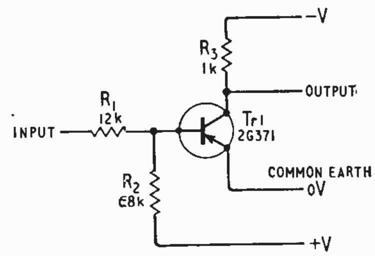


Fig. 3. Circuit of NOT gate. (Cost per gate: 3 resistors at 2d each, 1 transistor at 9d = 1s 3d)

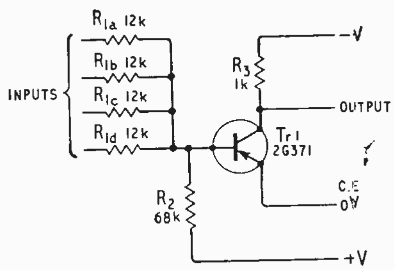


Fig. 4. Circuit of NOR gate. (Cost per gate: 1s 3d as for NOT gate plus 2d per extra input)

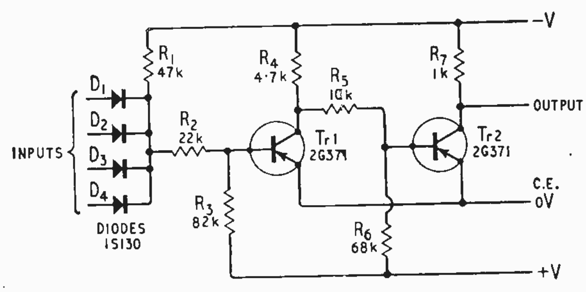


Fig. 5. Circuit of AND gate. (Cost per gate: 7 resistors at 2d each, 4 diodes at $4\frac{1}{2}$d each, 2 transistors at 9d each = 4s 2d)

tial by means of R_3. Tr1 is cut off and its collector is negative. This negative potential is felt at the base of Tr2 which is held in the conducting state and the output is therefore " down." An input going " up " results in its associated diode becoming reverse-biased; this, however, has no effect on the circuit as the other three input diodes remain forward-biased and Tr1 base is still held positive. When all the inputs are present, however, Tr1 base goes negative by virtue of the potential divider formed by R_1, R_2 and R_3 between the negative and positive supply lines. Tr1 switches on and its collector falls to almost 0V. Tr2 base goes positive, because R_6 is connected to the positive rail, and the output goes " up " as Tr2 switches off.

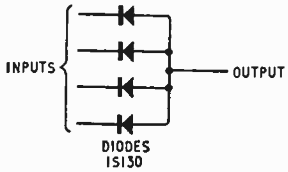


Fig. 6. Diode OR gate. (Cost 4½d per input)

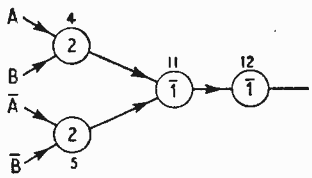


Fig. 7. Logical diagram of comparator.

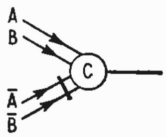


Fig. 8. Comparator symbol that will be used in system diagrams.

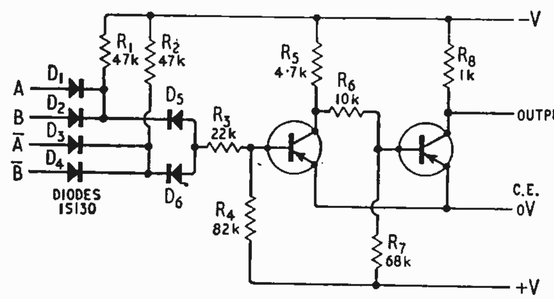
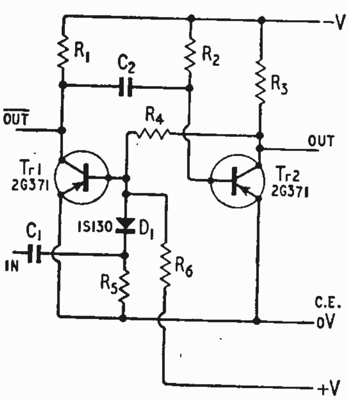


Fig. 9. Comparator circuit. (Cost per comparator: 8 resistors at 2d each, 6 diodes at 4½d each, 2 transistors at 9d each = 5s 1d)

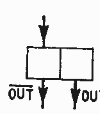


COMPONENTS LIST

	TYPE 1	TYPE 2	TYPE 3
R_1	1k	500Ω	270Ω
R_2	22k	10k	5·6k
R_3	1k	500Ω	270Ω
R_4	10k	4·7k	2·7k
R_5	4·7k	4·7k	4·7k
R_6	47k	22k	12k
C_1	0·01μ	0·01μ	0·01μ
C_2	*	*	0·047μ

* C_2 FOR TYPES 1 & 2 SHOWN ON LOGICAL DIAGRAMS

WHERE MORE THAN ONE TRIGGER INPUT IS REQUIRED REPEAT D_1, R_5 & C_1



SYMBOL USED FOR FF1 FF2 FF3

Fig. 10. Flip-flop (monostable) circuit and symbol (Cost per flip-flop: 6 resistors at 2d each, 2 capacitors at 4d each, 2 transistors at 9d each, 1 diode at 4½d = 3s 6d)

In certain parts of the computer, as has already been stated, a " down " signal can be an open circuit. In these cases the OR circuit shown in Fig. 6 is used. In all other cases the OR function is performed by a combination of the NOR and NOT gates as described in the " reminder " section on logic elements.

To enable the computer to multiply, as will be seen, it is necessary to compare the states of two circuits, *a* and *b*. We will call the outputs of these circuits (indicating their states) A, $\bar{A}$ and B, $\bar{B}$. A comparator is required for this purpose, and its output must be " up " when one of two conditions exists: A and B present or $\bar{A}$ and $\bar{B}$ present. The output must be down when the conditions $A\bar{B}$ or $\bar{A}B$ exist. There is a large number of possible ways of performing this operation, the logical layout used in the computer being shown in Fig. 7. The numbers written outside the gates are their circuit reference numbers, and this method of identification will be used throughout the computer description. The comparator is made up from two AND gates, one NOR gate and one NOT gate, the NOR and NOT gates forming an OR gate. When AB are present AND gate 4 will open and the output of NOT 12 will be " up." When $\bar{A}\bar{B}$ are present AND gate 5 will open and the output of NOT 12 will again be " up." For any other combination of A and B the output of NOT gate 12 will remain " down."

To economise on components the comparator is built as a single circuit and is depicted by the symbol in Fig. 8. the $\bar{A}\bar{B}$ inputs having a line drawn through them. The circuit is shown in Fig. 9. Here, assume that inputs A and $\bar{B}$ are " up " ($\bar{A}$ and B being therefore " down ") D1 and D4 will be reversed-biased by these input signals and D2 and D3 will be forward-biased by R_1 and R_2. As a result the lower ends of R_1 and R_2 will be connected to earth, together with the left-hand side of R_3 via D5 and D6. The base of Tr1 will be positive and the remainder of the conditions that exist will be as for the previously described AND gate. If the input condition changes to, say, A and B up, D1 and D2 become reverse biased, Tr1 base goes negative via R_1 and, as for the AND gate, the output goes " up." While this condition exists D6 becomes reverse-biased, preventing the input circuits from interfering with one another. The action for the input $\bar{A}\bar{B}$ is similar, R_2 providing the drive for Tr1 and D5 becoming reverse-biased.

Three types of flip-flop are used, the actual choice being determined by circuit requirements. The circuit, together with the symbol used in each case, is shown in Fig. 10.

The flip-flop provides a convenient means of introducing a time delay and obtaining reasonably shaped pulses of known width. In its stable state Tr2 will be held switched on by R_2, Tr2 collector will be at almost 0V and the base of Tr1 will be positive because of R_6 being connected to the positive line. The flip-flop will remain in this condition until a negative-going edge is applied to C_1 (a signal change of " down " to " up "). This will switch Tr1 on, resulting in a positive-going change at Tr1 collector, and this is felt at Tr2 base via C_2. Tr2 switches off and Tr1 is held on by R_4, Tr2 collector now being negative. The flip-flop will remain in this state for a length of time determined by the time constant of C_2 R_2, and when C_2 has discharged the flip-flop will regeneratively switch back to its original condition.

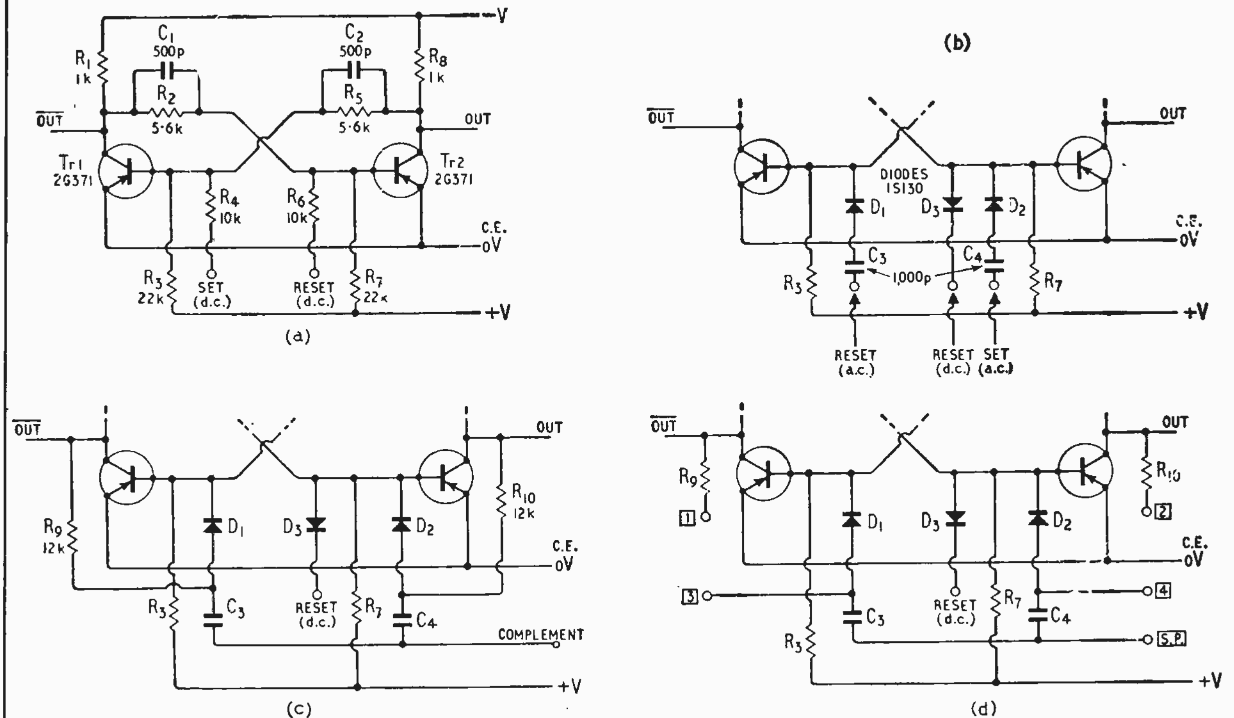


Fig. 11. (a) Set/reset d.c. bistable circuit (Cost per d.c. bistable: 8 resistors at 2d each, 2 transistors at 9d each = 3s 6d). At (b), modifications to (a) to produce set/reset a.c. bistable. (Cost per a.c. bistable = 4s 11½d). At (c), modifications to (a) to produce counter bistable. (Cost per counter bistable: 5s 3½d). At (d), modifications to (a) to produce shift register bistable. (Cost 5s 3½d)

A further note regarding terminology is in order here, the flip-flop has two outputs which are labelled $\overline{\text{OUT}}$ and OUT respectively. In the text they will be referred to respectively as the NOT output and the output. In the stable state of the circuit the output is " down " and the NOT output is " up." When triggered the reverse is true.

The bistable is used throughout the computer in large numbers. The basic circuit used is the same in all cases, though quite a number of variations occur. The actual type of bistable being used can be deduced from the number and nature of the inputs. However, four main basic types emerge, these being the set/reset d.c. bistable, the set/reset a.c. bistable, the counter bistable and the shift-register bistable. The circuits of all four types are shown in Fig. 11 (a) (b) (c) and (d).

The set/reset d.c. bistable (a) will be described first. When power is applied to the circuit it will assume a state determined by the various component tolerances. Let us assume that Tr1 switches on. Tr1 collector will be at nearly 0V so Tr2 base will be positive (R_7). The collector of Tr2 will be negative as will the base of Tr1 (R_5), holding Tr1 in the on condition. The circuit will remain in this state until something is done to disturb it. If a short negative pulse is applied to the reset terminal, this switches Tr2 on, and, by virtue of the cross coupling resistors, R_2 and R_5, Tr1 switches off. The switching action is a regenerative one in that the voltage changes at each collector are felt at the opposite base where they are in the right direction to assist the switching. Capacitors C_1 and C_2 are commutating or speed-up capacitors that reduce the switching time. When a pulse is applied to the reset terminal the output does " down " and the NOT output goes " up." When a pulse is applied to the set terminal the output goes " up " and the NOT output goes down.

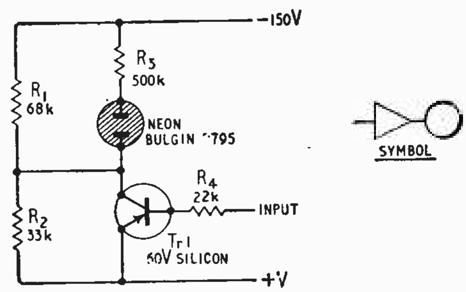


Fig. 12. Neon indicator stage and symbol. (Cost of stage: 4 resistors at 2d each, 1 transistor at 2s 0d and 1 neon at 2s 9d = 5s 5d)

The set/reset a.c. bistable (b) is very similar except that the trigger pulses are a.c. coupled as shown. Here a positive pulse is used, or a positive-going voltage change (" up " to " down "). This is applied to the base of the transistor that is switched on, the positive edge turning it off and reversing the state of the circuit. Because of this the set/reset connections are transposed. An additional reset input is provided which is d.c. coupled, and this enables the starting condition of the bistable to be established. In some cases it is necessary to provide more than one set or reset input; this is achieved by duplicating D1 and C_3 or D2 and C_4 as required.

The counter bistable (c) is an a.c. coupled bistable with a gating facility. The starting condition is established by applying a negative pulse to the reset terminal. Tr1 will now be switched off, its collector will be negative and its base positive. These potentials are applied to D1 in such a way as to reverse-bias it. D2 is, however, forward biased. If now a positive pulse is applied to the "complement" terminal, D2 can conduct, but the reverse-biased D1 cannot. In conducting D2 applies a positive pulse to the switched-on Tr2, turning it off and reversing the condition of the bistable. Now the state of affairs has changed and D1 is forward-biased and D2 reverse-biased, so a subsequent pulse applied to the "complement" terminal will be steered by the diodes to Tr1 base to turn it off. From this it can be seen that each positive pulse applied to the "complement". input will reverse the state of the bistable.

The shift register bistable has no additional components, R_9 and R_{10} are not connected to D1 and D2 but are brought out as outputs. The complement terminal has been relabelled S.P. This bistable will be dealt with fully in the appropriate section of the computer description.

Indication of the states in various parts of the computer is provided by neon lamps driven by 60V silicon transistors (Fig. 12). A fixed potential is applied across the neon and R_3 that is below the striking voltage, and another voltage, within the transistor rating, is applied across the transistor. Both voltages are supplied by the potential divider R_1 R_2. When a logical "up" signal is applied to R_4 the transistor switches on and the neon strikes and remains in this condition until the "up" signal is removed.

(Next month we will begin to consider the overall system of the computer.)

ELECTRONIC MUSIC

"IN *Mixtur* the sounds of a woodwind ensemble, a brass ensemble, and two string ensembles—one pizzicato—(seated in four groups around the audience) are picked up by microphones and put into four ring modulators; the four groups of microphones lead to four mixing tables, where sound engineers control the balance of the various microphones and the input levels for the ring modulators (during the last public performance of *Mixtur* at Stockholm in October, 1966, a total of 36 microphones was used, one microphone for each stand with two musicians). Four players, each using a sine wave oscillator with continuous frequency control, produce sine waves with which the instrumental sounds are modulated by the ring modulators. The results, reproduced over four separate loudspeakers, are blended with the orchestral sound. From each instrumental sound there arises a Mixtur-sound. (By 'Mixtur',one usually refers, regarding organ stops and also choral and orchestral melodies, to a mixing of parallel pitches. It is then a matter of timbral texture from overtones or parallels of chromatic intervals.) The fifth instrumental group of *Mixtur*, consisting of three percussionists each playing a cymbal and gong, is provided with contact microphones connected to three separate loudspeakers. So a composition of differentiated timbres—which I had heretofore been able to achieve only with electronically produced sounds—becomes possible with the use of instruments."

So writes the composer Karlheinz Stockhausen in the first issue of a new quarterly journal *Electronic Music Review*. Established "to provide a source of information and a means of discourse on all aspects of electronic music," this American publication contains a mixture of semi-technical articles describing equipment and general articles on organizational and æsthetic matters. "The multiplier-type ring modulator" shows how various audio signals, produced from conventional musical sources and from artificial sources such as oscillators and noise generators, may be modulated by each other to obtain various effects; while a group of articles on "programmed control" discusses methods by which sources may be controlled in pitch, amplitude, bandwidth, duration, etc., using punched tape, sequencing devices and computers. As one of the authors says, "the variety of results obtainable is as limitless as the imagination of the user." It will be interesting to see whether the use that is made of this superabundance of technical means in the years to come will refute the widely held belief that art thrives on restriction of means—or support it!

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Conferences and Exhibitions

EDINBURGH

Aug. 5-10
Information Processing Congress
(International Federation for Information Processing, 23 Dorset Sq., London, N.W.1)

LEEDS

Aug. 30-Sept. 6
British Association Annual Meeting
(British Association for the Advancement of Science, 20 Great Smith St., London, S.W.1)

OVERSEAS

Aug. 13-17 Miami Beach
Energy Conversion Engineering Conference
(M. Altman, Univ. of Pennsylvania, 113 Towne Bldg., Philadephia, Pa.)

Aug. 25-Sept. 3 Berlin
German Radio Exhibition
(Berlin Ausstellungen, 22 Messedamm, 1 Berlin 19)

Aug. 28-Sept. 2 Lausanne
Computation Congress
(International Association for Analogue Computation, Wasserwerkstrasse 53, CH 8006 Zurich)

Aug. 29-31 Ithaca, N.Y.
High Frequency Generation and Amplification
(H.F. Conference Committee, School of Electrical Engineering, Phillips Hall, Cornell University, Ithaca)

D.C. Power Supplies

Design aspects of d.c. power supplies to different sections of the colour television receiver

By T. D. TOWERS,* M.B.E.

IN colour receiver circuits you will find d.c. power supplies of the same kind as in black-and-white, but with some design features peculiar to colour. At the time of writing, most new designs use shadow-mask colour tubes and a combination of transistors and valves ("hybrid" sets). In these you will find five different kinds of d.c. supplies: (1) 25-kV e.h.t.; (2) 5-kV focus voltage; (3) 800-V boost voltage; (4) 250-V h.t.; and (5) 24-V transistor rail. The voltages quoted for each supply are typical design values, but may vary quite a bit in different receiver designs.

E.H.T. SUPPLY

It is in the e.h.t. supply for the final anode of the colour picture tube that you will see most difference from black-and-white sets. First, the three-gun colour tube calls for much higher currents, voltages and powers than the single-gun, black-and-white tube. Secondly, the e.h.t. voltage in the colour receiver has to be kept much more steady, and thus calls for a separate voltage regulator circuit, not found in black-and-white.

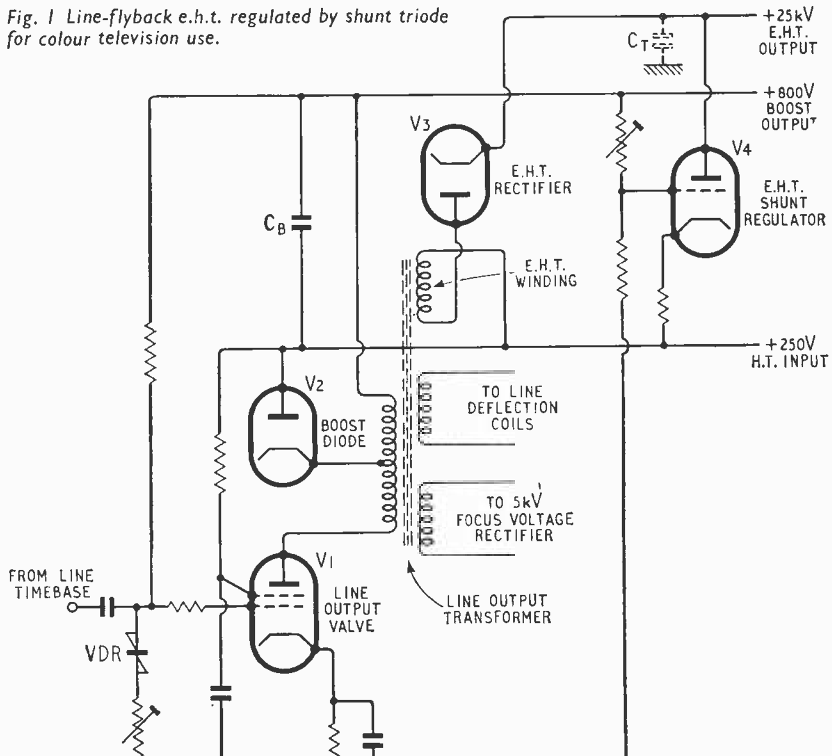
Now, three things can cause undesired changes in e.h.t. voltage, variation in picture tube current ("load"), mains voltage ("line") or component ageing ("drift"). Modern circuits take all these into account.

In Fig. 1 you will find a simplified circuit to illustrate the problems of providing an adequately stabilized e.h.t. supply for the colour receiver. As in black-and-white practice, the source of e.h.t. is a multiple-turn winding on the line output transformer. Of course, the first purpose of the line output valve, V1, is to provide forward sweep drive to the line deflection coils. However, when the valve is cut off on line flyback, a high positive back e.m.f. is generated across the e.h.t. winding. This is rectified in the diode, V3, to provide an e.h.t. d.c. supply.

Also, as in black-and-white practice, a diode, V2, and capacitor, C_b, in Fig. 1 provide a subsidiary "boost" high-voltage rail, several hundred volts higher than the h.t. input voltage. (The boost supply will be dealt with more fully later.) One function of the boost voltage is to feed back through a resistance to the voltage dependent resistor, *VDR*, in the grid circuit of the line output valve, and thus to provide automatic adjustment of the output of V1 for valve spreads and ageing as well as mains voltage variation.

So far, the e.h.t. circuits have been in essence the same as for black-and-white. On the other hand, the e.h.t. regulator valve, V4, in Fig. 1, is peculiar to the colour receiver, which cannot tolerate variations of e.h.t. that would be permissible in black-and-white. This is because colour "trueness" (as governed by purity and convergence) is sensitive to e.h.t. voltage changes. Also, as average scene brightness varies, the three-fold change of beam current inherent in the three-gun tube would, without the use of e.h.t. regulation, lead to unaccept-

Fig. 1 Line-flyback e.h.t. regulated by shunt triode for colour television use.



*Newmarket Transistors Ltd.

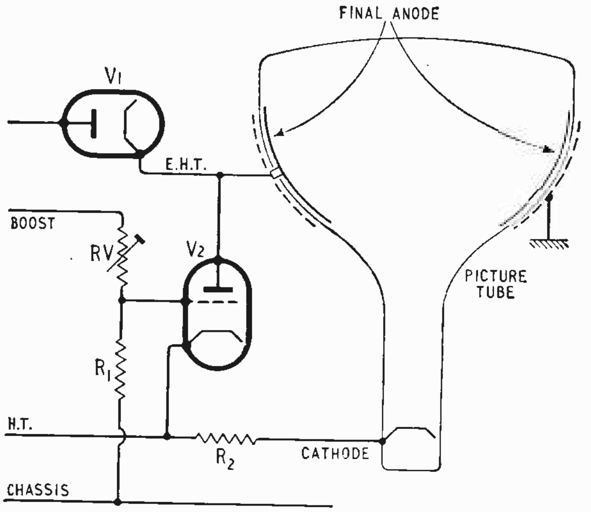


Fig. 2. Basic e.h.t. regulation circuit

able changes of picture size (height and width and focus.

Fig. 2 shows a simplified diagram of the shunt e.h.t. regulator basic arrangement employed in most colour sets. In this, the valve, V2, forms a variable path for current, placed in shunt with the picture tube cathode-to-final-anode across the e.h.t. rail. It operates to maintain a constant load on the supply. When scene brightness increases, the picture-tube beam current also increases, but the e.h.t. voltage does not fall because the anode current of the regulator decreases an equal amount to keep the total e.h.t. current load constant.

The regulator works like this. Its cathode is connected to the h.t. rail, and its anode to the e.h.t. Its grid voltage is obtained from a voltage divider between the boost rail and the h.t. negative. The bias is adjusted initially by *RV* with the picture cut-off (black) until the anode current of V2 equals the total maximum current that will be taken by the tube final anode on the brightest scenes to be displayed (typically 1.2 mA at 25 kV). In normal operation, if the e.h.t. starts to drop because the picture tube beam current rises, the boost voltage (being obtained by line-flyback from the line output transformer) also falls. As a result, the bias voltage on the grid of the regulator valve also falls and causes its anode current to fall. Thus, as beam current rises, the regulator current falls; and vice versa.

Apart from the special feature of the regulator valve, colour e.h.t. supplied, as mentioned earlier, have to deal with higher voltages, currents and powers than black-and-white. This means that the valves in Fig. 1 are much "beefier" than usual. An example of a special colour line-output valve is the Mullard PL505 with the massive ratings of 7.5 kV peak anode volts, 1.4 A peak anode current and 32 W anode-plus-grid dissipation. Matching this is the PY500 boost diode with peak ratings of 5.6 kV on voltage and 0.8 A on current, and the GY501 e.h.t. rectifier with peak ratings of 31 kV on voltage and 1.7 mA on current. The high e.h.t. of around 25 kV common in colour sets means also that the regulator circuit requires a special device such as the PD500 with 25 kV design-centre anode voltage rating and a 30 W (intermittent 35 W) dissipation rating. All these valves use B9D bases, and have top-cap anodes (except for the PY500, which has a top-cap cathode) together with a special construction to prevent internal flashover.

The e.h.t. smoothing is obtained by the capacitance, typically 2,000 pF, between the final anode coating on the inside of the picture tube bulb and the graphite coating on the outside, indicated by the dotted capacitor, C_T, in Fig. 1. The e.h.t. rectifier heater supply, not shown, is obtained conventionally from an isolated winding on the line output transformer.

The 25 kV of the e.h.t. brings certain practical difficulties. E.H.T. points and connectors are usually housed in small anti-corona shields so that they will not corona under humid and dusty conditions. Again, flashover is a greater danger than in black-and-white sets and special precautions against this, such as capacitors with inbuilt spark gaps, are often incorporated in colour designs. Furthermore, any stream of electrons with an accelerating voltage over 16 kV generates X-rays when the electrons collide with metallic electrodes and screens. This gives rise to an aggravated X-ray hazard in colour receivers. Special X-ray-absorbing lead glass is used for colour tubes and high-voltage valves and reduces the level of X-radiation outside the receiver to a safe level for domestic use. This "safe" level is of the order of 0.5 millirontgen per hour. *But service engineers should not spend long periods close to unshielded e.h.t. rectifier and regulator valves with the set outside its cabinet.*

FOCUS VOLTAGE SUPPLY

In Fig. 1 a separate winding is shown on the line output transformer, labelled "To 5-kV focus voltage rectifier." The flyback output from this winding is rectified and used to provide high voltage d.c. to the three commoned second anodes of the shadow-mask colour tube. Beam focusing is done by varying this second-anode d.c. voltage. The required voltage turns out to be in practice somewhere between 16.5% and 20% of the nominal 25-kV final anode voltage, i.e., 4.2-5.0 kV. Some modern black-and-white sets also use electrostatic focusing, but the focus voltages required have been low enough (at a few hundred volts) to be derived from the h.t. or boost rails. But colour sets with a 5 kV requirement have to have a separate special focus voltage supply not found in black-and-white.

Fig. 3 shows the basic arrangement for rectifying and smoothing the a.c. focus-supply voltage from winding (a) of the line output transformer. The high nominal 5-kV focus voltage calls for a special rectifier valve, V, such as the Mullard EY51. As with the e.h.t. rectifier,

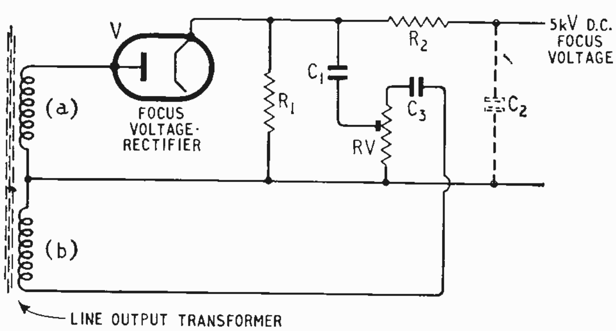


Fig. 3. Obtaining 5-kV d.c. supply for adjustable focus voltage on second anodes of shadow-mask colour picture tube.

an isolated special winding on the line output transformer (not shown) supplies the heater voltage for the focus voltage rectifier. However, the semiconductor rectifier is replacing the valve in this application because it does not need a heater and this lightens the load on the already hard-worked line output transformer stage. The resistor R_3 is designed to protect the picture tube from damage due to internal arcing. C_1, R_2 and C_2 form a smoothing filter, where C_2 represents the total of the picture tube second anodes capacitances, together with lead screening and other stray capacitances. The variable resistor, *RV*, enables the supply output voltage to be varied for focusing adjustment. This fine adjustment is made by varying the amplitude of the pulse applied to the reservoir capacitor, C_1, by the potentiometer connected across a separate winding (b) on the line output transformer. This arrangement reduces the voltage from the control spindle to chassis.

The method shown in Fig. 3 is only one of many different methods, some involving separate inductors, used to adjust the focus voltage safely.

BOOST VOLTAGE

In both colour and black-and-white television receivers, some circuits, such as the field timebase sawtooth oscillator, require a supply voltage higher than the 250-V h.t. rail. This requirement is met by a "boost" supply which in the colour receiver is derived in the same way as in black-and-white. The main difference between the two is that the boost voltage for colour generally lies around 800 V, while in black-and-white it is somewhat lower at around 500 V.

It must be made clear that although we are here primarily interested in the boost voltage as a supply for various circuits which need a voltage greater than that of the h.t. line, this is really a very secondary application. The boost circuit exists to obtain efficiency in the line scan circuit and without it the line scan stage would need an almost impracticable input power.

In the basic circuit of Fig. 4, we assume that the inductance of the transformer is very large compared with that of the deflection coils so that we can ignore the magnetising current in comparison with that in the coils. In practice, this is not true, but the difference is merely to modify the effective inductance of the circuit and to make the deflection-coil current a fraction of the total current. It does not affect the general way in which the circuit functions.

The deflection coil has inductance L and is shunted by capacitance C_s which is the total effective stray circuit capacitance. With a linearly changing current in L there is a constant inductive back-e.m.f. across L. This is stepped up by the transformer and acts against the h.t. supply so that V1 anode is some 80 V above cathode.

These conditions exist at the start of flyback, when V1 is suddenly cut off by a negative-going pulse applied to its grid. The current in L is then at its maximum and there is a large amount of energy stored in its magnetic field. Because V1 is cut-off this current starts to fall and it flows into the circuit capacitance C_s, charging it so that the voltages transferred to the anode of V1 and the cathode of V2 are reversed in polarity. This cuts off V2 and effectively the circuit comprises only L, C_s and the transformer.

The current in L continues to decay and the voltage on C_s to rise until the current is zero and the energy has been all transferred to the capacitance C_s. The voltage is now at its maximum and reaches about 6 kV on the anode of V1.

The valves are still cut-off and now C_s starts to discharge through L so that the current in L reverses in direction and grows from zero in that reverse direction. This continues until the voltage on V1 has fallen to about 80 V and the current in L is large and in the opposite direction to that at the end of the scan. If there were no circuit losses, it would be of the same magnitude as at the end of the scan, but in practice losses make it not quite so great.

This current cannot flow in V1 even if V1 now becomes conductive because it is in the wrong direction. It is the purpose of V2 to carry this current and it becomes conductive automatically at the right point when the voltage across the deflection coil, as modified by the transformer ratio, reaches the proper scan value. The current then flows through V2 into C_B, which is large enough for the voltage across it to be substantially constant, despite the energy which it gains or loses during the scan cycle.

The current now decays fairly linearly, because, in effect, V2 clamps the voltage across L at a constant level. V1 may or may not conduct while the current is decaying; usually it does and provides a small surplus of current which flows through V2. When the current in L becomes zero again, V1 drives current through L and this current now flows out of C_B, so this capacitor loses as much energy during this part of the scan as it gains during the first part.

Depending on the circuit design V2 may or may not be conductive during this second part of the scan, but it usually is conductive. C_B is of the order of 0.5 μF and in a colour receiver becomes charged to around 500 V. This adds to the h.t. supply of 250 V and so makes the effective h.t. supply for V1 about 750 V. A small current can be drawn off from this to operate other stages without seriously affecting the efficiency of the line output stage, and this boost voltage helps considerably in the design of the field timebase.

In practice, the circuit resistances complicate the action somewhat but the technique of line output stages of this kind is now so well known that the methods of compensating for the effects of resistance cause little or no difficulty. The circuit in one form or another has been used in every monochrome television set for well over 10 years. The differences in a colour receiver are merely

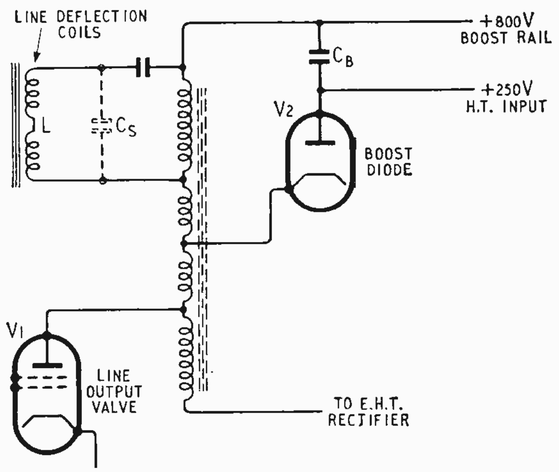


Fig. 4. How 800-V d.c. boost voltage rail is supplied by flyback current pulses through efficiency diode.

the modifications needed to adapt it to the much larger currents and voltages involved.

Because the arrangement of components in an actual colour circuit is a little different from that of Fig. 4 the boost rail has flyback pulses of about 700 V superimposed upon the 800 V d.c. value. For this reason, any supply derived from the boost line must be smoothed by the insertion of a series impedance.

We have used the term "boost" diode in this discussion, but you may come across the alternative names "efficiency," "economy" or "damper" diode.

Ultimately it is to be expected that a semiconductor diode will be found suitable for boost diode applications. At present, however, the tendency is to use thermionic diodes, although of a rather special kind. The PY500, given as an example previously, has a 6.3 V heater-to-cathode voltage rating, which enables it to withstand the very high peak back e.m.f. the cathode sees on flyback. The boost diode in colour differs in one other respect from black-and-white. With mean forward current of around 350 mA, it requires a higher current and power rating than before; so much so that some designs use two black-and-white boost diodes in parallel for colour.

H.T. VOLTAGE

Current colour receiver designs are now usually a.c./d.c. to avoid the use of a mains transformer. They tend also to standardise on a 250 V nominal h.t. rail voltage, which can be obtained by conventional direct rectification of the 230 V a.c. mains. Full or half-wave bridge rectification is used, and valves or semiconductor diodes may be employed. In practice, some arrangement is used on the lines of the basic circuit shown with semiconductor rectifiers in Fig. 5. The mains voltage, after rectification in the bridge circuit, is smoothed by a simple *RC* "brute-force" filter before being applied to the load, R_L.

The absence of a mains transformer means also that the heaters of the valves must be operated in series across the mains input, except where the heater voltages are derived from special windings on the line output transformer. The series-string valves are usually 300 mA heater type. (The 6.3 V, 0.9 A a.c. for the picture tube heater is usually supplied by a separate winding on the mains transformer used to provide the low voltages for transistor circuits described below.)

One last point worth noting is that many designs in textbooks show a 400 V h.t. rail, as this used to be a common American value. Current practice seems to be tending towards an h.t. rail of around 250 V, mainly because the efficiency of modern, scientifically designed, line-output auto-transformers, coupled with more efficient deflection yokes, makes it unnecessary to go to 400 V in the h.t. line to get the necessary deflection swing.

TRANSISTOR RAIL VOLTAGE

Now that transistors have filled over three quarters of the sockets in colour receivers, each design must include a low-voltage d.c. supply rail, usually around 24 V, for the transistor circuits. In an a.c./d.c. receiver without a mains transformer, it is difficult to provide such a low voltage without going to a small separate conventional step-down transformer supply of the basic form of Fig. 6(a). As most transistors used are n-p-n silicon, calling for a positive supply rail, the output is shown as +24 V d.c.

Much effort is being expended in finding a satisfactory

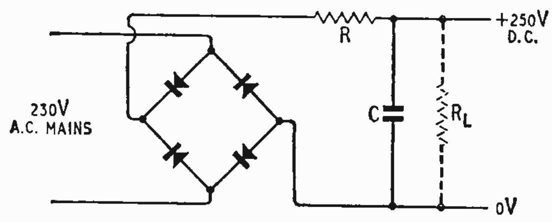


Fig. 5. Basic circuit of mains-derived h.t. d.c. supply for colour receiver.

alternative low-voltage d.c. supply without the use of a mains transformer. The trouble is that this can normally only be derived as a by-product of high voltage circuits, and these are left only in the power output stages. It is difficult to say how the problem will be finally solved, but the RCA circuit shown in Fig. 6(b) gives one interesting answer. The line output transistor (which could equally well be a valve, operating as it does from a 125-V h.t. rail) has a decoupling reactor in its emitter circuit which provides, at its two outputs, voltages of around 21/22 V for different sections of the receiver. This circuit functions as a true d.c. to d.c. transformer, drawing 200-350 mA (depending on the brightness setting) at 125 V and delivering 600 mA at 23 V.

COMMERCIAL DESIGN

To bring together the various points discussed in principle above, Fig. 7 gives relevant parts of the line output stage of a practical circuit of Mullard illustrating the commercial design of e.h.t., focus voltage and boost voltage supplies for a colour receiver.

Drive from the line oscillator comes in to the grid of the PL505 line output valve, which has the line output transformer in its anode circuit. The e.h.t. supply is derived during the line flyback period from the high-voltage overwind 12-13, the output of which is rectified in the GY501 to give the required 25 kV output with respect to earth. Heater voltage for the GY501 is provided from a separate isolated winding, 10-11, on the line output transformer. Smoothing of the e.h.t. is effected conventionally by the picture tube final anode capacitance.

The PD500 valve in Fig. 7 is the shunt e.h.t. voltage regulator, whose grid voltage (and thus quiescent anode current) is set by the 500-kΩ preset potentiometer from the boost voltage point, 6, on the line output transformer. The 1-kΩ cathode resistor of the PD500 introduces such a large amount of negative feedback that, provided the 500-kΩ preset is replaced by a 5% tolerance 470 kΩ, this is sufficient to take care of any spreads in the PD500 characteristics or fall of emission during life. The diode, BA115, in the grid circuit of the regulator valve prevents the grid-cathode potential from becoming excessive when the picture tube is heavily conducting; i.e. when the shunt regulator reaches the end of its control and is cut off. The circuit is designed to set the anode current of the PD500 at 1.2mA when the picture tube is cut off.

The heater of the regulator valve is connected in the series heater chain. Since its cathode warm-up time is shorter than that of the e.h.t. rectifier, the regulation circuit is ready for operation as soon as the e.h.t. becomes available.

Conventional voltage-dependent-resistor feedback stabilization of the line output stage is effected by the E298/ZZ06 in series with a 1-MΩ preset variable resistor across the grid circuit of the PL505, and fed from point 7 on the line output transformer.

The circuit detailed in Fig. 7 results in an e.h.t. line regulation of better than 5% for mains variation of 10% and a beam current load regulation up to 1.2mA better than 0.5%. What this means in simple terms is that an e.h.t. set at a nominal 25 kV at nominal mains voltage will not rise above 26 kV or fall below 24 kV for ±10% variation in mains voltage. Also there is no material change in picture height or width between zero and 1.2mA total beam current.

The 2-μF capacitor connected to point 8 on the output transformer serves a special purpose. To overcome the considerable difficulties of achieving satisfactory third-harmonic tuning and of reaching the necessary e.h.t. of 25 kV with 90° deflection, this Mullard design employs a special transformer using a large air gap, which ordinarily would imply a large VA requirement. To get around this, a "desaturation" technique is used by which the direct current from the PY500 booster diode to the PL505 pentode does not pass through any winding on the transformer, but through a separate choke, *L*. A.C. from the booster diode reaches the line output transformer through the 2-μF capacitor.

The focus supply voltage in Fig. 7 is obtained by rectifying and smoothing the high-voltage flyback pulses at point 8 on the line output transformer, which, as noted above, is effectively at the cathode of the PY500 boost diode. The rectifier, EY51, draws its heater current from the isolated winding 14-15 on the line output tranformer. The 500-kΩ preset potentiometer between points 1 and 5 on the line output transformer varies the amplitude of the flyback pulses applied at point "Y" to the 270-pF reservoir capacitor, and permits presetting of the resultant focus supply voltage level.

The boost voltage in the circuit of Fig. 7 is obtained from point 6 on the line output transformer, being rectified by the PY500 boost diode. This boost line has flyback pulses of about 700V superimposed on the 800V d.c. value, and the control supply to the grid of the PD500 regulator valve grid is smoothed by the 500 kΩ preset and the 0.33 μF to the 250-V decoupled h.t. rail.

FUTURE DEVELOPMENTS

For maximum usefulness this article has been written round current designs of hybrid colour sets as they are likely to come into use for the start of colour transmission in the United Kingdom, but, as semiconductors displace valves, fully transistorized receivers must take over in the end. Readers interested in how development trends are likely to affect the power supplies in colour sets should consult "A Development 15-inch Transistorized Color Receiver" by W. E. Babcock of R.C.A. in *I.E.E.E. Transactions on Broadcast and Television Receivers,"* July, 1966, Vol. BTR-12, No. 3, pp. 127-140.

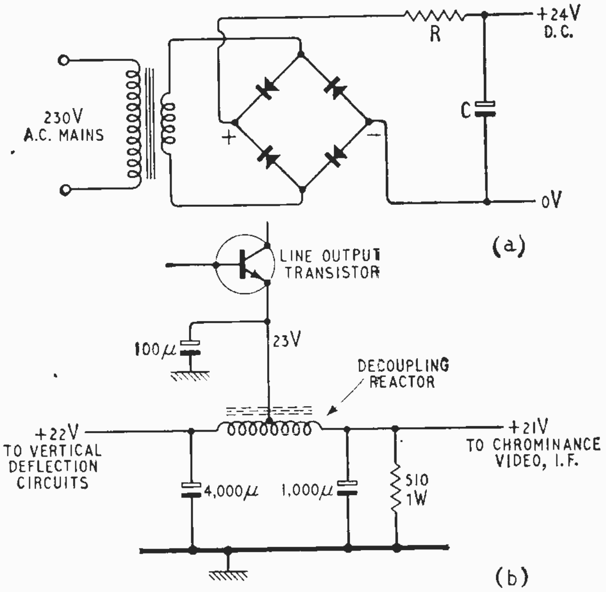


Fig. 6. Obtaining low-voltage, 24-V d.c. supply for transistor circuits in hybrid (transistor-cum-valve) colour receiver: (a) by mains step-down transformer, and (b) by decoupling reactor in line output transistor emitter (RCA circuit).

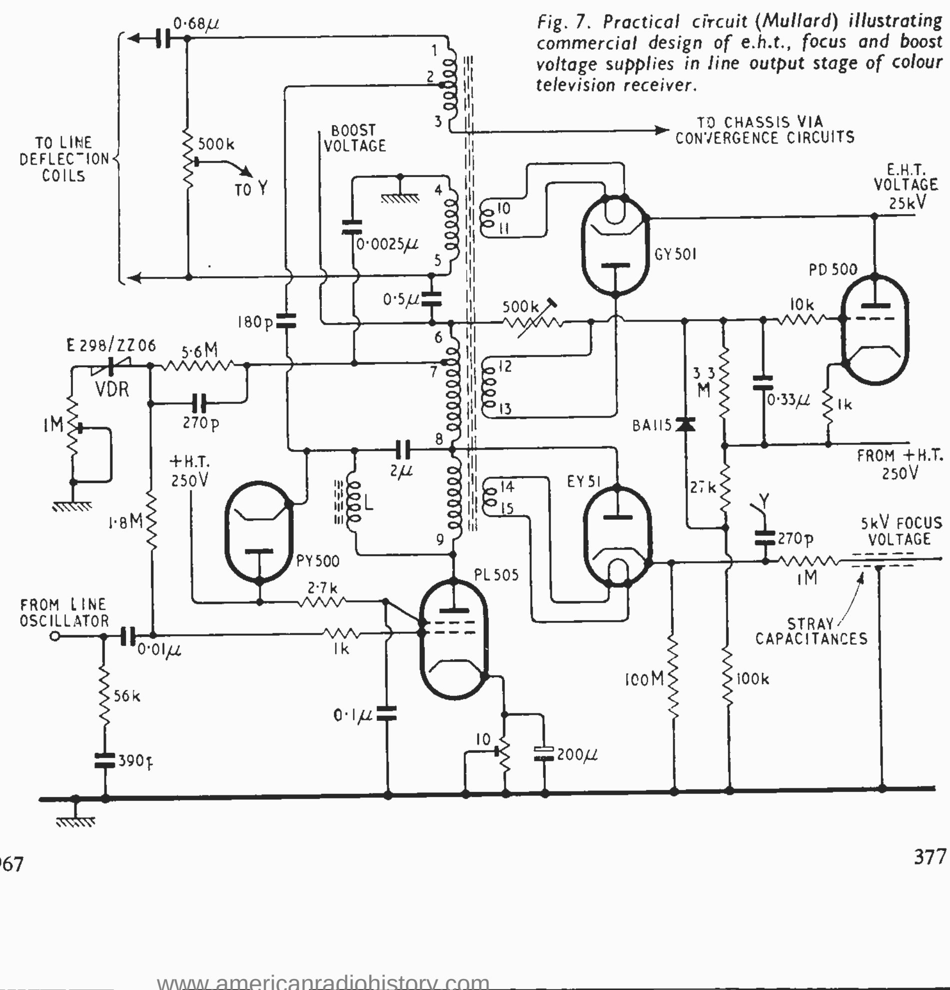


Fig. 7. Practical circuit (Mullard) illustrating commercial design of e.h.t., focus and boost voltage supplies in line output stage of colour television receiver.

WORLD OF WIRELESS

Three-Tube Cameras for Colour TV O.Bs

THE first advertised public transmissions of colour television, on 1st July, were B.B.C. outside broadcasts from the Wimbledon Lawn Tennis Championships. This was the beginning of what the B.B.C. are calling a "colour launching period"—a trial run for both the B.B.C. and the colour set users—before the official start of colour television on BBC-2 on 2nd December. The Wimbledon O.B. took place from a new mobile control room, the first of three to be equipped by the Corporation's Studio Planning and Installation Department during 1967/8. It is designed to operate with four cameras, and in this O.B. they were Peto-Scott 3-tube Plumbicon types. (Four-tube cameras will be used in later operations.) These 3-tube cameras are very much based on Dutch Philips technology, and use a Philips dichroic prism for colour splitting, but the maker's head amplifiers had been replaced by new f.e.t. ones, designed and built by the B.B.C. Research Department, in order to provide higher gain for conditions of low scene illumination. Zoom lenses with a range of 10:1 were fitted.

At a preview at Wimbledon, *Wireless World's* reporter was greatly impressed by the quality of the 625-line colour pictures seen on B.B.C. monitors, and the only minor faults were slight smearing on movement and the occasional appearance of thin colour fringes at the edges of the male players' white shirts. The 625-line compatible monochrome pictures—produced in this case by the combination of R, G and B signals (not from a separate luminance source as in a four-tube camera)—were extremely good, and there is no doubt that much of this quality must be attributed to the Philips "contours-out-of-green" electronic system of aperture correction used in the cameras. This system gets its name from the fact that the correction is performed initially only on the green pick-up tube signal but the correction signal so obtained is then added to the red and blue signals—an arrangement which avoids the need for separate aperture correctors for R, G and B signals. As a result the sharpness of the compatible monochrome picture is much improved. At a recent I.E.E. conference, C. B. B. Wood, of the B.B.C., said that the technique "goes quite a long way towards giving the three-tube camera the advantage of the four-tube camera in its resistance to loss of picture sharpness in case of misregistration."

The monochrome picture, however, was receivable only by BBC-2 viewers. The BBC-1 Wimbledon outside broadcasts came from a separate group of monochrome cameras and mobile control rooms.

More BBC-2 Transmitters to be Built

THE Postmaster General has approved in principle the following further group of high-power u.h.f. transmitting stations for BBC-2:—

Carmarthen	Ayrshire
Blaen-Plwyf (Cardigan)	North Kent
Dorset	Stockland Hill (East Devon)
Rosemarkie (Inverness)	Presely (Pembroke)
Caldbeck (Cumberland)	South Devon
West Cornwall	Arfon (Caernarvon)
Selkirk	West Sussex
Herefordshire	Dunbarton

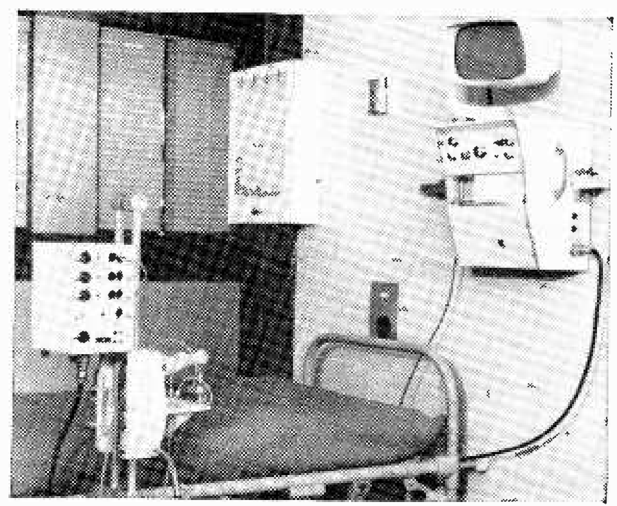
These 16 additional stations together with the 28 high-power stations already approved will make BBC-2 available to 90% of the population of the United Kingdom.

Twelve high-power stations are already in service. The station at Durris in North-east Scotland opens on July 29th and that at Tacolneston, near Norwich, will be brought into operation later this summer. A further two are expected to be ready for service by the end of this year, bringing the population coverage to about 70%. It is hoped to complete by early 1969, the remainder of the 28 stations previously approved, and the new group of 16 stations at the rate of three or four each year from 1970 onwards.

In addition to these high-power stations a number of low-power relay stations are being built to fill in the gaps in the coverage of the main stations.

Hospital-Built Intensive Care Monitoring Unit

AN intensive care monitoring system, which displays, measures, and records up to three blood pressure waveforms, and an e.c.g. has been developed by the medical electronics department of St. Thomas' Hospital, London. The criteria for the design were minimum space occupied at the bedside, use of i.cs where possible, and the employment of strain gauge pressure transducers. The decision to use these transducers led to the development of a simple, compact pressure signal pre-amplifier using integrated circuit amplifiers. Calibration of this unit is achieved by the automatic generation of a sequence of simulated pressure levels appropriate to the range in use. A check on system linearity, and unambiguous range identification in strip chart recordings are also possible with this calibration system. A parametric amplifier is used in the e.c.g. channel, and the fully isolated input circuit provides common mode rejection, and immunity from damage by transient high voltage inputs which can occur when a d.c. defibrillator is used. Two, three or four signals can be displayed simultaneously on the multi-trace c.r.t. and a stationary display of waveforms is possible.



Right: Intensive care equipment installed at St. Thomas' Hospital.

Domestic Newspaper by Radio

EVALUATION of an experimental system that will transmit news copy along with a normal television programme, and print it out in the living room, is to be carried out by the Radio Corporation of America. Permission to make actual on-the-air tests between New York and Princeton, New Jersey, has been requested of the Federal Communications Commission. Dr. Hillier of RCA Laboratories said that this system will require no additional allocation in the r.f. spectrum, nor will it affect television services or limit them in any way. "Printed copy is converted into a series of electromagnetic signals," which are inserted into the transmission during the television vertical blanking intervals (occurring 60 times a second). The present design permits simultaneous transmission of four different printed messages. Experimental printers in current use will produce the equivalent of a page from a standard paperback book every ten seconds by employing an RCA electrostatic printing process. This system which is said to be compatible with v.h.f. or u.h.f. transmissions will be tested over a six months period. Suggested services obtainable through this system—apart from the obvious ones of news briefs, sports scores, and stock market reports—are charts, cartoon strips, maps, television programme schedules, syndicated columns and news magazines. However even when proven it is stated that it might be a few years before this domestic facsimile system becomes generally available.

I.E.R.E. Membership Structure

THE Institution of Electronic and Radio Engineers intends to seek permission to amend its Bye-Laws so that existing Members can be regarded as Fellows and Associate Members as Members. This is in line with current thinking among the constituent members of the Council of Engineering Institutions towards standardizing their corporate membership designations.

It is also intended that the grade of Associate, which at present is limited to those persons who have the requisite experience but not the formal academic qualifications for corporate membership, will be broadened to include holders of the H.N.C., or the City and Guilds Full Technological Certificate in Telecommunications. The purpose of this change is to enable senior technicians to be associated with, and to participate in, the same Institutional activities as the professional engineers with whom they work so closely.

The Council of the Institution hopes that these changes will become effective before the end of this year.

At the time of going to press 22 manufacturers of broadcasting and ancillary equipment had taken space at the exhibition to be held in conjunction with the **International Broadcasting Convention** at the Royal Lancaster Hotel, London, from September 20th-22nd. Details of the Convention, which is being sponsored by the Electronic Engineering Association and the Royal Television Society, are obtainable from 166 Shaftesbury Avenue, London, W.C.2.

Station Engineers have been appointed by the B.B.C. for the first six of the nine stations to be built for the experimental **local broadcasting service.** They are Brighton, E. R. H. Castle; Leicester, J. B. Hawley; Merseyside, C. G. Wright; Nottingham, M. R. A. Edis; Sheffield, J. K. Beard; and Stoke-on-Trent, B. G. Lock. They will be responsible for installing, testing and commissioning the equipment for the local stations and for its subsequent operation and maintenance. Stations are also to be set up at Manchester, Leeds and Durham; the last two were announced by the P.M.G. on July 4th.

The author of the article **Design of Negative-feedback Equalizers** in the July issue has asked us to correct three errors in his text. A root sign was omitted from the denominator in eq. 3; for 220 kΩ in line 8, col. 1, p. 349 read 100 kΩ; and for Fig. 6 in the 9th line from the foot of col. 2 read Fig. 10.

In addition to those **domestic television and radio receiver** manufacturers listed in our June issue (p. 282) the following are also holding London trade shows:—

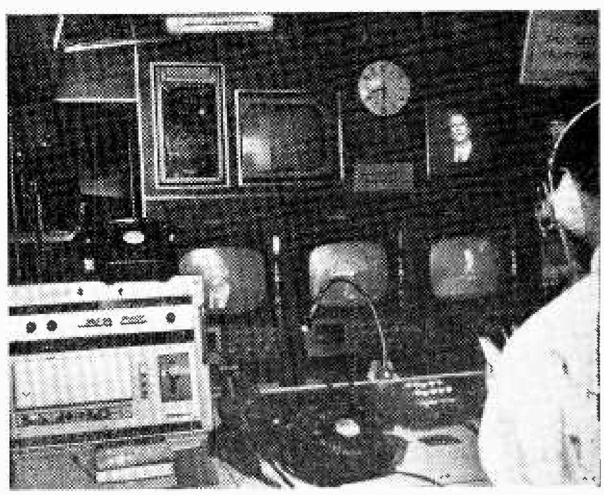
Baird: Seymour Mews House, Wigmore Street, W.1, 20-24 August.
Bang & Olufsen: 70/71 Welbeck Street, W.1, 21-25 Aug.
Bosch: Kensington Palace Hotel, W.8, 22-24 August.
National Radio: Playboy Club, Park Lane, W.1, 21-24 August.

Electronics E.D.C.—Four new additional members have recently been appointed to the Economic Development Committee for the Electronics Industry, bringing the total to 20 plus Sir Donald Stokes, the chairman. The new members are Viscount Caldecote (deputy managing director, British Aircraft Corp.), Dr. F. E. Jones (managing director, Mullard), A. D. Mackay (managing director, S.T.C.) and Professor G. D. Sims (Southampton University).

German Colour Television.—Both the German television organizations mentioned in our note in the July issue (p. 324) on the introduction of colour in Europe are, of course, in the Federal Republic. The A.R.D. (*Arbeitsgemeinschaft der Oeffentlich-Rechtlichen Rundfunkanstalten der Bundesrepublik Deutschland*) to which the broadcasting organizations serving the different *laender* and Berlin belong, operates the first and third television programmes, and Z.D.F. (*Zweites Deutsches Fernsehen*) the second (commercial) programme network. Colour is being introduced in both the first and third programmes on August 25th.

Colour Tube Guarantee.—"ColourScreen" television tubes (25 in) by Mullard are to be covered by a guarantee of one year, subject to the same conditions applicable to monochrome tubes. There will be an option of extending the guarantee for a further three years by paying an £8 premium.

Non-Destructive Testing is the title of a new quarterly journal to be published by Iliffe Science and Technology Publications Ltd. It will be concerned with new techniques, and economic and management aspects of non-destructive testing. The annual subscription will be £6 (£10 overseas). The first issue will appear on August 17th.



From this Marconi mobile control room relays (via Post Office networks) from Earls Court, London, to 25 centres, each equipped with Eidophor receivers, enabled some 100,000 people to participate each evening in the recent Billy Graham crusade. The most distant centres were Aberdeen and Belfast. Television Advisors Ltd. provided the technical facilities.

PERSONALITIES

Sebastian de Ferranti, chairman and managing director of Ferranti Ltd., has received an honorary degree of Doctor of Science of the University of Salford. The degree was conferred upon him by His Royal Highness the Duke of Edinburgh, who is the first Chancellor of the University.

J. Sharpe, B.Sc., F.I.E.E., has been appointed commercial director of E.M.I. Electronics Ltd. He joined E.M.I. in 1953 as chief engineer of the Valve Division, of which he later became sales manager. For the past year he has been general manager of the Commercial Division. Before joining E.M.I. he spent six years at A.E.R.E., Harwell, in charge of a group dealing with nuclear radiation detectors, having previously been eleven years at the G.E.C. Research Laboratories working on cathode-ray tubes.

J. Sharpe

G. D. Cook, M.I.E.E., has become assistant chief engineer, B.B.C. Television Operations in succession to **W. R. Fletcher,** B.Sc(Eng.), M.I.E.E., who has resigned from the Corporation. Mr. Cook joined the B.B.C. in 1947 as a maintenance engineer at Brookmans Park transmitting station and in 1955 transferred to the Television Service when he became assistant to the superintendent engineer, television (regions and outside broadcasts). He was head of engineering (Wales) from 1962 until earlier this year when he was appointed assistant chief engineer, television developments.

F. G. Sandman, director in charge of production and technical developments with Teleng Ltd., of Romford, Essex, for the past five years, has been appointed managing director. He succeeds **Jack Evans,** who is undertaking other responsibilities with the Telefusion organization of which Teleng, manufacturers of v.h.f. communal aerial systems, is a subsidiary.

Peter E. Trier, M.A., M.I.E.E., M.I.E.R.E., director of the Mullard Research Laboratories, Salfords, Surrey, is to give the 1968/69 I.E.E. Faraday Lecture which will be concerned with microelectronics. Mr. Trier, who is 47 and graduated as a wrangler in the mathematical tripos at Cambridge, spent nine years at the Admiralty Signal and Research Establishment before joining the Mullard Laboratories in 1950 as head of the communications and radar division. He was appointed manager of the laboratories in 1953 and became director in 1957. He has been on the board of Mullard Ltd. for the past five years.

P. E. Trier

Cornelius R. Webster, who was until recently with Visual Electronics International as European Manager, has been appointed managing director of the Crow Co. of Reading. The company has recently undertaken European representation for Richmond Hill Laboratories of Canada (instruments and control equipment); Sparta (cartridge tape recorders) and MVR Corporation (Videodisc recorders). Mr. Webster was at one time with Southern Television prior to which he was for three years senior video engineer of Ampex Great Britain.

Claude A. Lucas, A.M.I.R.E.E. (Aust.), who recently came to this country from his native Australia, has been appointed a sales engineer with English Electric Valve Company. He will be concerned chiefly with c.r. tubes and storage tubes. He served for 17 years with Philips Electrical (Australia), latterly as product manager for the professional tube division.

B. France, B.Sc.(Eng.), M.I.E.E., is the new chief engineer of Scotland Yard in succession to **W. M. S. Cawley,** C.B.E., B.Sc.(Eng.), who has retired.

G. Salter, A.M.I.E.R.E., is appointed by the B.B.C. as Head of Engineering, Wales, in succession to **G. D. Cook,** M.I.E.E., who has been appointed assistant chief engineer, television developments, London. Mr. Salter joined the B.B.C. in 1943 at the Wrexham transmitting station. Later he served in the Sound Service in Manchester and London and in 1950 transferred to the Television Service. In 1965 he was appointed assistant superintendent engineer, television recording, and later the same year became projects engineer, television, with responsibility for the operational planning of new television installations.

English Electric Computers have appointed **A. H. Headley** as manager of their Design Services and Automation Organisation which has its headquarters at Kidsgrove, Staffs. In 1959, Mr. Headley joined Leo Computers, which is now integrated into English Electric, and led the teams which developed the Leo 326 and System 4-50 computers. Before his present appointment he was head of the London Development Department of English Electric.

John Glaser was recently appointed sales director of J. Beam Aerials Ltd., of Northampton. He was formerly sales director of Antiference Ltd., from whom he resigned after nearly 20 years' service.

M. J. Carus-Wilson, B.Sc., has joined S.T.C. Semiconductors Ltd. as field sales manager at its Footscray, Kent, plant. For the past three years he has been responsible for sales of all Mullard products to the U.K. telephone industry. Mr. Carus-Wilson, who is 30, graduated from Trinity College, Cambridge, with an honours degree in metallurgy, and then spent six years on electronic circuit and systems development first with E.M.I. and later with Decca on radar display systems.

M. J. Carus-Wilson

C. T. Chapman, M.I.E.R.E., managing director of Derritron Electronic Vibrators Ltd. has been appointed to the main board of Derritron Ltd., the parent company of the Group. In 1950 Mr. Chapman formed his own company, C. T. Chapman (Reproducers) Ltd., which was acquired by Derritron in 1960.

C. T. Chapman

Dr. J. Ross Macdonald, director of Corporate Research and Engineering at Texas Instruments Inc., Dallas, Texas, has received his second doctorate from Oxford University. In 1950 he obtained a D.Phil. in Natural Philosophy (Physics) after attending Oxford as a Rhodes Scholar and in June received a D.Sc., in recognition of his work in a wide variety of areas of physics and electronics including dielectric response, linear systems, space charge, electrolyte double layers, and electrical effects of absorption. In 1944 he obtained a B.A. in physics from Williams College, and a B.S. in electrical engineering from M.I.T. where he taught at the Technical Radar School. He served in the U.S. Naval Reserve from 1944 to 1946 as a technical radar-radio officer, after which he returned to M.I.T. for postgraduate work. After receiving his 1950 doctorate he conducted research in experimental and theoretical physics at Armour Research Foundation. On leave of absence from Armour he served for a year as associate physicist at Argonne National Laboratory working on solid state physics problems. Dr. Macdonald joined Texas Instruments in 1953. He served as director of solid state research and as director of the Physics Research Laboratory from 1955 until 1963 when he became director of T.I.'s Central Research Laboratories. Since January, he has been director of Corporate Research & Engineering.

James J. Pinto has resigned his position as managing and technical director of K.P.E. Controls Ltd. of which he was founder member, and formed a new company—Process Auto-Control Electronics Ltd.—for design and development consultancy in the fields of semiconductor instrumentation and process automation.

K. J. Austin, M.A., M.I.E.E., is appointed head of the technical group in the B.B.C. Equipment Department. Mr. Austin joined the B.B.C's Planning and Installation Department in 1951 and since 1964 has been head of the general services section of the Designs Department. He succeeds **H. D. M. Ellis,** M.B.E., M.A., F.I.E.E., who has retired after 30 years' service with the Corporation. Mr. Ellis started in the Research Department but four years later transferred to the then, Station Design and Installation Department. Since 1955 he has been head of the technical group, Equipment Department.

E. J. Reid, B.Sc., A.C.G.I., M.I.E.E., recently became managing director of Data Recording Instrument Co. (a subsidiary of International Computers and Tabulators Ltd.) in succession to **R. E. Hutchins** who has taken up a new appointment as manager of I.C.T.'s Peripheral Equipment Organisation but remains a director of D.R.I. Mr. Reid, who is 46, is an honours graduate in electrical engineering from the City and Guilds College, London, and prior to joining D.R.I. was with the General Electric Company.

Dynamco Ltd., who recently announced that they were to establish engineering and assembly facilities in North Sydney, Australia, during 1967 with D. E. J. Rawlings as managing director have now appointed **M. F. Marks,** M.I.E.E., as technical director of the Australian subsidiary. Mr. Marks was previously with Elliott Automation where he was latterly responsible for the "Hydroplot" shipborne data logging system.

THE QUEEN'S BIRTHDAY HONOURS

RECIPIENTS of honours on the occasion of H.M. The Queen's Birthday included:-

Knight Bachelor

Cecil Mead, lately chairman, International Computers and Tabulators Ltd.

K.C.M.G.

Dawson Donaldson, chairman, Commonwealth Telecommunications Board.

Leonard James Hooper, C.B.E., director, Government Communications H.Q.

C.B.

W. H. Penley, C.B.E., deputy controller of electronics, Ministry of Technology.

C.M.G.

A. G. Touch, Government Communications Headquarters.

C.B.E.

R. H. C. Foxwell, chairman, Wayne Kerr Company.

W. J. Sharpe, O.B.E., director of communications, Foreign Office.

R. Telford, F.I.E.E., managing director, Marconi Company.

Howard Thomas, managing director, A.B.C., Television.

E. V. Truefitt, director of scientific research (electronics), Ministry of Technology.

O.B.E.

Commander A. C. S. Gower, D.S.M., A.M.I.E.R.E., Royal Navy (Retd.).

A. Lightbody, technical executive, Hawker Siddeley Dynamics Ltd.

R. E. Reason, research manager, Rank Taylor Hobson, Leicester.

M.B.E.

R. H. Bradley, chief engineer, Television Wales and the West.

H. A. Edwards, lately divisional chief inspector, Systems Group, Plessey Co.

A. H. Ellson, B.Sc., M.I.E.E., divisional manager, Mullard Equipment Ltd.

A. E. Evans, chairman and managing director, Evans Electroselenium Ltd.

J. R. Hawkins, chief radio officer, s.s. *Canberra,* Peninsular and Oriental Steam Navigation Company.

L. H. Robinson, B.E.M., lately Technical Officer Grade I, Telecommunications Engineering Establishment, Board of Trade.

S. J. Sellwood, executive engineer, Radio Planning and Provision Branch, G.P.O.

G. E. Tagholm, engineer-in-charge, Croydon transmitting station, I.T.A.

E. C. Walton, head of Department of Electrical Engineering and Physics, Leeds College of Technology.

B.E.M.

F. T. Fenner, instrument maker/mechanic, Standard Telecommunications Laboratories Ltd.

V. E. Massen, supervisor instrument maker, Independent Television Authority.

G. M. Tranter, chief model maker, Whiteley Electrical Radio Co.

R. H. C. Foxwell (C.B.E.)

R. Telford (C.B.E.)

A. H. Ellson (M.B.E.)

A Variable Voltage Stabilized Power Supply

By ALISTAIR SHARP,* B.Sc.

AN essential requirement for an experimenter working with transistor circuits is a variable voltage, stabilized power supply capable of delivering high currents. This article describes a stabilized supply the output of which could be varied from 6 V to 20 V and which would deliver a minimum current of 400 mA at any voltage. It also discusses the procedure used in designing it, thus making it possible to modify the design to individual requirements. Both series and shunt stabilizers were considered, and although the shunt stabilizer always draws maximum power, it was felt that this circuit with its inherent over-current protection, and using fewer semiconductors, was a more practical proposition.

The circuit of the prototype is shown in Fig. 1. The resistor chain R1, RV1, R2 serves a dual purpose. First, by means of RV1, the output voltage can be controlled. Secondly, once the output voltage has been set, it senses errors in the voltage, and controls Tr1 accordingly. Tr1 thus provides an error signal, and this is amplified in Tr2. The output of Tr2 controls the collector current of Tr3. If the output voltage rises, the circuit increases the collector current in Tr3. This increases the voltage drop across R3, thus reducing the output voltage. If the output voltage falls, the reverse occurs. ZD1 should be chosen to have a Zener voltage considerably lower than the minimum voltage that the stabilizer is required to supply to ensure Tr1 does not cut off. For the prototype, the minimum voltage was 6 V, and ZD1 was a 3.3 V Zener diode. R4 ensures that the Zener always gets sufficient current to provide a good 'knee.'

Thus $R4 = V_{min}/I_{z\ min}$

where V_{min} is minimum output voltage, and $I_{z\ min}$ is minimum Zener current. Maximum current flows in R4 when the power supply is maximum, V_{max}.

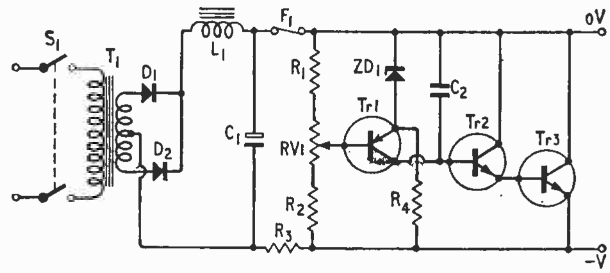
So maximum power in $R4 = (V_{max} - V_z)^2/R4$ where V_z is Zener voltage. R3 is found by taking the voltage drop across it when the supply is giving maximum voltage and maximum current,

i.e. $R3 = (V_t - V_{max})/I_{max}$

where V_t = unstabilized input voltage, and I_{max} = maximum current that the supply is required to deliver. The maximum dissipation in R3 is found from the equation,

Max. power in $R3 = V_t^2/R3$

The short circuit output current is fixed, since when a short circuit exists, R3 will drop the entire unstabilized input voltage. The resistor chain R1, RV1, R2 must draw sufficient current to provide thermal stabilization of Tr1 at under the minimum voltage. When this current is fixed, the total resistance of the chain is fixed. R1 should, at this current, drop the Zener voltage plus the base-emitter voltage drop of Tr1, to ensure that Tr1 does not cut off, thus R1 can be defined. RV1 should be about half the total resistance of the chain, so RV1 and R2 are fixed. C2 is present to prevent high frequency oscillation and should be about 0.1 μF. The transistors are chosen as follows: Tr3 must be capable of dissipating the entire power supplied by the stabilizer, so the space available for heat sink limits the choice of transistor. In the prototype, sufficient space was available to permit the use of a BUY11. The power dissipated in Tr2 is given by the power dissipated in Tr3 divided by the current gain of Tr3. The minimum current gain of the transistor chosen for Tr3 must be used for this calculation. In the prototype, this was 400 mW, and a 2S712 was used. The dissipated power in Tr1 is unlikely to be critical, but may be determined from the power in Tr2 by the method just used for Tr2. Account must also be taken of collector currents and voltages, and care should be taken that the manufacturer's recommendations are not exceeded. One p-n-p and two n-p-n transistors were used in the prototype, but this can be altered to one n-p-n and two p-n-ps if desired, by simply inverting the rail voltages. The diodes, D1 and D2, must, of course, be capable of passing the short circuit current of the supply.



Test Results on Prototype				
Voltage setting on no load	O/P at 100 mA	O/P at 200 mA	O/P at 300 mA	O/P at 400 mA
20 V	19.8 V	19.6 V	19.0 V	17.7 V
15 V	15.0 V	15.0 V	14.9 V	14.6 V
10 V	10.0 V	10.0 V	10.0 V	9.9 V
6 V	6.0 V	6.0 V	6.0 V	6.0 V

The apparently bad stabilization at an output voltage setting of 20 V and high load current is that, of the 400-mA current allowed at this voltage setting, a large fraction is being used by the Zener diode, and so is not available as output. Ripple voltage at 10 V output on 400 mA load = 40 mV. The short circuit output current is 1.2 A. When 200 mA was drawn at 10 V for 30 minutes, the voltage fell by 0.1 V and the current by 6 mA.

COMPONENTS LIST

Resistors
R1 = 300Ω 0.1W
R2 = 510Ω 0.1W
R3 = 27.5Ω 40W
R4 = 66Ω 6W
RV1 = 1kΩ 0.25W, w/w

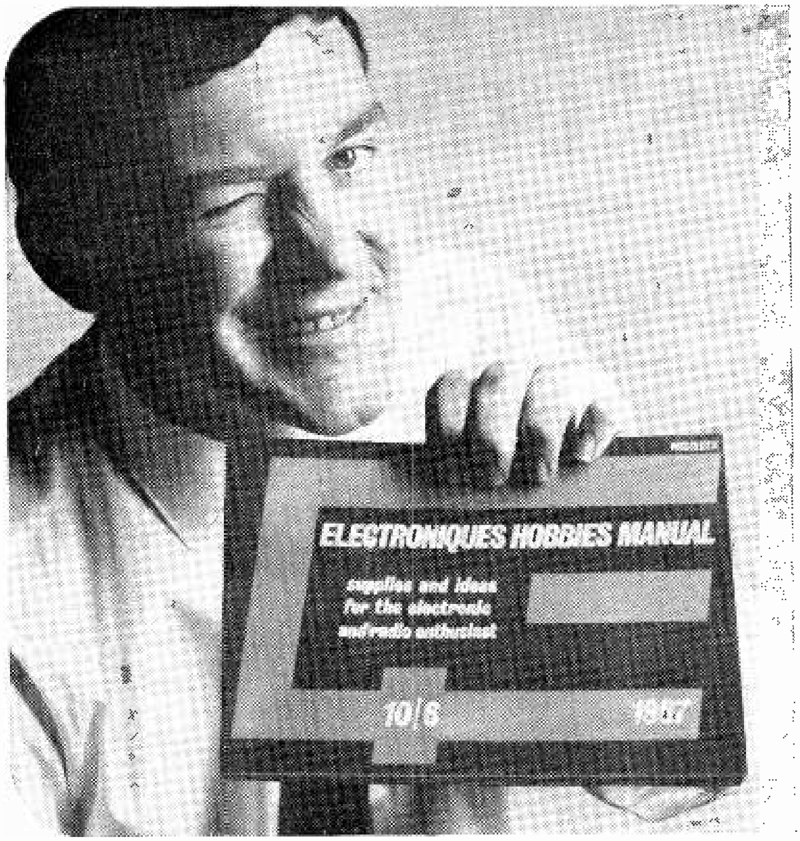
Capacitors
C1 = 1000 μF, 50 V
C2 = 0.1 μF

Inductor
L1 = 1 mH, 2A

Semiconductors
ZD1 = Z2A 33F
D1, D2 = RAS 310 AF
Tr1 = 2S302
Tr2 = 2S712
Tr3 = BUY11

Radiospares 30V rectifier transformer, using 25 V secondaries; DPST mains switch and 1.5 A fuse.

* Standard Telephones & Cables Ltd., Basildon, Essex.



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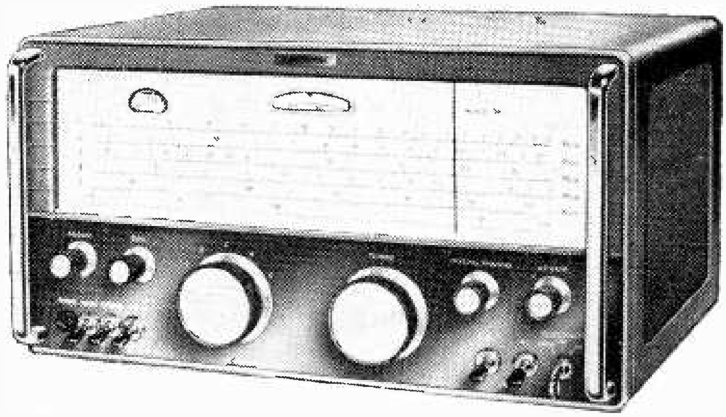

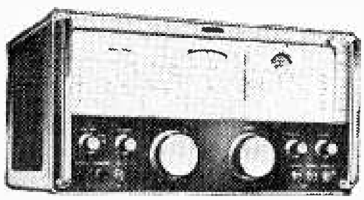

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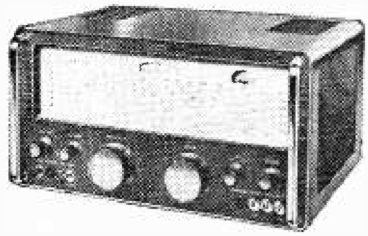

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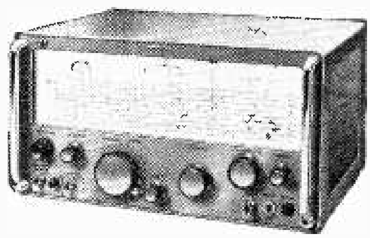

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Low-Distortion RC Oscillator

Silicon transistor design covering 1.5 Hz to 150 kHz

By P. F. RIDLER, B.E., M.I.E.E.

IN the oscillator described the requirements of reasonable output voltage, wide frequency coverage, low distortion and good stability are met. The performance is summarized in the panel on p. 386.

The general procedure followed is to first design a very low distortion amplifier and then, by using positive feedback through a frequency-selective network, make the amplifier oscillate at a frequency which is determined by the feedback network. It is absolutely necessary that the amplifier have extremely low distortion, for the distortion of the ouput waveform of the oscillator will always be greater than the distortion of the amplifier from which it is made. (Ref. 1)

Starting the design of the amplifier from the output end and referring to Fig. 1, the total load seen by the output stage is the parallel combination of the 600 Ω load, the feedback networks and the 1,000 Ω output level control; these total around 300 Ω. For an output voltage of 2·2 V r.m.s. the peak voltage will be $\sqrt{2} \times 2{\cdot}2 = 3{\cdot}1$ V, so that the peak current which must be supplied by the output stage is 3·1 V/300 Ω ≈ 10 mA. If the output stage is to operate in class A in order to have low distortion, the direct current through it must be equal to or greater than the peak alternating current, so that this sets a lower limit to the drain from the supply battery, although twice this current was used to ensure low distortion.

An emitter follower is chosen as the output stage for the dual purpose of providing low output impedance and low distortion, and it can be seen that the d.c. load in the emitter of this transistor Tr3 is another transistor Tr4. A resistor could have been used here, but as this would also have been in parallel with the load, it would have raised further the peak alternating current to be supplied by the output stage and thus the drain on the supply battery. If we remember that the collector impedance of a transistor is very high, it is obvious that this can be used as the tail resistor of the emitter follower, without requiring the extra supply voltage which would be necessary if a high value passive resistor were to be used for this purpose. The impedance of Tr4 at its collector is about 100 kΩ and there is 20 mA passing through it with a voltage drop of 13 V: an ordinary resistor of this value would require a supply voltage of 20 mA × 100 kΩ = 2,000 V! A voltage divider R_2, R_3 supplies the base of Tr4 with 3·0 V from the supply line and as the direct voltage at the emitter of Tr4 will be almost exactly 0·6 V less than this for a silicon transistor, a resistor between here and earth will define the current through itself, Tr4 and Tr3. Quite obviously, this resistor R_1 must have a value of 2·4 V/20 mA = 120 Ω, fixing the current at the value which we have calculated previously to be necessary.

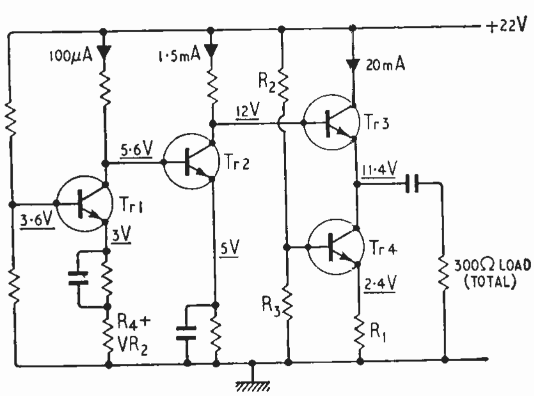


Fig. 1. Basic circuit of amplifier section.

The voltage conditions in the output stage are not critical as the distortion is almost unmeasurable unless clipping occurs, so that as long as the junction of Tr3 emitter and Tr4 collector is at about half the supply potential, all will be well: this also equalizes the dissipations in Tr3 and Tr4 and keeps each comfortably within the ratings of the transistors specified.

The driver stage has to supply an alternating current to the base of Tr3 and as the voltage gain of an emitter follower is unity, the voltage at which this current is to be supplied into the input impedance of Tr3 is fixed. The input impedance of Tr3 is approximately its current gain h_{fe} multiplied by its load resistance of 300 Ω. This could be as low as 30 kΩ for a low limit transistor, although it is more likely to be about twice this figure. Taking the low figure for the sake of safety, the peak signal current to be produced by the driver transistor Tr2 will be 0.2 mA together with the signal current flowing in its load resistor. At a reasonable estimate, a direct current of 1.5 mA through Tr2 should give sufficient signal current capability and will give a fairly high value of current gain, but for this stage we must be very careful about distortion as there is no local feedback as there was in the output stage. Reference 2 shows that for any collector current there is a definite value of collector load resistance which minimizes the distortion of the stage. Fortunately this distortion minimum is fairly broad as the load is varied and the distortion is not critically dependent on the operating conditions. For the 2N3707 specified, operating at a current of 1.5 mA, the optimum load resistance is about 5.2 kΩ and varies very little from sample to sample of transistor. The nearest standard value of resistor 5.6 kΩ is used, and this in combination with the 1.5 mA direct current gives a voltage drop of 10 V across the load and makes the collector voltage 12 V: this value of collector voltage also permits Tr2 collector to be directly coupled to Tr3 base as it is roughly half the supply voltage. If the emitter of Tr2 is held at 5 V this will give plenty of room for the 6.2 V peak-to-peak swing at the collector, and the emitter resistor must be 5 V/1.5 mA = 3.3 kΩ.

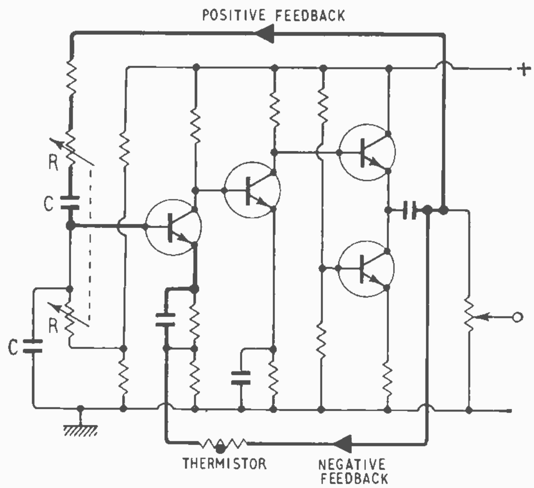


Fig. 2. Basic circuit of oscillator showing feedback loops.

Bearing in mind that any unwanted output, and not just harmonics of the required frequency, can be classed as distortion, we must ensure that the noise at the output of our amplifier is negligible compared with the oscillatory voltage, and as the first stage normally determines the noise output of the whole amplifier we must choose a low-noise transistor for the first stage and operate it under optimum conditions. Fortunately, the transistor manufacturer usually recommends suitable conditions for low-noise operation and for the device used these are a collector current of 100 μA and a collector to emitter voltage of 5 V. The first stage is run at a current of 500 μA in order to give a slightly higher gain than would have been obtained at 100 μA although this makes very little difference to the noise.

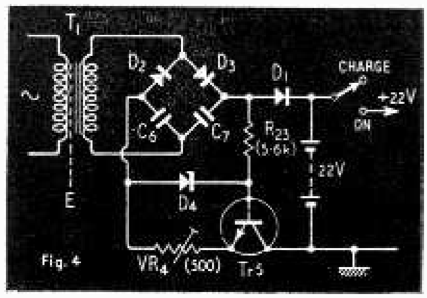
The voltage across the first stage load is the potential difference between the supply rail and the base of Tr2, i.e. $22-5.5=16.5$ V, and the current through this resistor is 500 μA, giving a resistance of 33 kΩ. The emitter of this stage should be 5 V below the 5.5 V at its collector but this would make its emitter resistor very small and stability would suffer. Accordingly the collector-emitter potential is reduced to 3 V which has very little effect on anything except the output voltage capability; as the output voltage is only 20 mV, this is hardly important. The emitter of Tr1 is now at 2.5 V and the emitter resistor should therefore be 5 kΩ, but as part of this is incorporated in the feedback resistors R_1 and VR_2, 4.3 kΩ is used.

The reactance of the bypass capacitors should be low compared with the reciprocal of the g_m of the transistor they bypass. A 100 μF capacitor has a reactance of 1 kΩ at 1.5 Hz and the g_m of a transistor at 1 mA collector current is about 40 m℧ ($1/g_m=25\Omega$) so it can be seen that the bypass capacitors would have to be very large indeed. A bit of cut and try shows that 100 μF capacitors do not degrade the performance excessively and do make the instrument more compact.

FEEDBACK LOOPS

The amplifier as it stands (i.e. without feedback) has a gain of about 3000 and a distortion of approximately $\frac{1}{2}$% when delivering its rated output. The criterion for oscillation, of course, is that the gain around a feedback loop is unity at the same time as the total phase shift is zero. This means that we have to attach to our amplifier some sort of feedback network which has a phase shift of zero

RANGE	1	2	3	4	5
FREQUENCY (Hz)	1·5 – 15	15 – 150	150 – 1·5k	1·5k – 15k	15k – 150k
CAPACITANCE C	2μ	0·2μ	0·02μ	2,000p	200p (170p IN PARALLEL WITH 3-30p)



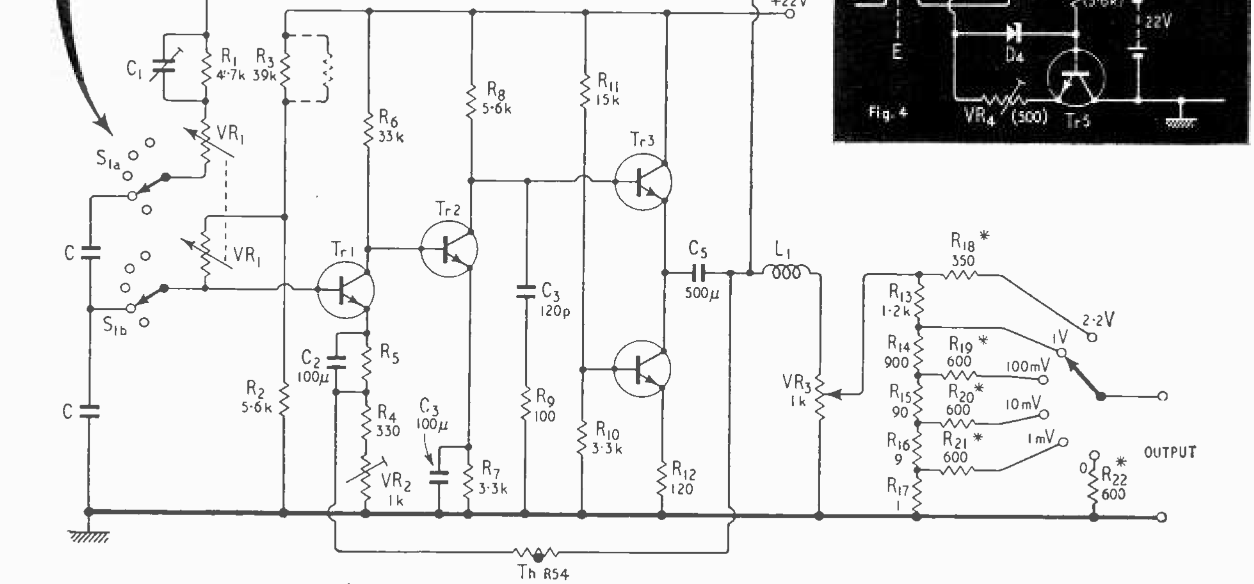


Fig. 3. Complete oscillator circuit with power supply circuit inset (Fig. 4). Note that Ni-Cd battery can be trickle charged by leaving connected to power supply when oscillator is switched off. R_5 should be 4·3kΩ.

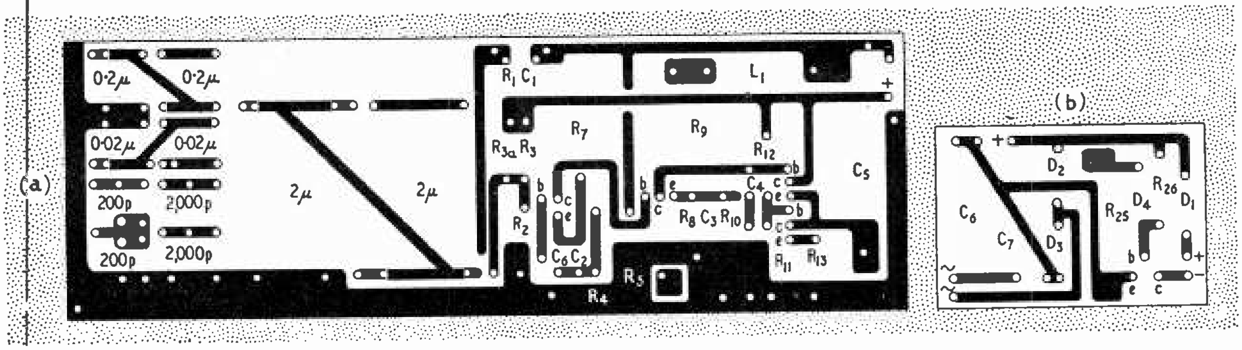


Fig. 5. Suitable printed board layouts. The smaller is the power supply board.

occurring at only one particular frequency. There are a number of networks which will satisfy this condition, but the simplest, although not the best from the distortion point of view, is the Wien network which is simply the reactive side of the Wien bridge. The network has a phase shift of 0° at a frequency $f=1/2\pi RC$ and at this frequency it has a gain of one-third. The frequency for zero phase shift can be varied by changing either R or C, and the fairly obvious choice is to use a two-gang potentiometer to give variable frequency and switch the capacitors to change from one range to the next. By using a square-law resistance track in the potentiometers, the frequency dial can be made approximately logarithmic, giving more or less the same setting accuracy, percentage-wise, at all points on the dial. As the Wien network has a phase shift of zero we must connect it between two points on the amplifier which also have zero phase difference: this is easily arranged by putting it between the output and the first base. As a ten to one ratio of frequencies is required in each range it will be necessary to include some fixed resistance in series with each section of the variable ganged resistor and this series resistor should have a value of about 5 kΩ; the exact value does not matter very greatly but it should be a high stability type so that the eventual calibration will stay put. In the series arm of the Wien network this series resistor will be a single resistor, but for the shunt arm it can be made to fulfil a second purpose of furnishing the correct bias to the first stage of the amplifier. By making this resistor from two separate resistors in series and connecting the outer ends to earth and supply respectively, with the variable resistor attached to the midpoint, the correct bias is presented to the base of Tr1 and the minimum resistance of the shunt Wien arm is fixed simultaneously. As the bias at Tr1 base is to be 3 V, the ratio of the two resistors should be 3 : 19 and the resistance of the two in parallel must be 5 kΩ. This calculation produces the values of 5.6 kΩ and 39 kΩ but in practice this is all a bit hypothetical, as the easiest way of getting the voltages throughout the amplifier correct in the final setting-up is to adjust one of these resistors as outlined in a later paragraph.

The amplifier gain is extremely critical, for if the gain is fractionally less than three, oscillation will either not start or having started will die out again, and if the gain is more than three the output voltage will keep rising until one of the amplifier stages limits.

The gain can be stabilized if the negative feedback resistor is made to have a resistance which is dependent upon the voltage across it. One such type of resistor is the thermistor which has a resistance which falls very steeply as its temperature increases; indirectly, of course,

COMPONENTS

C	see table, Fig. 3.			
C_1	3-30 pF	air trimmer	Tr1, Tr2	2N3707
C_2	100 μF	6 V electrolytic	Tr3, Tr4	2N3704
C_3	100 μF	15 V „	Tr5	2N706
C_4	120 pF	silvered mica	D1-D3	1S130 or similar
C_5	500 μF	15 V electrolytic		
C_6, C_7	50 μF	25 V „	D4	6 V Zener diode
R_1-R_{12}		½ W, 5% high stability		
R_{13}-R_{17}		½ W, 1% „ „	L1	100 μH r.f. choke
R^*_{18}-R^*_{22}		½ W, 5% „ „	T1	13V secondary
VR_1	50 kΩ	2-gang square law wirewound (Colvern CLR 40/193/17)	Battery	22V Ni-Cd sealed accumulator
VR_2	1 kΩ	wirewound	Th	Thermistor type R54 (STC)
VR_3	1 kΩ	wirewound	S1	2-pole 5-way range switch
VR_4	500 Ω	wirewound	S2+	To suit requirements

*Can be omitted if 600 Ω attenuator is not required.
+Supply switching can be combined with attenuator switch if desired.

its resistance will depend very much on the voltage across it, for the higher the voltage the greater the power dissipated in the thermistor and the higher the temperature. The thermistor is connected from the output back to the emitter of the first amplifier stage, as shown in the simplified circuit of Fig. 2. The thermistor has a resistance of approximately 1.5 kΩ at the working voltage so that the unbypassed emitter resistor (R_4+VR_2) must be half of this to give a gain of three minutes. A combination of a fixed 330 Ω and a 1 kΩ variable will give the desired variation.

Several small capacitors for various purposes are included in the complete circuit (Fig. 3). The 120 pF capacitor in series with the 100 Ω resistor from collector of Tr2 to earth is to tailor the high frequency response of the amplifier so that the negative feedback loop will not oscillate at a high frequency. The trimmer across the 4.7 kΩ section of the Wien network allows the upper end of the high frequency range to be adjusted precisely to 150 kHz. A small r.f. choke is also included between the output and the attenuator: without this, capacitive loads may also cause the negative feedback loop to burst into unwanted oscillation.

The final part of the oscillator is some form of output attenuator. Here, one can use a straight potential divider which does not present a constant output impedance or alternatively a 600 Ω step attenuator.

The circuit of Fig. 3 shows a 600 Ω attenuator, but the author installed the simpler potentiometer circuit because a separate attenuator was available, and this merely involves omitting resistors R_{18} to R_{22} replacing them by short circuits.

The power supply circuit (Fig. 4) consists of a full-wave voltage doubler feeding a constant-current battery charging circuit. The author claims no originality for the charger circuit, which is used in several pieces of commercial equipment, but the action is not particularly obvious and is worth a few words of explanation. The series transistor feeds the battery from its collector which is a high impedance point and can be used to set a current fairly independently of the circuit voltages. In the tail transistor of the amplifier a simple potential divided set the current but this is not feasible here because we are now dealing with unregulated voltages which would cause the charging current to vary between unacceptable limits. The answer to this is to stabilize the reference voltage with a Zener diode. If the current through the regulator falls slightly then the voltage across the emitter resistor also falls, which increases the base-to-emitter voltage, restoring the current to its original value, although changes in the emitter resistor will naturally allow the charging current to be adjusted within wide limits. This current is normally adjusted so that when the oscillator is operated from the mains, the oscillator is supplied from the charger and the battery is trickle-charged at about 2 mA, but when the oscillator is turned off and the power supply is still on the battery is charged at the full current of 25 mA.

Construction is not critical and any reasonable layout will do. The author followed the normal practice of input on the left-hand side and output on the right and used standard $\frac{3}{4}$ in spaced binding posts which also take a 4 mm plug. The case earth is divorced from the circuit earth as this is a considerable advantage when dealing with very low level circuits, when the case may be earthed to prevent electrostatic pick-up without introducing hum voltages which might be present on an imperfect earthing system. A printed circuit was used by the author, as his colleagues in the printing department are favourably disposed towards electronics, and this printed circuit board is shown in Fig. 4. However, Veroboard or simple tagstrip would be equally satisfactory and the same general form of layout could be used.

Adjustment and Calibration: For calibration there is no substitute for a digital frequency meter in the matter of speed and convenience, and the author suggests that the local technical college be approached on this subject, but do have the oscillator complete and functioning in advance and do not expect to be supplied with mains leads, soldering irons, connecting leads, etc., or you will find their co-operation becomes a little strained. If the layout is mechanically stable with rigid leads, then the frequency stability will be quite within the accuracy of commercial instruments costing a great deal more, so

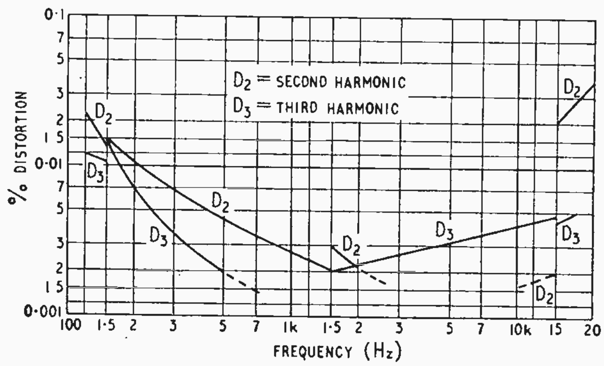


Fig. 6. Distortion characteristics.

PERFORMANCE

Coverage:	1.5 Hz to 150 kHz in five ranges
Frequency stability:	<2% for ± 10 deg C
	<1% for ± 10% mains supply variation
Output voltage:	2.2 V r.m.s. into 600Ω
Output voltage stability:	<3% for ± 10 deg C
	<1% for ± 10% mains supply variation
Output voltage variation with frequency:	<1% 1.5 Hz—150 kHz
Distortion:	see Fig. 6 (<0.01%)
Power supply:	mains and/or 22 V Ni-Cd battery

that a little trouble in calibration is well worthwhile.

It is highly advantageous to have an oscilloscope to make sure that the output is sinusoidal, as some gross mistake in the wiring could permit the oscillator to produce an output which is far from pure, but failing this a pair of high-impedance phones across the output will give a reasonable idea of correct functioning; if the tone sounds pure it is probably all right. A high resistance voltmeter should now be connected between the emitter of Tr3 and chassis earth and the value of R_2 varied until this voltage is within 0.5 V of 11.5 V. This adjustment of R_2 is done by paralleling a second resistor across R_3 and the printed circuit has provision for this. This process is considered easier than selecting a single resistor, as the second resistor can simply be held in parallel with the first until the right value is found and the whole outfit does not have to be switched on and off each time a new value is tried.

The variable emitter resistor of Tr1 is next adjusted so that the r.m.s. output voltage is 2.2 V. This measurement may be made with any good multimeter with the oscillator output set to maximum and the frequency at about 1 kHz.

The dial should now be calibrated on the 150-1500 Hz range, and if 1% tolerance capacitors have been used this calibration will hold within 2% for all ranges with the exception of the 15-150 kHz range. To adjust this range, the dial should be set to the 15 kHz dial mark and the two trimmers adjusted until the frequency is correct with both trimmers set at approximately the same capacitance. This process compensates for variations in stray circuit capacitance and is only necessary on the high frequency range as on the others the tuning capacitance is larger and swamps the strays. The dial is now set to the 150 kHz mark and the trimmer across the 4.7 kΩ resistor adjusted to obtain the correct frequency. These two operations interact a little and should be repeated a couple of times until the frequency is correct at both ends of the scale.

A milliammeter should be connected in series with the battery and R_{25} adjusted until the current is 2-3 mA with the oscillator operating. This adjustment should be made with the battery fully charged.

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IN Mr. Ridler's article, "**Regulated Power Supply with Overload Protection,**" in last month's issue the +10 V line in Fig. 3 should, of course, have terminated at R_{15} and not continued to R_C. In the Fig. 4 the positions of capacitors C_8 and C_9 should be transposed; R_7 should be connected to the +10 V rail; and the upper connection of Tr1 should be marked e.

Improved Biasing for High-Speed Current Switches

By B. L. HART, B.Sc., A.M.I.E.R.E.

The switching time of current steering circuits is directly related to the charge which must be removed from the base of one transistor and transferred to the base of the other. This may be reduced if the bias of the off transistor is set at zero and even further reduced, without significantly affecting d.c. stability, if the off transistor is held at the edge of conduction. The circuit techniques described (which are the subject of a British Patent Application) permit biasing arrangements to achieve these objectives.

WHEN high-speed switching in the nanosecond and sub-nanosecond region is required current mode switching (or steering) is extensively employed. There are a number of sound reasons for this choice of circuit configuration. The underlying philosophy is that it is easier to divert, or re-route an existing current than to establish one *ab initio*. In this arrangement the transistors do not saturate (i.e. they are called upon to support the minimum excess minority-carrier base charge for a given switched current), the output changes are well defined and—a factor of great practical importance—the loading of the power supply units is approximately constant. This arises from the balanced nature of the circuit. Regardless of which transistor is conducting, the total current supplied is the same, and this means that decoupling problems—by no means easily overcome in high speed logic circuits—are eased. (In an unbalanced system 10 mA demanded in 1 ns develops around 0.2 V across only 1in of wire because of inductive effects).

In spite of these advantages the full speed capabilities of the current steering circuit are seldom realized because the requirements for d.c. stability (with respect to temperature effects and component tolerances) result in the charge stored in the emitter depletion layers of the "off" transistor being greater than, or at least comparable with, the base charge required to maintain the full collector current.

In this article, which is basically concerned with principles rather than detailed circuit applications, a biasing scheme that minimizes this effect is discussed. The scheme takes advantage of the close matching which can be obtained between the characteristics of discrete semiconductors made during the same manufacturing process, and the even closer matching which is exhibited by the characteristics of semiconductor devices of the same type fabricated on the same slice of material (i.e. semiconductor integrated circuits).

THE BASIC CURRENT SWITCH

To understand how improved biasing can help increase circuit speed it is necessary to review the elementary current switch. A basic form of this circuit, using two discrete transistors of the same type, is shown in Fig. 1, and its d.c. behaviour is summarized in Fig. 2. The horizontal scale on this diagram has no fundamental significance but is included for clarity of presentation—the inter-relationship of nodal voltages is more readily seen. The charge required to bring Tr1 to the edge of conduction is

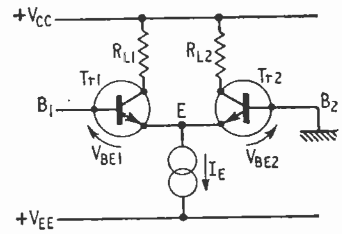

Fig. 1. Basic current switch.

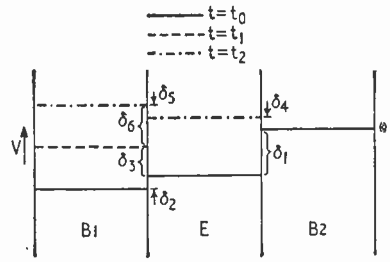

Fig. 2. D.C. behaviour of current switch of Fig. 1. Full line represents condition at each node when $t=t_0$ (Tr1 off, Tr2 on). Broken line corresponds to $t=t_1$ ($t_1>t_0$) when Tr1 is on verge of conduction. For $t_1>t>t_2$, Tr1 is fully on and Tr2 is off, shown by chain line.

$$Q_A = \int_{-\delta_2}^{+\delta_3} C_{te}(v_{BE})dv_{BE} + \int_{V_1}^{V_2} C_{tc}(v_{CB})dv_{CB}$$

where C_{te}, C_{tc} are the emitter and collector transition region capacities and $V_1 = (V_{CC} + \delta + \delta_2)$ and

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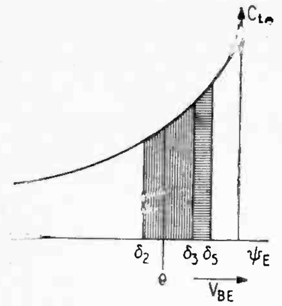

Fig. 3. C_{te} as a function of V_{BE}. When base-emitter junction is forward biased infinite capacitance point is never reached since I_E increases rapidly with V_{BE}.

$V_2 = (V_{CC} + \delta_1 - \delta_3)$. The remaining charge required change Tr1 to the fully conducting state is

$$Q_A = \int_{v_2}^{v_3} C_{tc}(v_{CB})dv_{CB} + I_C\tau_C$$

where $V_2 = (V_{CC} + \delta_1 - \delta_3)$ and $V_3 = (V_{CC} + \delta_1 - \delta_3 - \delta_6 - I_C R_{L1}$. τ_C has its normal charge control theory significance but is referenced to the threshold of conduction level δ_3. With germanium transistors finite δ_2, δ_4 are necessary for accurate definition of switched current (as temperature varies) and adequate noise margins. In that case it is usually arranged that $\delta_4 \approx \delta_2$.

In certain cases the variation of junction capacitance with voltage may be expressed in closed form[1] i.e.

$$C_{tj} = \frac{C_{tj}(0)}{(1 - V/\psi)^n}$$

where $C_{tj}(0)$ = junction capacitance at zero applied voltage
V = applied voltage
ψ = diffusion potential (related to doping levels either side of the junction).
$n = \frac{1}{2}$ for an abrupt junction, and $\frac{1}{3}$ for a linearly graded junction.

(In practice, of course, the depletion layers of diffused transistors do not behave in this ideal way. Should an analytical expression be required a computer curve-fitting procedure must be used to establish the value of n for the real junction after measurements have been made —or see reference 2). Experimental data gives curves of the general shape shown in Figs. 3 & 4. Q_A is given by the sum of the vertically shaded areas in Figs. 3 & 4; Q_B is given by the sum of the horizontally shaded regions. Rigby[3] has shown that it is possible to have $Q_A \approx 4Q_B$ if OC170s are used. With silicon planar devices the ratio is not always so spectacular since C_{te}, τ_C tend to decrease together; however if $Q_A \approx Q_B$ a disproportionate amount of the base charge supplied is still being used to overcome charge storage in the depletion layers arising from bias considerations.

Ways of diminishing this effect have been discussed by

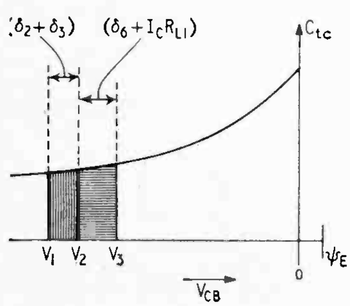

Fig. 4. C_{te} as a function of V_{CB}. Collector-base junction is always reverse-biased in current steering circuits.

Rigby[3] and Miles[4]. The simplest of these—using a resistor between base and emitter of Tr1—is attractive in principle but can lead to practical difficulties. The improved biasing arrangement discussed next effectively allows the resistor scheme and variations of it to be employed.

IMPROVED CURRENT SWITCHES

If two transistors made as an integrated pair are operated at constant emitter currents at the same V_{BE} and the same V_{CB} then the temperature coefficient of their differential base-emitter voltage drops is approximately zero:—

$$\frac{d}{dT}(\Delta V_{BE}) \approx 0 \quad .. \quad .. \quad .. \quad .. \quad (1)$$

This fact was appreciated by Hoffait & Thornton[5]. In their work on d.c. amplifiers they found that, under the above conditions, for 2 × 2N2484 the value of the temperature coefficient was 4 μV/deg. C. Theoretical reasoning (see Appendix) which gives equation 1 also shows that if the base-emitter voltages are not equal, then

$$\frac{d}{dT}(\Delta V_{BE}) \approx + \frac{\Delta V_{BE}}{T} \quad .. \quad .. \quad .. \quad (2)$$

Equations 1 & 2 suggest how the biasing of a current switch may be improved. Consider Fig. 5 in which Tr1, Tr2 and Tr3 are made on the same slice.

I_3 may be chosen so that $V_{BE3} = V_{BE2}$. This has two effects. Firstly (see Fig. 2) $\delta_2 = 0$ and remains so as T varies. The leakage current of Tr1 increases by 15% per deg. C. but it is very small at $V_{BE1} = 0$ so variations have no significant effect on the emitter current of Tr2. Secondly the charge component Q_A is reduced. By suitable choice of I_3, Tr1 can be biased so that $BV_{BE1} = \delta_3$. This last quantity is not well defined. It is ≈ 0.5V for silicon but the value as used in practice depends on the permissible current in Tr1 when this transistor is in the off state. Again, as T varies, the V_{BE} changes in Tr1 and Tr3 tend to cancel but in this case equation 2 shows that the cancellation is not so good as for the previous condition. The charge component Q_A is, however, further reduced as is the input voltage swing required for switching.

In Fig. 5 the collector of Tr3 may be strapped to its base if desired. This is allowable since base width modulation factors are such that differences in collector-base voltage have only a minor effect. (Slight differences in junction temperature then arise between Tr2 and Tr3, since Tr3 operates with lower collector dissipation, but this is also a minor effect, if the three devices are on the same substrate).

The base of Tr1 may be driven by another current switch or from a voltage source. Whichever is used, the current in R_b does not flow into the collector of Tr4 and rob the switched-on transistor of current during the period of the switching pulse; this is one shortcoming of a simple base-emitter resistor. If a voltage source is employed care must be taken in design since any inductance in the base lead can give rise to ringing and oscillation effects and although these may be overcome by using a series resistance this has a detrimental effect on switching speed. Similar considerations apply to the circuits of Figs. 6 and 7, which are suitable for positive going and negative going input pulses respectively.

Direct coupling is used to avoid problems of pattern sensitivity which can arise if the switching waveform is a.c.-coupled. Neglecting recombination in the switch-over process, the switching time will be that required for

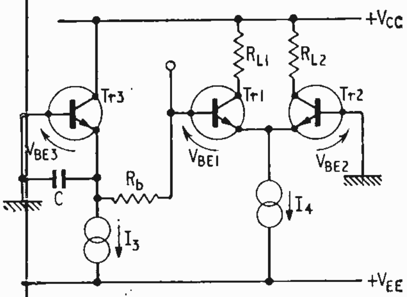

Fig. 5. Current switch biasing in which Tr1, Tr2 and Tr3 are on the same silicon slice.

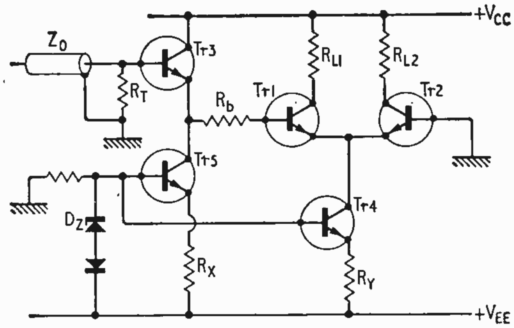

Fig. 6. Current switch for positive input pulses.

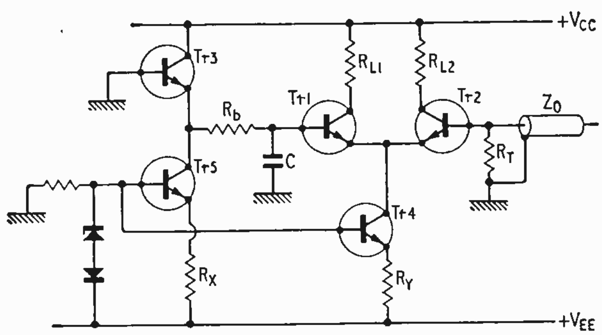

Fig. 7. Current switch for negative input pulses. Capacitor C is a decoupling component.

$Q_A + Q_B$ to be removed from the base of Tr2 and established in the base of Tr1. This is limited by the nature and magnitude of the impedance elements in the series path between the bases. (To achieve a low impedance the un-switched base lead must be short for small inductance; if a base resistance is connected to it as in Fig 7 this must be decoupled by a low-inductance capacitor connected to the ground plane). Exact calculation of the switching time is very complex because of conductivity modulation in the bases and emitter crowding effects. It has been discussed by Ghosh[6].

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APPENDIX

Variation of differential base-emitter voltage with temperature at constant emitter current.

An n-p-n silicon planar transistor having an emitter injection efficiency ≈ 1 and a low base-width modulation factor has an $I_E - V_{BE}$, relationship of the form [7] :—

$$I_E = qAn_i^2\left(\frac{D}{\int_o^W N(x)dx}\right)\left[\exp\left(\frac{qV_{BE}}{kT}\right) - 1\right] \;..$$

where q =magnitude of electronic charge
- A =area of emitter.
- n_i =intrinsic carrier concentration
- D =electron diffusion constant in the base
- $N(x)$ =impurity concentration in the base
- x =base-width variable
- W =base width
- V_{BE} =base-emitter voltage
- k =Boltzmann's constant
- T =absolute temperature

For most practical purposes the relation may be written

$$I_E = qAn_i^2\left(\frac{D}{\int_o^W N(x)dx}\right)\left[\exp\left(\frac{V_{BE}}{V_T}\right)\right] \quad .. \quad \text{A.1(a)}$$

where V_T = thermal voltage.

Now for two transistors close to each other on the same slice of material it may be assumed, as a first approximation, that D, $N(x)$ and W are the same. Thus, if the two devices are operating at the same junction temperature, any differences in their emitter currents for a given V_{BE} is mainly due to differences in emitter areas produced in the masking process.

$$n_i^2 = \text{const. } T^c \exp\left(V_{go}/V_T\right) \quad .. \quad .. \quad .. \quad \text{A.2}$$

In this equation c and the constant are temperature independent and V_{go} = apparent energy gap at 0°K. Combining the last two equations yields, for each transistor.

$$I_E \approx \lambda A T^c \exp\left[-(V_{go} - V_{BE})/V_T\right]$$

in which λ is a constant, not a function of T.

Taking logarithms, differentiating w.r.t. T, and setting $\partial V_{BE}/\partial T = 0$ gives

$$\left.\frac{dV_{BE}}{dT}\right|_{I_E} = -\frac{kc}{q} - \frac{(V_{go} - V_{BE})}{T}$$

If the two transistors are operated at base-emitter voltages V_{BE1} and V_{BE2} then

$$\frac{d}{dT}(V_{BE1} - V_{BE2}) = \left.\frac{d}{dT}(\Delta V_{BE})\right|_{I_E} = \frac{\Delta V_{BE}}{T}$$

This equation demonstrates the validity of Eqn. 2 in the text. Eqn. (1) obviously follows for the special case $V_{BE1} = V_{BE2}$.

Although attention has been directed to two transistors, a simple extension of the above argument shows that it is applicable to three devices manufactured in close proximity.

Variable Selectivity Audio Filter

Theoretical and practical considerations are applied to the design of a narrow band audio filter for the reception of Morse signals

By R. E. SCHEMEL, B.Sc. (Eng.)

THE Nyquist rate of signalling—concerned with the maximum number of changes per second, which can be transmitted over a channel, and yet be received with an arbitrarily low error rate—suggests that Morse signals can be received in a much smaller bandwidth than can be readily obtained at intermediate frequencies. On the short-wave bands at least, there will a real advantage in using as narrow a bandwidth as possible to remove interference, whilst in the reception of weak signals every 2:1 reduction of bandwidth gives a 3 dB improvement in the signal to thermal noise ratio. One way of overcoming the many difficulties involved in constructing a narrow band filter would be to use a much lower i.f. (Now a beat frequency oscillator when functioning correctly acts as a frequency changer, the audio stages acting as a second or third i.f. amplifier). At audio frequencies the construction of such a filter becomes practicable. We can obtain variable selectivity characteristics by cascading a number of circuits similar to those of Fig. 1. At the same time it is possible to build up complex filters by employing this type of circuit. The more sections there are, the more closely does the selectivity characteristic approach the desirable rectangular box shape. However, cost and complexity also increase. In this design two sections have been chosen, and no attempt has been made to obtain a maximally flat response. A Morse bit is best received with a linear phase (constant group delay) filter, and to provide best performance at the narrowest bandwidths each section is stagger tuned.

Fig. 2 shows the circuit of the complete filter, which

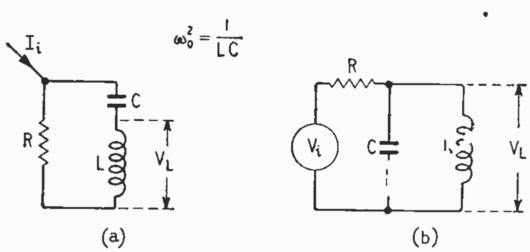


Fig. 1 Basic filter circuits. Variable selectivity can be obtained by cascading several.

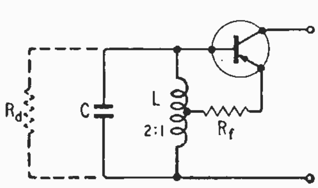


Fig. 2 Essentials of the Q multiplying circuit.

R. E. Schemel, after graduating from King's College, London, worked at Standard Telephones and Cables for several years on microwave relay systems specializing in frequency modulation and demodulation with particular reference to the problems of non-linear distortion. He was then for some time in the future projects department of Hawker Siddeley Dynamics, working on the assessment of multiple access schemes for communication satelites. He is now employed by the European Space Research Organization.

has been designed as an add-on unit to feed headphones. An alternative which many readers might wish for is the filter's inclusion into the audio stages of an existing receiver. This is possible using an emitter follower between the audio driver and the filter input. R_1 is omitted and the output taken from the emitter of Tr2. Another, and perhaps better, alternative might be to use the circuit of Fig. 1(a), the audio driver providing a ready-made current source. There are no problems of instability and the layout is entirely non-critical, except to note that when originally constructed, some forward coupling tended to mar the selectivity. This was traced to a common earthing path between input and output, where high signal currents flowed in the 3.3 ohms load resistance.

One problem which has to be solved is the production of high effective Q inductances. Table 1 shows that a Q of 390 at 1 kc/s is obtainable, but that a cheaper and smaller ferrite has a Q of perhaps 100 at the same frequency. If such coils were used, the minimum bandwidth would be larger than the minimum of 5 c/s. Fortunately this difficulty can be overcome with the highly stable Q multiplier circuit of Fig. 3. The emitter follower type amplifier is capable of giving very stable Q multiplications of up to 100. In the final circuit a minimum Q of about 1000 is all that is required for the circuit to function correctly, and it will not matter if the Q varies or even becomes negative (corresponding to an oscillatory condition without the loading of *RV*) provided that its magnitude is above this figure.

The two tuning capacitors *C* are nominally of value 1.25 μF. The usual type of bypass paper capacitor is quite adequate, but the tolerance of 20% is more than can be taken up within the tuning range of the ferrite. The solution adopted was to make up the necessary value from a 1 μF with a 0.25 μF padder. The output transformer is not specified, as it will depend on secondary impedance and what the constructor has available. The primary impedance should be in the region of 2 kΩ.

The grade of ferrite chosen is left to the reader, as the fact that Q multiplication (Fig. 3) is used does mean

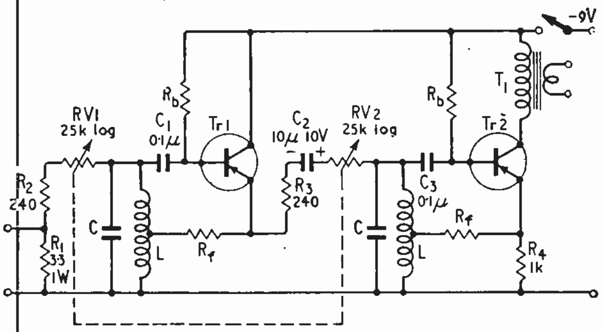
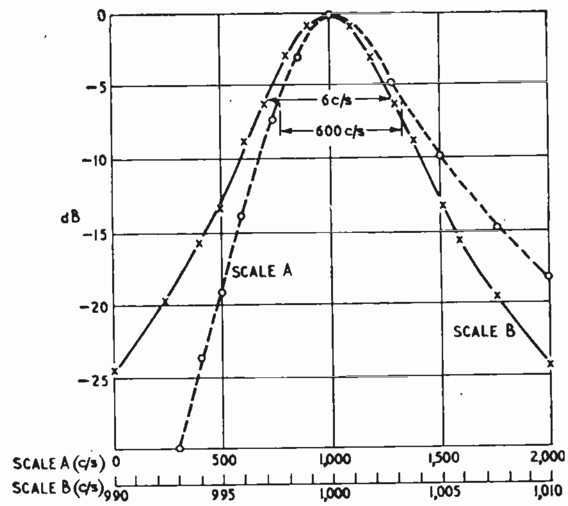


Fig. 3. *Complete Q multiplier circuit. All resistors are* $\frac{1}{10}$*W±10% unless stated otherwise. Transistors are OC7's or similar l.f. types.*



(Right) Fig. 4. *Filter characteristics at minimum and maximum bandwidth positions.*

that a very wide range of materials is suitable. As an example, the original breadboard of the circuit used old television line transformers pressed into service as ready made inductances, and the performance was equal to the finished model using the best available ferrites. Table 1 gives winding details for an inductance of 20.5 mH. The coil is centre tapped.

Some thought has been given to the range of bandwidths, 5 c/s is too narrow to receive morse at 12 w.p.m., and frequency stability problems may also cause difficulties, but nevertheless this bandwidth has been provided so that the reader can determine what really is a minimum. If required, the minimum bandwidth can be made as low as 1 c/s by increasing *RV* by a factor of five. The maximum bandwidth, whilst rather wide for Morse reception, does permit the reception of intelligible speech.

TABLE I

	Q	N	S.W.G.
Mullard LA2131 35 mm violet range	390	98	22
,, LA2331 25 mm ,, ,,	210	122	26
,, LA2433 21 mm ,, ,,	100	166	29
,, LA2002 45 mm yellow ,,	300	138	21
,, LA2103 35 mm ,, ,,	200	155	24
,, LA2303 25 mm ,, ,,	100	192	28

This data was taken from the 1963 Mullard Vinkor handbook, and the quoted codes include a complete mounting assembly.

Calibration of the bandwidth control is carried out experimentally using an ohmmeter. A logarithmic type was chosen, as this gives in theory a constant percentage bandwidth increase with angular rotation. However, a linear potentiometer is equally suitable. The 6 dB designation is a little inaccurate at the maximum bandwidth setting because the circuit really possesses geometric symmetry about the centre frequency. Table II gives the calibration data.

TABLE II

6 dB bandwidth c/s	5	6	7	8	9	10	12	15
Resistance k ohms	25	20.5	17.5	15.4	13.4	12.5	10.2	8.1
6 dB bandwidth c/s	20	30	40	50	70	100	120	150
Resistance ohms	6.0k	3.9k	2.9k	2.3k	1.75k	1.0k	800	585
6 dB bandwidth c/s	200	300	400	500				
Resistance ohms	375	167	62	0				

Low resistance coincides with the expanded resistance part of the potentiometer.

The bias resistances R_b and feedback resistances R_f are both dependent on the type of ferrite used. To set these values, choose resistances for R_f of about 2 to 3 kΩ. Measure the potential of the emitters of Tr 1 and Tr 2 to earth, and adjust these separately to be 4.5 V by means of R_b. Temporarily disconnect R_f from the emitter of Tr 2 and one end of RV_1. R_f may now be adjusted to the nearest standard value in the 5% range so that stage one is just oscillating. It is not important for the circuit to be exactly at the oscillation point. Recheck the emitter voltage of Tr 1; if this is between 3 and 6 V, R_b need not be adjusted further, otherwise the above procedure is repeated for a second time. Disconnect RV_2 and adjust the feedback resistance of the second stage until it is also gently oscillating. With or without introducing a small degree of coupling to the second stage, it should be possible to hear the oscillation of each stage separately. By means of the cores in the ferrites, adjust one of them so that the oscillations are about 5 c/s different in frequency; the setting is not critical. This gives the overall filter response a constant group delay characteristic at the narrowest band setting. A spacing of 7 c/s will give a maximally flat response at 5 c/s and a linear phase response at 7 c/s bandwidth respectively. Some readers may prefer this refinement. Reconnect RV_1 and RV_2. The oscillations should cease, and the filter is then complete and ready for use.

Fig. 4 shows the response of the filter in its two extreme bandwidth positions. There is some asymmetry at the wideband setting for the reasons given earlier about the geometric response, but the slight shift in centre frequency that it causes is unnoticeable in practice. One advantage is that beat notes below about 300 c/s are suppressed very effectively. The finished filter also has a small gain variation as the bandwidth control is altered. This has not been explained, but it is of little consequence in use.

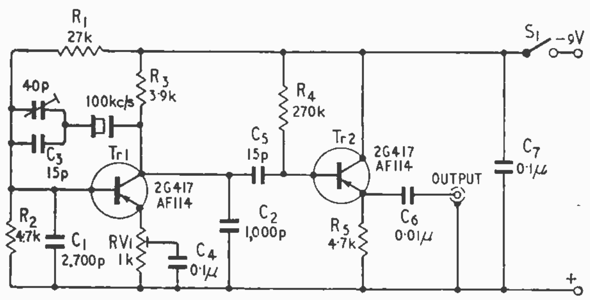
The completed filter has been tested by three competent Morse operators, and their comments are interesting. All found the filter very useful and expressed surprise that such a narrow bandwidth was possible. One operator felt that the minimum setting was of doubtful value and preferred between 15 and 20 c/s. The remaining two found that the minimum was 10 c/s. One thought that a three-section filter might be desirable. All agreed, rather surprisingly, that the widest position was beneficial in the reception of speech. The feature of variable selectivity was considered a major asset.

Transistor Crystal Calibrator

This circuit employs a parallel resonant low-cost surplus crystal

THE crystal oscillator shown may be used to provide reliable calibration signals for communication receivers and signal generators. The circuit is intended for use with surplus 100 kc/s crystals. These are usually DT face shear or +5° X bar extensional types, and are intended for so-called parallel resonant operation. There is considerable variation of the activity and the required load capacitance of the various types. The gain control RV1 and the variable 40 pF capacitor will enable most crystals to be used.

The relatively large capacitors C1 and C2, the high d.c. stability, and the use of a v.h.f. transistor improve the frequency stability by reducing transistor capacitor changes and their effect. The circuit is capable of maintaining the required frequency for reasonably long periods, providing the crystal is kept at a constant temperature. If the circuit is required to drive a clock or is to be used as a laboratory standard the whole circuit should be mounted in an oven. The trimmer should be adjusted until the calibrator signal is at zero beat with the signal of the Light Programme (200 kc/s) or other standard frequency service. The amplitude of oscillation, and the harmonic content is controlled by the gain control RV1. Where the highest possible frequency stability is required, the gain control should be adjusted to provide the lowest amplitude level that is consistent with reliable starting. Low distortion and low amplitude are aids in achieving good frequency stability due to the resulting reduction of crystal ageing and harmonic intermodulation. When the gain control is carefully adjusted, full output is not obtained until about 15 seconds after switch-on.

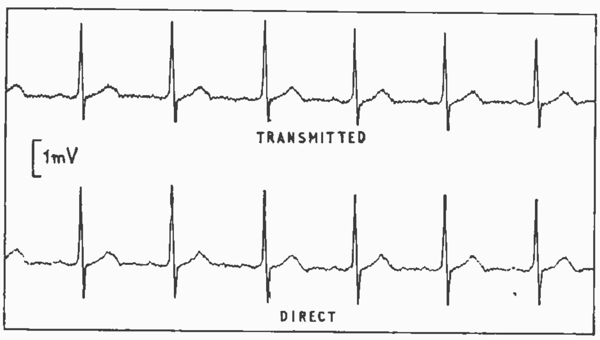


When the circuit is used to provide calibration signals for communication receivers, long term stability will not be essential. The gain control may therefore be adjusted to provide the greater amplitude and harmonic level that may be required. Even then the signal from the calibrator will not be easy to find at high frequencies unless the receiver and calibrator are well screened. If the harmonics prove to be particularly difficult to detect, a Schmitt circuit may be used to provide the harmonic level required.

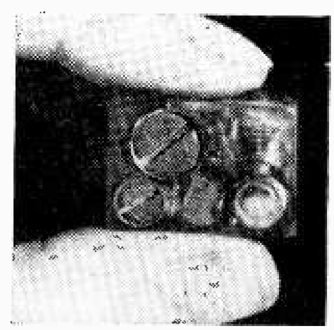
The construction presents no special problems but the wiring to the crystal and adjusting capacitors must be particularly short and rigid. J. ROBERTS

Subminiature Biological Transmitter

A TRANSMITTER with the overall dimensions of 1.5×2.5×0.7 cm has been developed for medical telemetry applications by the Department of Physiology, Royal College of Surgeons. Essentially, it is a tunnel diode oscillator, controlled by a microminiature crystal 6 mm in diameter and 2 mm high. Amplification of input signals is achieved by a four-stage differential amplifier (integrated). This device has been working at distances up to 100ft for the transmission of electrocardiograms operating within the band allocated by the Post Office (102.2 to 102.4 Mc/s) for medical and biological telemetry services. Completely encapsulated, in Araldite resin, this transmitter can be coated with a silicon rubber solution for implant in animals. A pen recorder is connected to an f.m. receiver to reproduce the transmitted biological signal. The frequency response is said to be flat from d.c. to 65 kc/s. It is expected that by using different transducers, other functions such as arterial pressure and respiration rate will also be transmitted.



Left: The tracing shows a transmitted electrocardiogram compared with a simultaneous direct recording.



Some idea of the size of the amplifier can be gained from this photograph.

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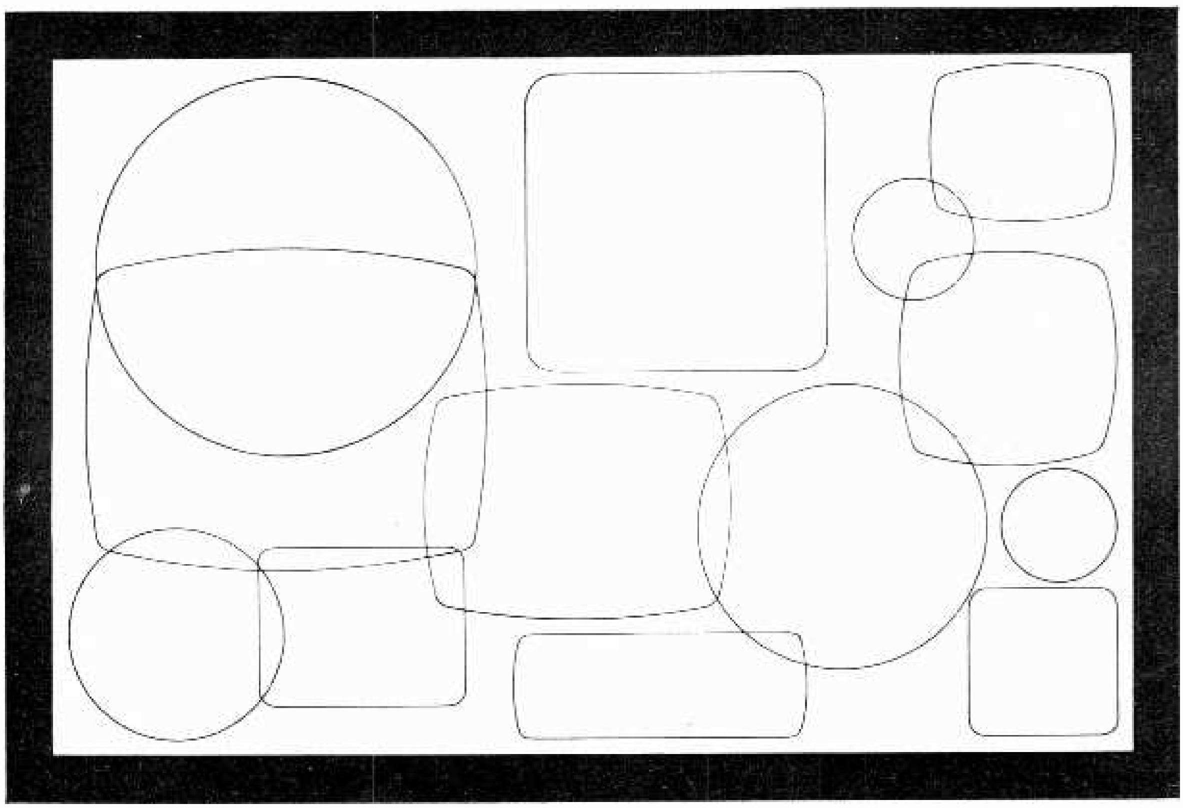

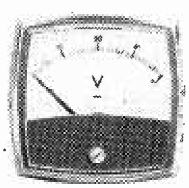

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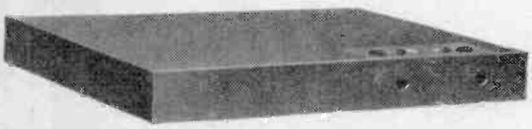

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NOISE FIGURE MEASUREMENT

Theory of noise figure in active devices: Measurement method using signal generator.

By C. N. G. MATTHEWS

THE conventional method of determining noise figure is to use a calibrated noise source with or without an automatic noise figure indicator. An alternative is to use a signal generator and calibrated attenuator. This will give reasonably accurate absolute measurements and remarkably precise comparisons.

Before going into the actual business of measurement it might be advisable to consider the theory of noise figure in active devices, a topic about which remarkably little has been written. Yet the noise figure of a system is an important quality criterion. With the advent of colour tv and the increasing sophistication of communications and telemetry equipment it is likely to assume even greater importance in the future.

If we define noise as spurious signals of random frequencies and amplitudes extending over the entire bandwidth of a system, one difficulty becomes apparent at once. To separate signal from noise and measure the two individually is not a practical proposition. How can we filter off spurious signals which cover the whole bandwidth of a device under test?

A more promising approach is to consider the relationship between input and output signal-to-noise ratios.

If the input signal and input noise powers are S_i and N_i the input signal-to-noise ratio is S_i/N_i.

In a device which generated no internal noise the output signal-to-noise ratio would be gS_i/gN_i where g is the power gain.

So, if S_o and N_o are the signal and noise power outputs,

$$\frac{S_o}{N_o} = \frac{gS_i}{gN_i} = \frac{S_i}{N_i}$$

In other words, the input and output signal-to-noise ratios of a perfect device are equal.

But unfortunately we have not yet learned how to make perfect devices, so a certain amount of internally generated noise is added to the amplified input noise.

The added noise *reduces* the signal-to-noise ratio by a factor F, giving us:

$$\frac{S_i}{N_i} = F\frac{S_o}{N_o} \quad \text{so } F = \frac{S_i/N_i}{S_o/N_o}$$

The factor F is a measure of the deterioration of signal to noise ratio within the system. It is the ratio of input signal-to-noise ratio to output signal-to-noise ratio.

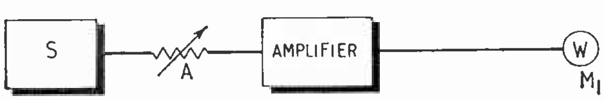
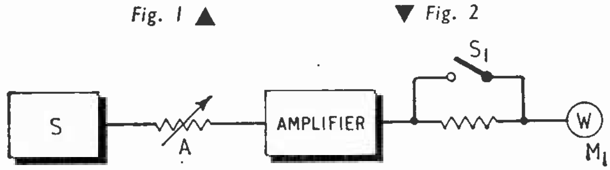


Fig. 1 ▲ ▼ Fig. 2



The noise figure, N.F., of a system is this ratio expressed in dB:

$$\text{N.F.} = 10\log\left(\frac{S_i/N_i}{S_o/N_o}\right)$$

$$= 10\log\frac{S_i}{N_i} - 10\log\frac{S_o}{N_o} \quad \ldots \quad \ldots \quad (1)$$

Although signal and noise cannot be measured separately it is a simple matter to measure the total power output, P_o. Clearly, $P_o = N_o + S_o$ and therefore $S_o = P_o - N_o$.

We can now substitute for S_o in equation 1

$$\text{N.F.} = 10\log\frac{S_i}{N_i} - 10\log\left(\frac{P_o - N_o}{N_o}\right)$$

$$= 10\log\frac{S_i}{N_i} - 10\log\left(\frac{P_o}{N_o} - 1\right) \quad \ldots \quad (2)$$

The input presents as much difficulty in separating signal from noise as the output. But although we cannot directly *measure* S_i/N_i we can *change* this ratio by varying the input signal amplitude. Furthermore, we can monitor the change with a precision limited only by the accuracy of the equipment available.

Now consider Fig. 1, in which the system under test is an amplifier. A signal source S is connected to the input via a variable attenuator A and the output is monitored by a power meter M_1.

Suppose we set A so that the amplifier is operating at the lower limit of normal working conditions. Since this is virtually a no-signal condition the input signal is N_i and the output is N_o.

If we now reduce the attentuation the ratio of the new to the old signal is S_i/N_i. Thus the dB change in attenuator setting is 10 log S_i/N_i. Meter M_1 now indicates $(N_o + S_o) = P_o$.

Now if we make the change such that the meter indication is exactly doubled, P_o becomes $2N_o$. So for this condition we can rewrite equation 2 as

$$\text{N.F.} = 10\log\frac{S_i}{N_i} - 10\log\left(\frac{2N_o}{N_o} - 1\right)$$

$$= 10\log\frac{S_i}{N_i}$$

As we have seen, 10 log S_i/N_i is the amount by which the attenuator setting was changed. So noise figure is read directly as the dB increase in input necessary to double the output power.

The practical circuit shown in Fig. 2 is identical with the first circuit except that a switched 3 dB attenuator is included between amplifier and power meter.

To make a noise figure measurement S_1 is closed and the attenuator is adjusted so that the amplifier power output is about half that for normal working. The meter range is selected to give something like full-scale deflection. S_1 is now opened and the attenuator is adjusted till the original meter indication is restored.

Switching in the 3 dB attenuator, of course, halves the power passed on to the meter. So to restore its initial deflection the amplifier output must be doubled. This means that the amplifier is operating at its normal level when the measurement is made. The switched 3 dB attenuator is an improvement because it enables the meter to be used as a reference level indicator rather than as an absolute measuring instrument. This eliminates errors from such causes as poor scale shape or bolometer non-linearity. Provided that the signal generator and input attenuator are adequately screened, this method gives results that agree reasonably closely with those taken on an automatic noise figure indicator using a calibrated diode noise source.

Slightly less accurate measurements can be taken by using a suitable r.f. voltmeter instead of the power meter. The voltmeter has the advantage of greater stability, but to double the power output its indication must be squared. Under these circumstances the 3 dB attenuator cannot be used and the possibility arises of errors due to non-linearities, especially as the measurements may be taken on two different ranges of the meter. With care though, the voltmeter is useful, especially for optimisation.

Choice of voltmeter is important, because a peak-responding instrument calibrated in r.m.s. will not respond to the entire noise signal. Consequently the instrument will read low and noise figure measurements will be high. A meter that responds to r.m.s. volts is essential. In tests on this method a Boonton Electronics 91DA voltmeter was used. This is true r.m.s. responding at input levels below three volts.

In an experimental comparison of the two methods, the noise figures of three 30 Mc/s, 40 dB amplifiers were measured using a recently calibrated Magnetic AB automatic noise figure indicator with a specified accuracy of ±0.5 dB. The measurements were then taken again, first with a power meter and then with the voltmeter. Here are the comparative results:

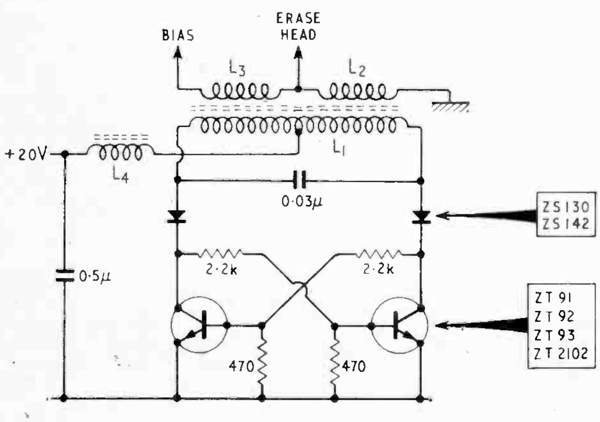
Amplifier	A	B	C
Automatic	7.0 dB	7.3 dB	6.8 dB
Power meter	7.6 dB	8.0 dB	7.5 dB
Voltmeter	8.0 dB	8.2 dB	8.0 dB

None of the discrepancies can be called glaring, but the general trend is for the power meter to give slightly high results and for the voltmeter to measure a little higher.

General Purpose Class D Oscillator for Tape Recorder

SINCE publication of the articles describing a silicon transistor tape recorder in the July and August 1965 issues another oscillator design has been evolved by Ferranti which can be used with a variety of erase heads. The oscillator is a class D type of the current steering or switching variety; being chosen for efficiency, good stability, and low even-order harmonic content. (The term class D, incidentally, was originally suggested by P. J. Baxandall in 1959 for describing the operation of oscillators in which the transistors act as bottoming switches, the voltage across the transistors being substantially zero when the current is finite and the current being negligible when the voltage is finite, thus leading to low device dissipation and a high potential efficiency. Current switching oscillators of this type have had efficiencies as high as 97%.)

The circuit shown is designed to be used with almost any erase head without major modification and can supply up to 3 W of power. The transformer primary is tuned to 55 kc/s and by making the Q of this circuit a reasonable value, low distortion can be obtained. (Waveform distortion is related to Q in this type of oscillator.) The two diodes are used to protect the transistors against negative voltage. Inductor L_1 (2.5mH) approximates a constant current source and if wound on a Mullard LA2 ferrite core with 34 s.w.g. wire should be 100 turns. The transformer, also wound on an LA2 core and with 34 s.w.g. wire, should have a primary of 36 turns and be centre-tapped. The number of turns on the erase winding depends on the head used and the table gives a guide for a number of heads. (If other heads are to be accommodated, the erase turns can be made equal to about 1.3 times the required erase voltage.) The number of turns used for the bias winding (L_2+L_3) again depends on the head. The arrangement used in the original design may be adopted, i.e. of developing a greater voltage than required and attenuating with a series capacitor and potentiometer.



Head type	Tape deck	Erase voltage	Erase turns
Bogen UL117	—	38	25
,, BL210	—	28	20
,, BL216	—	28	20
Bradmatic CE1	Collaro Studio II Magnavox 363 (2 track)	28	24
Marriott X/ES/11	Collaro Studio Magnavox 363 (4 track)	20	16
Reuter	Collaro Studio I Collaro Transcriptor Mk. III & IV (2 track)	28	25

Short-wave Disturbances

Their correlation with solar activity

By L. J. PRECHNER,* B.Sc., A.M.I.E.E.

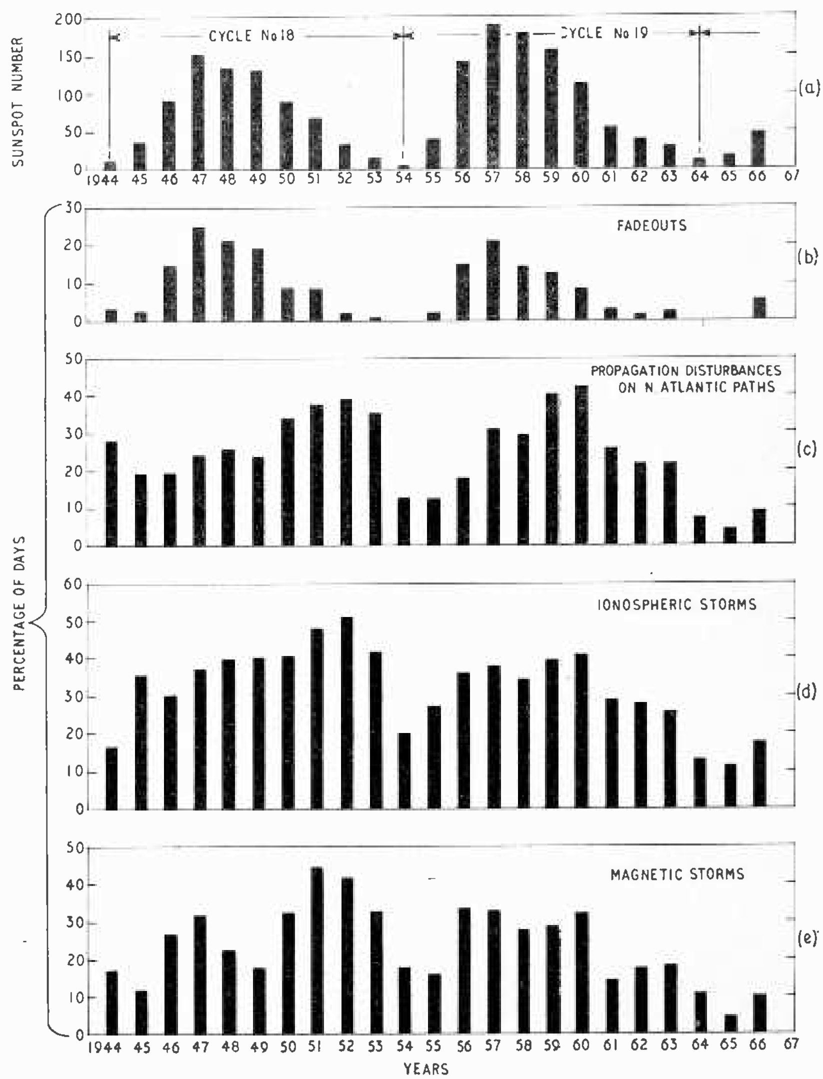
IONOSPHERIC storms are perhaps the most common cause of interruption in short-wave propagation. They result from the failure of the ionosphere to reflect high-frequency transmissions, usually because of the decrease in the ionization density of the top (F_2) layer. The so-called short-wave fadeouts (SWFs) are the second major cause. They result from an abnormal increase in the absorption of high frequencies in the bottom (D) layer of the ionosphere. This increased absorption is caused by the hard X-rays emitted by solar flares, which are short-lived "hot regions" associated with sunspots. These fadeouts rarely last more than 1 or 2 hours.

The accompanying histograms illustrate the marked correlation between the short-wave disturbances and the solar activity of the last two sunspot cycles (1944-54 and 1954-64). A similar pattern is also becoming apparent in the current cycle.

Diagram (b) shows the annual percentages of days during which short-wave fadeouts were recorded in Southern England by Cable & Wireless Ltd. and the B.B.C. It will be seen that they occurred on about 20-25% of all days during the sunspot maximum years (1947 and 1957) but they were negligible at the minima (1944, 1954 and 1964). Thus the correlation of the incidence of fadeouts with the solar activity (diagram a) was very marked.

Diagram (c) shows the percentage of days in each year from 1944-66 during which short-wave propagation disturbances on the North Atlantic paths were recorded by the B.B.C. Receiving Station at Tatsfield, Kent. It is worth noting from diagrams (a) and (c) that the least disturbed years corresponded with the years of low solar activity, but the most disturbed years occurred a few years after the sunspot maxima.

There were similar trends in the incidence of ionospheric storms (diagram (d)) as shown by the decrease in the critical frequencies of the F_2 layer recorded by the Radio and Space Research Station near Slough, Bucks., and also in the incidence of magnetic storms (e) recorded by the Royal Greenwich Observatory at Hartland Point, Devon. This delay (compared with peak solar activity) in reaching the peak of ionospheric and magnetic storms is due to the so-called "recurrent storms." These storms are thought to be caused by long-lived sources of corpuscular radiation ("M" regions) on the sun and they therefore recur at intervals of about 27 days, corresponding to the synodic period of axial rotation of the sun. These disturbances sometimes recur over periods of a year or more (e.g., from the autumn of 1962 to the early part of 1964) but are often most prevalent during the few years immediately following the sunspot maximum.



Transistor Line Output and E.H.T. for Colour TV

IN deciding to make their first colour television receiver an all-silicon solid-state design* instead of the more common valve/transistor hybrid, British Radio Corporation set themselves the particular problem of how to get adequate power from transistor line output and e.h.t. stages. Something in the region of 4,000 VA was needed to drive the line deflection coils and about 1500 VA for the 24-kV e.h.t. supply for the 25-in shadow-mask cathode-ray tube. The problem was solved by a combination of careful, slightly unconventional, engineering and the fortuitous arrival on the market of some new silicon power transistors of high performance (see table). The results can be seen in the two circuits published here. It will be noted immediately that there are, in fact, *two* units. B.R.C. decided at an early stage that it would be impracticable to attempt to generate the e.h.t. from the flyback voltage pulses in the line output stage, as is normal practice in valve circuits, so they designed a separate, voltage-regulated, e.h.t. generator—although this is nevertheless driven from the same 15-kc/s oscillator as the line output stage.

This line-scan oscillator can be seen as Tr2 in Fig. 1, the line timebase circuit. The oscillator is a modified Hartley type, with frequency determined by an inductor and two capacitors, and oscillates by virtue of a 180° phase difference between the collector and base voltages. It might be described as an "over-coupled Hartley oscillator," since for part of each cycle it operates in the normal Hartley manner while for the remainder of the cycle the transistor is cut off and the mode of operation is that of a blocking oscillator. The on/off time of the transistor is determined by the 2.7-kΩ base resistor. This particular configuration was adopted, in preference to a normal blocking oscillator, because of the need to obtain good frequency stability in order to avoid increasing the ratings of the line output transistors. Frequency change with "hold" control variation is adequately catered for.

Transistor Tr1 is the control stage for the oscillator and its output is fed to the base of Tr2 via an OA81 diode, which prevents over-running of the base-emitter voltage of Tr2. Transformer T1 gives the current step-up (6:1) necessary for driving power transistor Tr3, while the voltage-dependent resistor is used for limiting spikes of damped oscillatory voltage which appear on the leading edges of the 15-kc/s pulses at the collector of Tr2. Line driver transistor Tr3, cut off in its quiescent condition, is turned on hard and bottomed by the pulses applied to its base. The resulting square waveform at the collector is fed to the primary of T2, which gives a 10:1 current step up to three secondaries driving the bases of two line output transistors, Tr4 and Tr5, and that of the e.h.t. generator transistor (Fig. 2). Peak currents of about 1 amp flow into these bases. Spikes on the pulse waveform are prevented by a 0.01-μF capa-

* A brief description of the set was given in the June 1967 issue, p. 305.

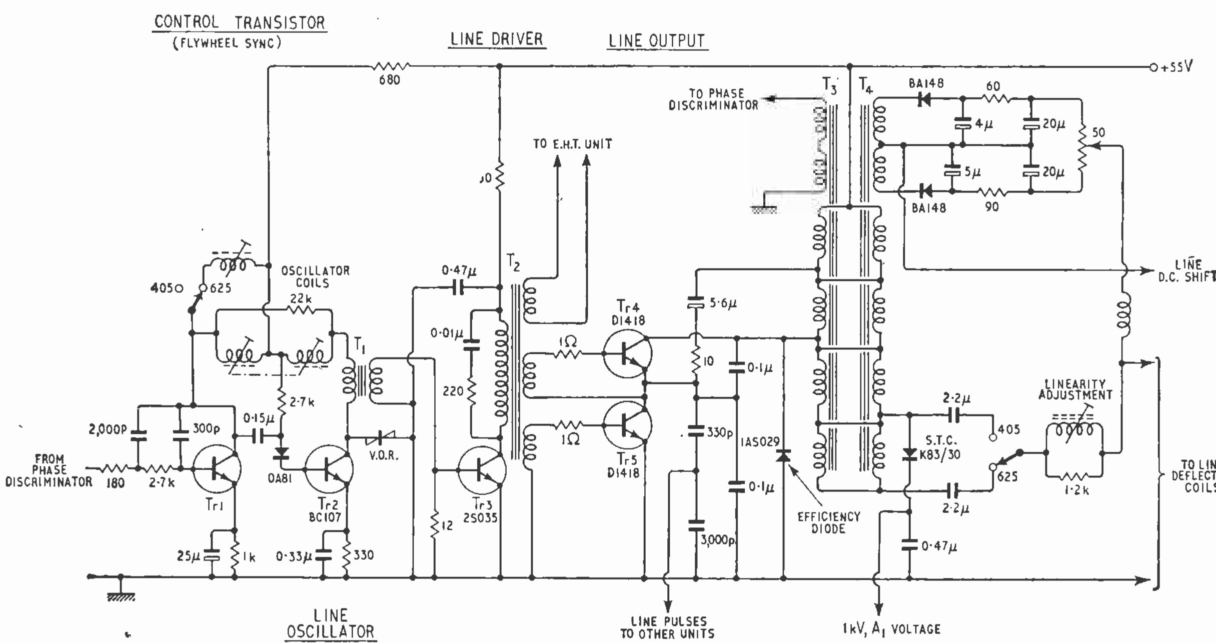


Fig. 1. Part of circuit of line timebase (not including all oscilliator control circuitry and all 625/405 switching)

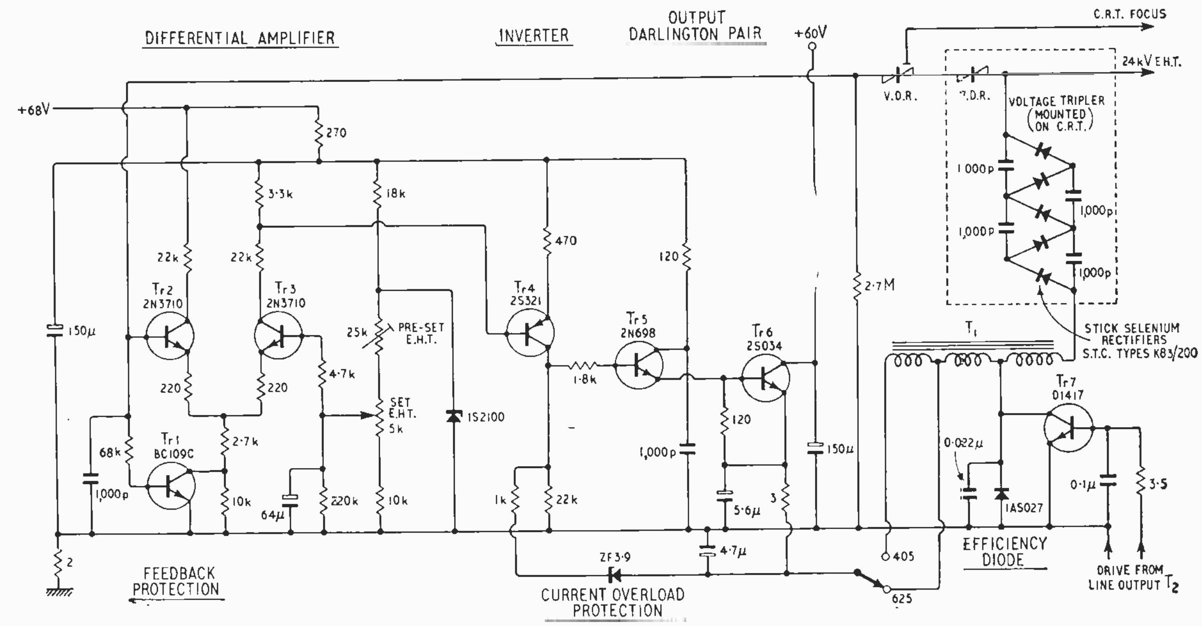


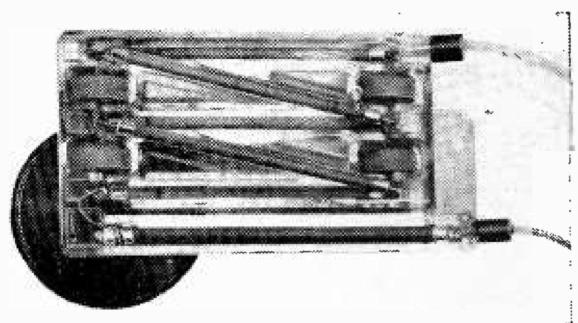
Fig. 2. Circuit of e.h.t. unit. The e.h.t. generator itself is on the right and the closed-loop voltage-regulation circuitry on the left.

citor and a 220-Ω resistor across the primary of T2.

Two line output transistors in series are used in order to withstand the back e.m.f. from the paralleled line output autotransformers, T3 and T4, which amounts to about 600 V. It was decided to have two paralleled transformers, instead of the usual one, in order to allow smaller gauge wire to be used for the windings: this facilitates manufacture of the windings and provides the closer coupling between turns necessary for obtaining the required inductance. The autotransformers are actually the "jelly-pot" type similar to that used in current B.R.C. monochrome receivers and have E ferrite cores.

In operation the line output stage is similar to the circuit commonly used in monochrome receivers, and has an efficiency diode (the 1ASO29) to provide the first part of the scan. When Tr4 and Tr5 are switched on by a positive pulse there is a linearly rising current through the inductance of the primary of autotransformer T3/T4 (connected between the collector of Tr4 and +55 V) and this induces a corresponding ramp function in the secondary (lower sections of the windings shown), which is fed through the deflection coils connected between the "hot" (lower) end of the windings and chassis. When the transistor is cut off current continues to flow in the deflection coils because of their magnetically stored energy, but it flows into the capacitance in shunt with them. The sequence of events from this point is explained in principle on p. 375 of this issue, in the series "Colour Receiver Techniques," and from this it will be seen how the efficiency diode conducts and produces the first part of the scan waveform. The total excursion of the current ramp function in the primary of T3/T4 is 12 A on 625 lines (10 A on 405 lines).

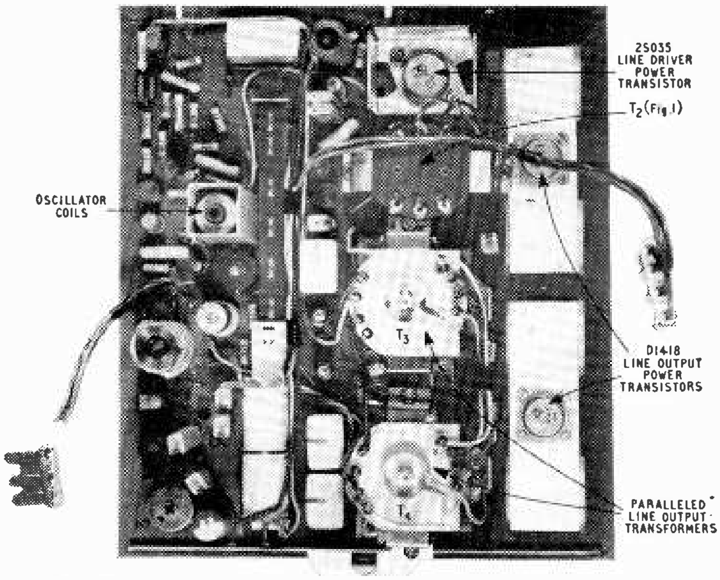
On the left-hand side of T3/T4, the 5.6-μF capacitor and 10 Ω resistor constitute a feedback network for balancing the two transistors Tr4 and Tr5, while the two 0.1-μF capacitors are tuning capacitors. On the right-hand side of T3/T4 the network at the top is for producing a direct current for picture-shift through

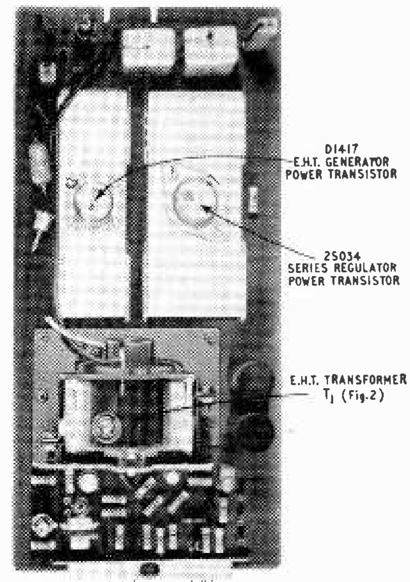


The e.h.t. Cockcroft-Walton voltage tripler is a separate sealed unit which is attached directly to the c.r.t final anode by a clip connector (in the centre of a black circular insulating cover which can be seen underneath)

the deflection coils, while below is a selenium rectifier (K83/30) and capacitor for obtaining a 1000-V supply for the first anodes of the colour tube. The adjacent 2.2-μF capacitors in the deflection coil circuit are waveform correctors, as is also the "linearity" variable inductor.

Passing now to the e.h.t. unit, Fig. 2, it will be seen that the actual e.h.t. generator operates on the "ringing-choke" principle familiar in line-flyback e.h.t. systems. The power transistor Tr7, normally cut-off, is turned on by positive pulses at the base from the line driver stage in Fig. 1 and the resulting current pulses in the collector circuit pass through the primary of autotransformer T1. The high rate-of-change of current at the rear edge of each pulse (at which Tr7 is cut off) induces a high-voltage pulse of 7-8 kV in the secondary (which gives a voltage step-up of 30:1 from the primary). The primary current waveform is a sawtooth, as in a line output stage, and the beginning of the ramp is again formed by use of an efficiency diode (the 1ASO27), the purpose being here to obtain a sufficiently large current excur-





Left: Layout of line timebase unit on replaceable printed-circuit module. Right: The e.h.t. unit on its replaceable printed-circuit module.

sion in the "flyback" period. (Energy storage in this case occurs in the transformer.) In this circuit, however, the rate-of-change of this current excursion has been made only about half that of the flyback in the line timebase (Fig. 1). This has been done to obtain higher efficiency (e.g. by virtue of the T1 ferrite core working with lower frequency components and so reducing losses), so that only one transistor need be used instead of two. The 7-8 kV pulses from the high voltage winding of T1 are then fed to a Cockcroft-Walton voltage tripler to produce the required 24-kV e.h.t. supply. This technique has been used in B.B.C. monochrome receivers for over two years.

Voltage regulation in the unit is achieved by measuring the e.h.t. voltage and controlling the supply voltage to the e.h.t. generator transistor Tr7 accordingly—an increase in e.h.t. causing a reduction of supply voltage, and vice versa. The e.h.t. voltage is measured by tapping off a sample from a potential divider connected across the 24-kV line (two voltage-dependent resistors and the 2.7 MΩ in Fig 2) and comparing this with a reference voltage (from the 50-kΩ "set e.h.t." potentiometer) in the differential amplifier Tr2/Tr3. The d.c. error signal resulting from this comparison is amplified in Tr4 and applied to the Darlington pair Tr5/Tr6, of which Tr6 is actually a series-regulator power transistor controlling the voltage supply to Tr7. Voltage-dependent resistors have been used in the 24-kV potentiometer because, by virtue of their non-linear characteristic, they provide some degree of stabilization themselves and also magnify the voltage changes before these are applied to the differential amplifier. One of the v.d.rs has a metal slide tapping which provides the focus voltage for the c.r.t.

Transistor Tr1 is part of a safety circuit which prevents the closed-loop system from driving up the e.h.t. voltage to an excessive value if there is a loss of measurement information coming into the base of Tr2. Any such loss of voltage at this point causes Tr1 to cut off and as a result an extra 10 kΩ is introduced into the common cathode circuit of Tr2/Tr3. This has the effect of turning off Tr3 and, in effect, lowering the reference voltage. Flashovers in the colour c.r.t. can cause temporary drops in e.h.t., and to cope with this situation the ZF3.9 Zener diode is used to limit the current fed into Tr7 (to about 1 A) when such a flashover occurs.

The silicon power transistors D1417 and D1418, developed by Texas Instruments in conjunction with B.R.C., have an n-p-n triple diffused mesa construction which results in high switching speed and good transient power handling capacity. The etched mesa construction provides a measure of junction shaping which concentrates avalanche currents in the bulk of the semiconductor rather than at the surface where localised heating would occur. The resulting robustness makes possible operation without failure during standards switching and c.r.t. flashover with minimum protective circuitry. The transistors are housed in TO-3 cans. The device specifications are:

	D1417	D1418
Maximum ratings:		
V_{CBO}	325 V	300 V
V_{CEX}	325 V	300 V
V_{EBO}	8 V	8 V
$I_{C(PEAK)}$	10 A	10 A
$I_{C(D.C.)}$	7 A	7 A
$I_{B(PEAK)}$	3 A	5 A
T Case (normal operation)	100°C	100°C
Typical performance:		
$V_{CE(SAT)}$	1.0 V	1.0 V
$V_{BE(SAT)}$	1.0 V	1.2 V
Fall time	0.5μs	0.5μs
Storage time	2.0μs	1.7μs

Measurement conditions:
D1417: $I_C=4$ A, $I_{B_1}=0.5$ A, $I_{B_2}=1.0$ A
D1418: $I_C=7$ A, $I_{B_1}=1.4$ A, $I_{B_2}=3.0$ A
Note: B1 indicates turn-on base current and B2 indicates turn-off base current.

The type numbers are provisional developmental ones, and eventually the D1417 will be issued as a BU106 and the D1418 as a BU107.

An additional safety circuit, not shown, operates from the voltage developed by the generator current flowing through the 2Ω resistor (extreme left) connecting the common negative suply to chassis. The object of this circuit, a beam current limiter, is to prevent the tube manufacturer's beam current rating from being exceeded by too high a screen brightness. The nearer the brightness control potentiometer is moved to its maximum position, the more the potentiometer element is shunted by a transistor being progressively turned on, so that the control has progressively less effect. The 2-Ω resistor is used to ensure that the voltage across it produced by the mean e.h.t.-generator current turns on the shunting transistor abruptly only when the beam current exceeds 1 mA.

The performance of the closed-loop e.h.t. regulation system is such that a 10-20 μA change in the feedback current flowing into the base of Tr2 (normally about 100 μA) results in a 30-V swing in the supply voltage to Tr7. Such a 30-V swing would be initiated by a 1 mA variation in tube beam current resulting from a change of picture content from peak white to black.

Literature Received

Equipment for **monochrome and colour television measurements** is described in the 41-page brochure "Philips Measuring Equipment for black/white and colour television" issued by M.E.L. Equipment Ltd., Philips Electronic and Control Division, Manor Royal, Crawley, Sussex. These Philips instruments are divided into two categories—professional and service. All the instruments offer dependable performance even with fluctuations in mains voltage up to 20%, and in a temperature range from −10 to +40°C. All television pulse inputs are insensitive to superimposed hum and variations in input level. Among the professional equipment are pattern generators, N.T.S.C. and PAL encoders, sub-carrier generator, and i.f. modulator. The service instruments include a black/white and colour pattern generator (PAL), a television service generator (PAL), wobbulator, and u.h.f. generator.

WW 330 for further details

Directional h.f. aerials for point-to-point or sector coverage in moderate environments, omnidirectional h.f. aerials for broadcast or communications with mobile terminals, steerable h.f. aerials and other types are presented in the 14-page "**Communications Equipment**" short-form catalogue received from Granger Associates Ltd., Russell House, Molesey Road, Walton-on-Thames, Surrey. Each technical specification is accompanied by a simple pictorial diagram and an elevation-plane radiation pattern. Balun transformers, transmitting multicouplers, broadband h.f. power amplifiers, ionosphere sounders for h.f. communication and research, and aviation communications equipment are also described.

WW 331 for further details

A four-page data sheet available from Gresham Lion Electronics Ltd., Twickenham Road, Hanworth, Middlesex, describes the GX range of **modular constructed power supplies.** A functional diagram is given with a full specification that includes input and output parameters, temperature factors, overload protection, and rack mounting details.

WW 332 for further details

G. W. Smith & Co. (Radio) Ltd., of 3 and 34 Lisle Street, London, W.C.2, have just issued their **first comprehensive 153-page catalogue** of electronic components and equipment ranging from aerials to Zener diodes.

WW 333 for further details

"Courier" No. 12 gives details of both EMT and Studer equipment (German) designed to meet the **requirements of modern recording studios.** Professional turntables, track position indicators, cutting characteristic dummy network, tape recorders (including a machine for simultaneous four-channel recording on one-inch tape with four full tracks), reverberation plates, mixers, polarity tester, wow and flutter meter, and a stereo transmission monitor are all described in this eight-page newsletter available from F. W. O. Bauch Ltd., Holbrook House, Cockfosters, Herts.

WW 334 for further details

The use of n-p-n and p-n-p silicon planar epitaxial transistors is fully discussed, and circuits and component lists given in the 47-page Ferranti book **High Fidelity Audio Designs.** It includes pre-amplifiers, 7, 15, 30 and 150 W a.f. amplifiers, associated power supplies, and tape circuits (some of which have already appeared in *Wireless World*). These designs include a playback amplifier, recording amplifier, record level indicator, bias/erase oscillator and an f.m. tuner. Guidance on layout and assembly for printed circuit boards is also provided. From Ferranti Ltd., Gem Mill, Oldham, Lancs.

WW 335 for further details

Leaflet giving details of Siemens **silicon planar diodes for tuners in radio and TV receivers** is available from Cole Electronics at 7 Lansdowne Road, Croydon CR92HB. The four-page data summary is written in both German and English and includes variable capacitance diodes for tuning in v.h.f. receivers. The most important characteristics (reverse bias *v* capacitance, series resistance and Q) are given of tuning diodes BB103, BA138, BA139 and BA140. Also included are the BB104 double tuning diode for push-pull tuning and the BA136 diode intended for waveband switching.

WW 336 for further details

Three leaflets describe apparatus for the **manufacture of printed circuit boards.** The units are a wash-off unit in stainless steel or p.v.c., a tank with motor for rapid developing of rigid and flexible printed circuit boards. There is also a dip coater designed for resist coating of boards. Procirc Co. Ltd., Station Road, West Haddon, Near Rugby, Warwicks.

WW 337 for further details

Over 150 **electronic kits** are mentioned in the latest 32-page edition of the Heathkit catalogue 87/2. As well as stereo and mono equipment there are domestic and communication receivers, test and laboratory instruments, intercom, and educational kits. Daystrom Ltd., Gloucester.

WW 338 for further details

A variable **u.h.f. tuner of quarter wavelength design** is fully discussed in the 12-page technical bulletin, ref. 2-6300-124 from Siemens, Germany. This tuner operates with the mesa transistors AF239 and AF139. Information is provided on the mechanical design of the tuner, and characteristic measured values. Sections are devoted to such aspects as noise and gain, band-pass filter, cross-modulation, temperature drift, interference due to radiation, oscillator, and mixing stages. The text is in German. Siemens Aktiengesellschaft, Technischer Pressedienst, 8000 Munchen 1, Postfach 463, Oskar-von-Miller-Ring 18.

WW 339 for further details

"**The Wells Series of Nucleonic Instrument Modules**" is the title of a 15-page brochure covering mains supply, high voltage, and power units. Linear, quench, and head amplifiers, scaler/timers, pulse height analysers, ratemeters, and other apparatus are also fully described. E.M.I. Electronics Ltd., Hayes, Middlesex.

WW 340 for further details

TECHNICAL NOTEBOOK

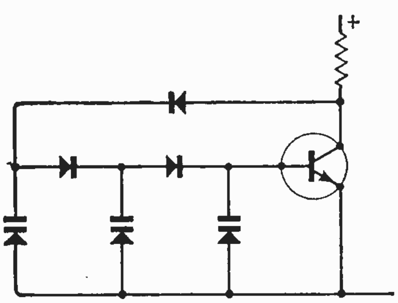
S.C.Rs in Integrated Circuits

ONE of the problems associated with silicon integrated circuits is the difficulty involved in fabricating p-n-p transistors on the same chip as n-p-n transistors. (P-n-p transistors are valuable for shifting voltage levels and replacing resistors.) One method which has been used to make p-n-p transistors is to arrange the layers laterally rather than vertically. Emitter and collector are formed on either side of the base by photomasking rather than by the normal vertical diffusions. (Transistors made this way are limited in gain and frequency response, however.)

This process has made possible the fabrication of s.c.rs having a n-p-n-p structure, the device being a combination of a lateral p-n-p and the basic diffused n-p-n transistor types. Such controlled rectifiers have been made which can handle currents of 3 A and at frequencies of 200 kc/s—low by i.c. standards but high by discrete thyristor standards, which are typically limited to around 20 kc/s. Six s.c.rs have so far been fabricated on a single chip, but the developers, Westinghouse Molecular Electronics division, are now working on a 20 s.c.r. chip.

Novel *RC* Sweep Oscillator

PHASE-SHIFT resistance-capacitance oscillators in which both *R* and *C* elements are varied have recently been developed and used in commercial sweep generators. Unusually, the basic three-section *RC* oscillator uses p-i-n diodes for the resistances and voltage-variable capacitance diodes for the capacitors. The p-i-n diodes act as voltage-sensitive resistances thus making the *RC* product variable in all three sections. Normally p-i-n diodes find application in the microwave region, but their use at lower frequencies (v.h.f.), mainly as attenuators, is becoming increasingly apparent. The skeleton circuit is shown in the Fig. and this arrangement is said to give a 20:1 swing in frequency, approximately in the region of 10-250 Mc/s. Commercial sweep generators employing this type of circuit are being manufactured by Kruse-Storke Electronics of California, according to a recent issue of *Electronic News*.



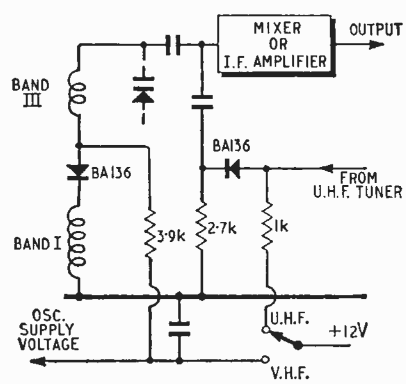
Diode Switching in Tuners

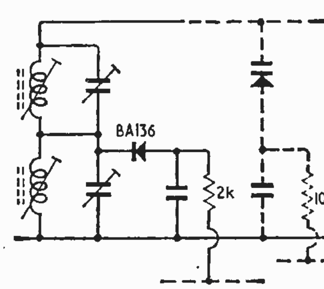
VARIABLE capacitance diodes are now frequently found in v.h.f. and u.h.f. tuner designs, both for sound and television receivers, and particularly in those designs produced on the Continent. Another obvious application for diodes in tuners is in channel or range switching and a number of makers (e.g., Siemens, S.T.C.) now produce diodes tailored specially for v.h.f. switching use.

In a v.h.f. tuner design by Hopt (a German company with a new British offshoot Hopt Electronics Ltd.—see p. 352, July issue) two diodes are used for switching from v.h.f. to u.h.f. and not for channel or band switching (Band I-III switching is achieved by magnetically operated reed switches). A separate u.h.f. tuner is used with this arrangement. The mixed transistor (AF 106) of the v.h.f. tuner is fed with the signal and oscillator frequencies for both Band I and Band III reception but acts as an i.f. amplifier for u.h.f. reception. Switching to u.h.f. disconnects the Band I and III tuned circuits (which are in series) by removing the forward bias from the BA136 diode (see Fig.). A second diode connects the output of the u.h.f. tuner to the AF106 by forward biasing the second BA136 diode.

In a circuit developed at Siemens, diodes are used for channel switching in a v.h.f. tuner (see lower Fig.). In one design five BA136 diodes are used to short-circuit part of the tuned circuits (which are voltage-tuned, incidentally). A reverse bias of 30 V applied to the diodes results in a capacitance of less than 2 pF (typically 1 pF) while a forward bias gives a dynamic forward resistance of 0.45Ω (at 100mA. At 10 mA this rises to 4.5Ω).

A BA143 diode has recently been announced (by STC) and this, with its low series inductance of 2 nH, should enable switching to be carried out in the u.h.f. region. We understand that a combined u.h.f./v.h.f. tuner using this diode will be introduced toward the end of this year.





A Linear Scaled Ohmmeter

Resistance measurement using a constant current source and a high-impedance voltmeter

By J. B. WILLIAMS, B.Sc.(Eng.), A.C.G.I.

IT could be argued that the normal method of resistance measurement is perfectly satisfactory, it is however open to certain objections not least of which is the backward-reading non-linear scale. The circuit described here, although open to the criticism of complexity, does away with special meter scaling, achieves greater accuracy when measuring high resistances and is suitable as the basis of a digital ohmmeter. If supplied from a stabilized power supply, resistance measurements better than 2 % can be made though this performance dropped away on the bottom range of the prototype mainly due to the resistance picked up in the check/read switch. If batteries are used or the setting up is not quite correct the measurements over most of the range should still be to this accuracy.

The principle normally employed by ohmmeters is to have a fixed source of voltage, the meter in fact measuring the current through the unknown resistance, thus from Ohm's Law.

$$I \propto 1/R$$

giving an awkward scale. If the current through the unknown resistor is fixed and the voltage across it is measured a linear scale is achieved as

$$V \propto R.$$

The configuration is shown basically in figure 1(a). The constant current source may be provided in a number of ways using transistors. The scheme used in the instrument is the simplest possible, namely a transistor with its base held at a fixed potential as shown in figure 1(b). The collector load is supplied with a reasonably constant current the value of which may be altered either by changing the base voltage or the value of the emitter resistor. More complex constant current sources could be used to obtain greater accuracy, some of those recently published in *Wireless World* would be suitable, but most are difficult to vary over a wide range of currents or need a large voltage across them to function.

VOLTAGE MEASUREMENT

If the unknown resistance is low a fairly high current is used and a standard multimeter used as a voltmeter will give good results. When high resistances are being measured the voltmeter impedance must be greatly increased. This is achieved by using a number of transistors connected in the "super alpha" configuration as an emitter follower with an extremely high input impedance. This method of connection, shown in figure 2, has a major drawback in that there is a considerable voltage between the base of the first transistor and the emitter of the last transistor. This voltage is the sum of the base emitter voltages of all the transistors as defined by the I_b/V_{be} curves and of course varies in a non-linear fashion with current. This "offset" voltage may be compensated for by a fixed voltage and one varying with the voltage input, or the current taken by the transistors. A scheme to achieve this is shown in figure 3.

The potentiometer provides a fixed voltage while the variable part of the compensation is provided by the resistor R in the collector lead of the last transistor. Figure 3 also shows the component values used on the prototypes, where R was 68 Ω in parallel with 470 Ω. The first three transistors are silicon types while the output transistor is any germanium one that will pass 15 mA and dissipate 100 mW. The linearity of this circuit when carefully set up is better than 1 % from 1 V to 10 V if a stabilized supply is used, while the accuracy is much better than 1 % of f.s.d., which is as good as most ordinary meter movements. The input impedance is in the region of 50 MΩ which is satisfactory for resistance measurements of up to around 1 MΩ.

The circuit of the constant current source and range switching is shown in figure 4. This is coupled to the voltmeter section of figure 3 to form the complete instrument. The scheme adopted for the actual instrument was as follows. There are five decade ranges covering the range from 10 Ω to 1 MΩ. The lowest range will read down to zero ohms with reduced accuracy. The check/read switch is of the biased type and is normally in the check position connecting a standard resistor in place of the unknown resistance. This corresponds to the highest resistance in that range so the meter should be adjusted to read f.s.d. by altering the 250 Ω potentiometer. The circuit may be run from two stout 9 V batteries, in which case it is necessary to include the 100 Ω resistor and zener diode to keep the

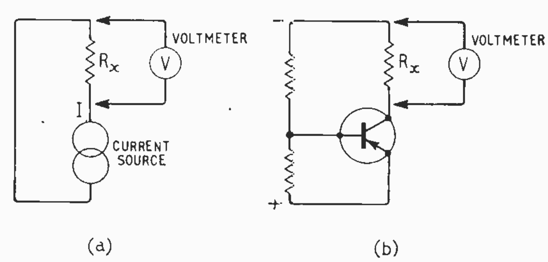


Fig. 1. (a) The measuring technique employed. (b) A transistor as a current source.

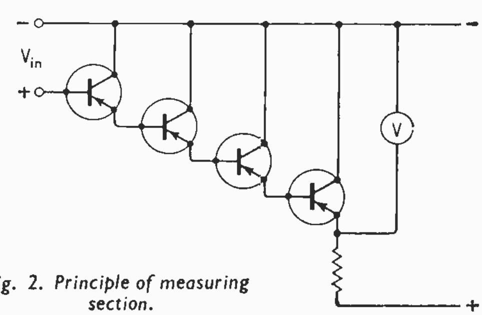
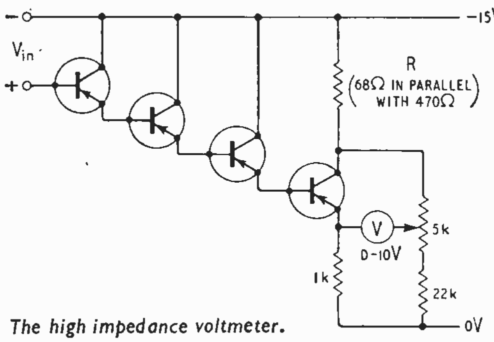


Fig. 2. Principle of measuring section.



ig. 3. The high impedance voltmeter.

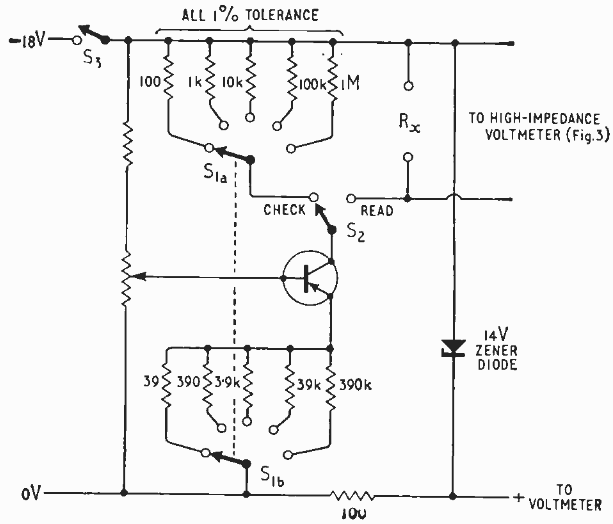
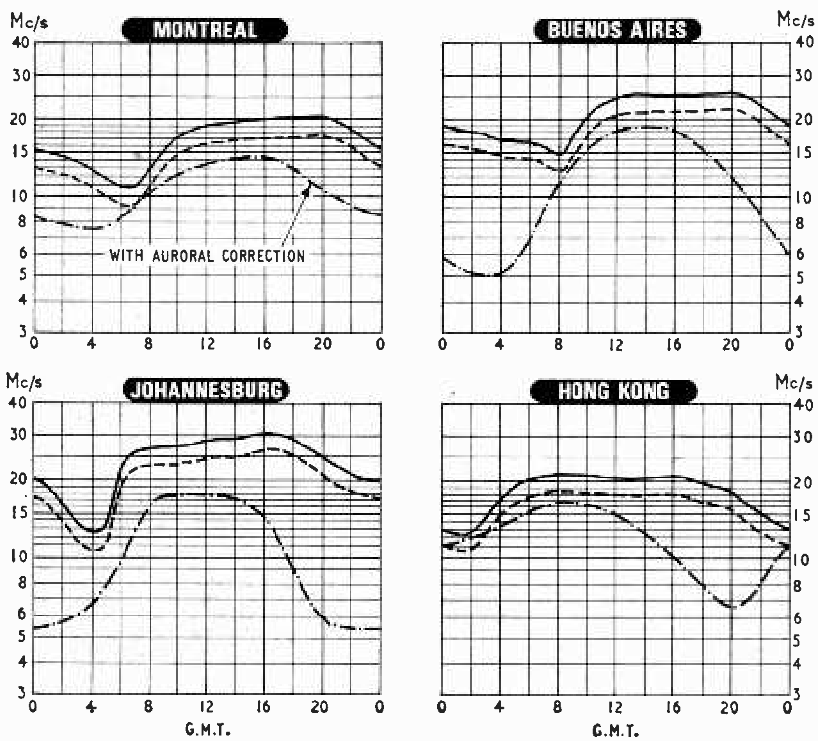


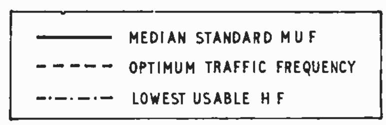
Fig. 4. (right) Constant current source and range switching.

supply voltage constant. If great accuracy is required these components may be removed and the whole circuit run from a stabilized supply of around 15-18 V.

The only setting up that is needed is the setting of the 5 kΩ potentiometer in the amplifier and any adjustment of *R* that may be necessary. The amplifier should be supplied from a variable voltage source (unknown resistors on a low range can be used for this) and the input and output voltages of the amplifier adjusted with the 5 kΩ potentiometer to be the same. If on setting this at one reading the readings at the other end of the scale are incorrect the resistor *R* should be altered to compensate for this. Once this has been done the readings over the 1-10 V range should be accurate.

H. F. PREDICTIONS—AUGUST





Daytime maximum usable frequencies (MUFs) will be 2 to 3 Mc/s higher than for the previous month, and optimum traffic frequency (FOT) curves are unlikely to be modified greatly by sporadic-E ionization.

Lowest usable frequencies (LUFs) show little change, being more dependent on e.r.p. and the type of modulation. Those shown are for commercial telegraphy using several kilowatts and rhombic aerials, but their relative proximity to associated FOTs is an indication of performance for any type of service.

The graphs, which were prepared by Cable & Wireless Ltd., show median standard MUFs, optimum traffic frequencies and lowest usable frequencies for reception in the U.K.

WW—086 FOR FURTHER DETAILS

LETTERS TO THE EDITOR

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

Pros and Cons for Using Integrated Circuits

MAY we congratulate the Rev. A. J. McEvoy on his bold approach to the design of audio circuits described in the July issue? It is always interesting, and often rewarding, to put aside the "official" descriptions of circuits and devices and to observe their behaviour in other configurations.

While designing a video mixer for a closed-circuit television system, another digital i.c. was investigated for possible use as a wideband amplifier. (It might be difficult to apply heavy signal feedback to the CD2200 because of its multi-stage design.) Although this other circuit was not used in the final version, brief details may assist anyone interested in spreading their experimental wings.

The circuit (Fig. 1) is the Fairchild μL900 Buffer (which was purchased from a recent advertiser in this journal for less than 10s). It is similar to the group of transistors Tr 6, 7, 8 in the circuit mentioned above (July p. 315). With the addition of feedback and coupling components as in Fig. 2 it is seen to be a development of the circuit variously known as the "Bowes brick," "d.c. feedback pair," etc. It is one of the most common forms of integrated wideband amplifiers, and with the component values shown, a power gain of 16 dB was obtained with a band width >25 Mc/s. It should be noted that the manufacturers optimize the circuit for pulse operation, and responsibility for any departure from this is always that of the user. We can however, say that we have used these circuits at up to four times the rated supply voltage with no apparent damage.

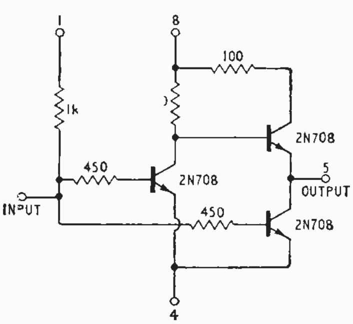


Fig. 1. Circuit of the Fairchild μL900 buffer.

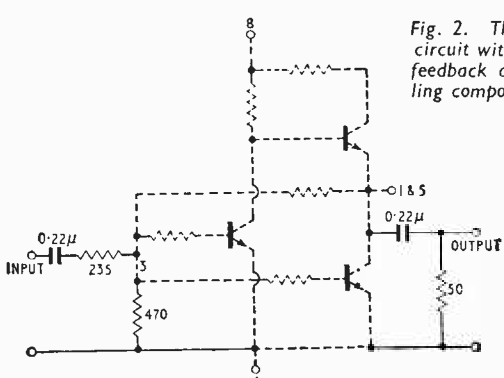
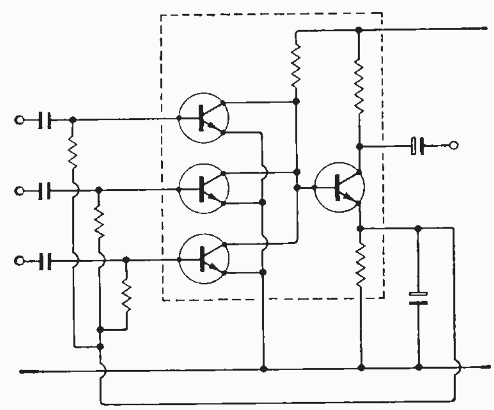


Fig. 2. The μL900 circuit with additional feedback and decoupling components.

Paisley College of Technology. J. H. EVANS and PETER WILLIAMS

THE Rev. A. J. McEvoy's adaptation of a commercial digital i.c. is attractive for its ingenuity, but it fails the acid test of economics.

One half of a discrete-component circuit which does the same job is shown in the accompanying diagram. The



portion equivalent to the i.c. is boxed. The component bill for the relevant parts in a *pair* of such circuits is:

8 planar transistors @ 3/-	24	0
6 resistors @ 6d.	3	0
	27	0

The total cost is 18s less than the maker's one-off price for the i.c. The millennium has not yet arrived.

G. W. SHORT

Croydon.

Sound Quality

APROPOS the recent correspondence on sound reproduction, my experience, starting in the early 1920s, is that design is not at fault, in either radio or television, but production (costs) cheese-paring to which the unfortunate designer's work is subjected for purely commercial reasons. Many attempts have been made by conscientious firms to make receivers which would produce quality sound output. In most cases they have failed commercially or have been forced by the economics of results to fall in line with the "average."

Quality sound, hi-fi and other similar terms are probably the most mis-leading that could have been chosen for the reproduction of sound. The quality of sound reproduction cannot be judged by frequency response curves alone. The final and deciding factor is human reaction and, since it is true to say that no two human beings react similarly to any set of circumstances, it is virtually impossible to set up a "standard" for "quality sound."

Since the great mass of television and sound receivers are used in normal dwelling houses, mostly filled with all types of bric-a-brac which resonate at their own particular frequencies, it is almost impossible to reproduce "high quality" sound in the average lounge or living room. By far the greater problem is the human element in control of the equipment. The "boom" produced by full top cut is beloved by the great majority, in my experience. The "pop" addicts must have "screech" and volume. "Groups," specialize in sound levels approaching and beyond the "pain" level. But, let's face it, these are the people who keep manufacturers in business.

It is only in recent years that any standard for loudness could be decided on. The much more complex subject of quality still defies definition since it is governed largely by the individual's reactions. The human voice, orchestra or solo instrumentalist which is aesthetic to the senses of one person is anathema to another.

To sum up the whole question of "quality sound" reproduction, there is no doubt that present engineering practice can produce sound equipment capable of a very high standard of reproduction; but engineers must also remember that there is no known means, at present, of measuring what the artistic interpretation of the resultant output is when analysed by the human ear and brain.

Finally, the testing of equipment in anechoic chambers is, academically, excellent, but where can those conditions be repeated in the normal dwelling house in order to iron out the "bugs" which are normally there? The "good quality" in the days of earphones, of which we were so conscious, was most probably largely contributed to by the fact that practically all extraneous damping and resonances were eliminated.

JOHN D. BENSON

Bedford.

Electronic Devices

"VECTOR" hardly went far enough in his July contribution. One of the greatest difficulties manufacturers have is that other people make equipment which is alleged to do what is wanted, but is in fact designed to do something else.

The optical people are pretty bad in this way, but one outstanding example exists in electronics applied to photography. Photographic exposures should be graded in the camera or enlarger on a log scale—intervals of sixth-root-of-two are as convenient as any (an electronics man might call this 1 dB, though what connection time and decibels have with each other is mysterious)—but with one exception the timers on the market have arithmetic scales. This means that students have to be taught to work out a set of log times and chalk them up on the wall behind the enlarger so that they can usefully use a timer that is supposedly designed to time exposures adequately.

The one exception to this is a German device, which does give "1 dB" intervals on times between 1 and 200 seconds. Unfortunately, it is not sufficiently stable in use to give values better than plus or minus 5%, and if one continually uses a single time setting, then the time dispensed will increase systematically.

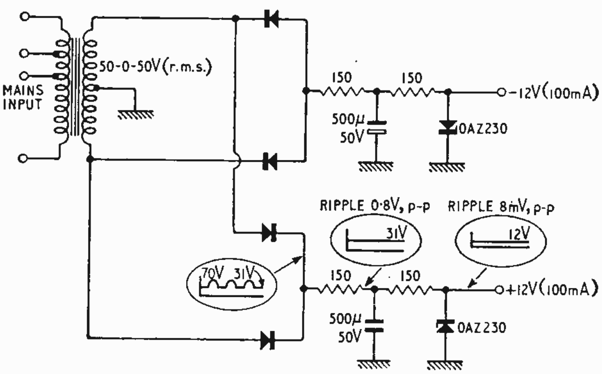
Part of the difficulty here, of course, is that if any timer were not described as being "transistorized," customers would think it was no good. It is remarkable how a bellpush, a Leclanché cell and a bell have developed into an electronic signalling system. In my youth we called them bell wiring.

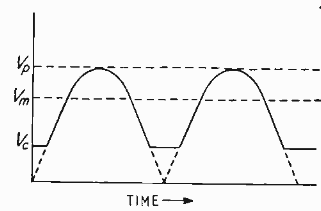
P. C. SMETHURST

Bolton, Lancs.

Low-voltage Power Supply

I RECENTLY required a power supply to give equal positive and negative outputs while maintaining a low output impedance at normal operating currents and which could also be short-circuited without harm, and, in addition, recover without external aid. The circuit which was developed and may be of interest to readers is shown in the diagram. Additional features are the low





V_c AS % OF V_p	V_m AS % OF V_p
0	64
20	65
48	71
64	78
78	85
89	91
96	97
100	100

ripple current flowing in the smoothing capacitors and the low diode peak currents.

The Table gives the mean voltage (V_m) at the filter input for different values of V_c, the capacitor d.c. voltage. The voltage across the capacitor V_c forms the limit at which the diodes cease conducting. Ideal diodes and transformer are assumed.

A. SANDHMAN

Harlow, Essex.

Producing Printed Circuits

I WOULD like to thank Mr. P. St. J. R. French for his rapid method of producing printed circuit layouts given in his article on an audio indicator in the June issue and in return to offer a suggestion of my own which I have found very useful.

When the printed circuit board has been prepared by whatever means is used, and after drilling has been carried out, I clean the copper surface with fine steel wool and then spray liberally with Messrs. Holts "Dampstart" (usual disclaimer). This forms a coating which prevents oxidation of the copper but is no barrier to soldering. If the equipment is to be used in a damp atmosphere then a further coating with "Dampstart" after mounting the components is very helpful. This can be given to both sides but if this is done care must be taken to protect any adjustments, as otherwise they will become gummed up.

M. C. MATTHEWS

Dorchester.

Novel Tone Generator

THE article by Mr. French in your June issue brings to mind a circuit I used as an engaged tone generator a few years ago when transistors were a rarity. I prefer to call it a "trivibrator" but it has also been referred to as a donkey simulator.

Three transistors are arranged in a triangle (electrically) and each side of the triangle is wired as an astable multivibrator. (See Fig. 1.) The approximate frequencies of the sides A-B and A-C are respectively 700 c/s and 500 c/s while the B-C side "ticks" at about 1.5 c/s.

Consider the B-C multivibrator. When B is off then A-C can oscillate at 500 c/s, and when C is off then A-B can oscillate at 700 c/s. The output, taken from the collector of A is a rectangular waveform alternating between 500 and 700 c/s.

As the feedback path round the triangle would effectively short the h.f. cross coupling, the l.f. multivibrator is coupled from the collectors of B and C to a tap on the respective base resistors. The final circuit is as shown in Fig. 2.

The output is now fed to a 150-Ω earpiece similar to that described by Mr. French, and the device is currently serving as an alarm clock sounder.

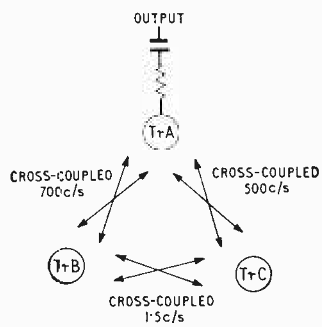
Any supply from 3 to 15 V will suffice and a flat battery makes the output sound even more interesting!

The transistors were bought as "red spot audio types" and β is approximately 30-50.

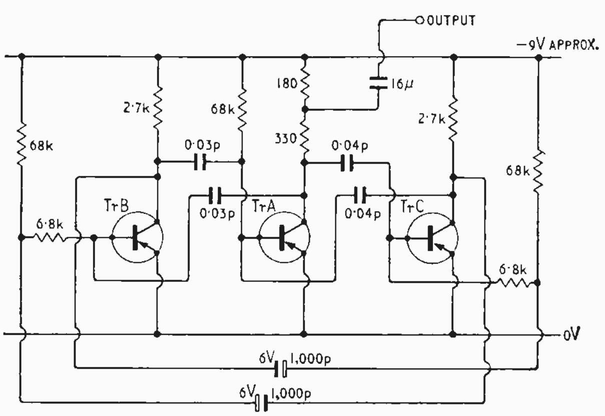
The device has frequently been used as an identification generator for checking wiring, as the signal it produces cannot possibly be confused with any other tones which may be around ones equipment.

A. B. BLACKWELL-JONES

Evesham, Worcs.



(Right) Fig. 1. The "trivibrator" described by Mr. Blackwell-Jones.



(Below) Fig. 2. The complete circuit.

Ohms per Square

WHEN dealing with problems concerned with the resistance of sheet materials, e.g. metal foils, resistance papers (Teledeltos), etc., the resistance of the material is frequently given in "ohms per square."

For the benefit of anyone puzzled by this expression and tempted to ask "per square what?" I would explain the reason for the answer being simply "ohms per square" as follows:—

The resistance R of any electrical path is given by the expression

$$R = \frac{\rho L}{A}$$

where ρ = specific resistance of material
L = length of path
A = cross section area of path.

In the case of sheet material of sensibly uniform thickness

$$A = \text{Width } (W) \cdot \text{thickness } (t)$$

And for a square

$$R = \frac{\rho L}{Wt} = \frac{\rho}{t} \text{ since } W = L$$

and hence R is independent of the size of the square, and for practical purposes will be constant for any given material.

E. ELLIOTT

Rochdale, Lancs.

Active Element Terminology

BEING afraid that someone may suddenly introduce yet another active circuit element and invent a new terminology for its various ports, may I make a belated plea for retention of the original "electrodes" of the thermionic triode (as indeed has been done for semiconductor diodes), abandoning the term "grid" which was descriptive of appearance rather than function, and substituting "archode" for that famous third electrode.

This move would not only give f.e.ts, transistors, metal rectifiers, diodes and thermionic triodes a common terminology, but could even bring Fluid Logic into the fold, with the difference that the collective "-odes" would rightly be called "pneumodes."

This proposal may not be new—"Free Grid" must surely have had something to say about it, even despite the danger of spoiling his *nom de plume*—but I sincerely hope that the restatement will encourage its acceptance.

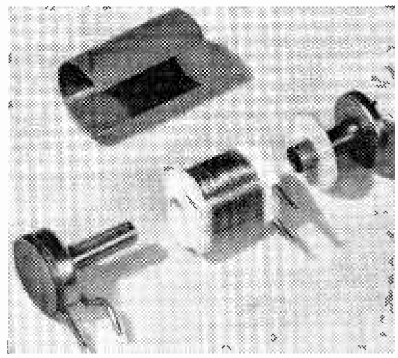
D. A. G. TAIT

Farnham,
Surrey.

NEW PRODUCTS

Diaphragm Relay

FLEXIBLE metallic diaphragms are employed as moving contacts in the STC diaphragm relay 1M. It is less than 10.5 mm high when mounted on a p.c. board, and has a switching capability of 30 W, maximum current being half an amp, while the maximum voltage can be either 150 V d.c. or 250 V a.c. Tests indicate that it conforms to DEF5011 vibration and bump conditions, at −65°C and +100°C. This specification is said to be a result of the low mass of this moving contact, and the fact that the contacts of this relay are in a separate hermetically sealed switching element, containing a non-oxidizing gas. The switching element (looking like a flat headed rivet in the illustration) is a circular steel enclosure in which the diaphragm is held close to a central fixed contact formed by the end of a cylindrical rod. This protrudes to act as a core for the relay coil. When the coil is energized, the diaphragm is attracted to the fixed contact, and makes firm contact over a wide area, (at least 5 sq. mm). Normal excitation creates a contact force of about 25 gm: restoring force is 10 gm. Operating time is 1.5 ms

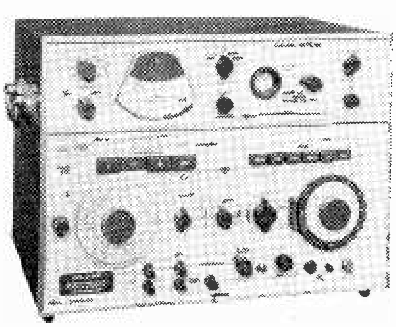


including contact bounce, and release time is of the order of 500 μs. Mating surfaces are gold coated, and typical contact resistance is 30 mΩ. Operating life when switched 60 V d.c. at half an amp (resistive) is said to be at least one million operations, rising to five million at 60V 100 mA. The 1 M can be produced in multiple forms. Overall dimensions are less than 1.5 cu. cm. STC Electro-Mechanical Division, West Road, Harlow, Essex.

WW 301 for further details

Versatile Inductance Bridge

INDUCTANCE values from 0.3 μH to 21 kH can be measured with the new Marconi Instruments Inductor Analyser type TF2702. A c.r.t. indicator is used for preliminary balancing, final balance being indicated on a moving-coil meter. The meter also reads the applied a.c.



inductor voltage. The c.r.t. indicator can also be used to display the overload characteristic of inductor cores.

The current rating of 10 A extends only to 21 H, at 3 A to 210 H, at 0.3 A to 2.1 kH and at 30 mA to 21 kH. Voltage of 500 V is available at 10 H. The detector permits an accuracy of −1% for frequencies up to 3 kc/s and is −1.25% up to 20 kc/s. Frequencies of 50 c/s, 1 kc/s and 10 kc/s are available internally and other frequencies can be supplied by external means. High power measurements are possible with addition of an a.c./d.c. mixer unit TM8339. Inductors can be measured in the series or parallel mode, thus eliminating the need for a calculation using f and Q. Marconi Instruments Ltd., St. Albans, Herts.

WW 302 for further details

SOLAR CELL POWER SUPPLY

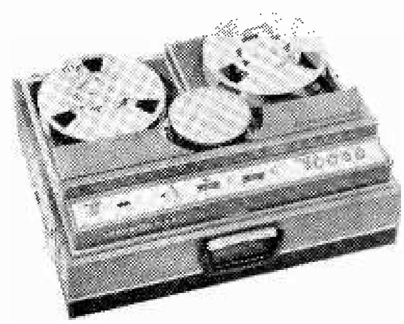
TWENTY semicircular solar cells are packed into a heat resisting acrylic resin case to form the Type S 224 solar cell power supply introduced by Photain Controls Ltd. All cells are mounted on a shock-proof printed circuit board. Anti-reflective coatings are used to prevent loss of sunlight, and the electrical contacts of each cell are made of treated alloy to reduce losses due to internal series resistance. Output current varies linearly with illumination intensity up to a maximum of over 90 mA with an optimum operating current of 86 mA. The output voltage remains almost constant irrespective of changes in illumination intensity. The unit has an optimum operating voltage of 4·2 V, with a maximum of over 5·5 V. Operating temperature range for continuous use is from −50 °C to +75 °C. Maximum output power is over 360 mW. Photain Controls Ltd., Randalls Road, Leatherhead, Surrey.

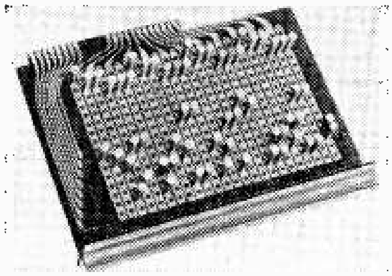
WW 303 for further details

Video Tape Recorders

TWO Optacord video tape recorder models, the 600 and the 600S, have a playing capacity of one hour, and are similar in style to an audio tape recorder. Warm-up time of three seconds, and high speed rewind (two minutes) are also common to both models, but in addition, the 600S has still-picture and slow-motion play-back facilities. On both these models, by Loewe-Opta, the video frequency response is 10 c/s to >2·5 Mc/s, and the audio frequency response is 50 c/s to 10 kc/s. There are two inputs which are switchable for television picture/television sound, and camera picture/microphone. The video output is 0 to 4 V pk-to-pk at 75 Ω, and the audio output is 0 to 1 V at 200 Ω. The recording system used is a one-head system. Price of the 600 is £850, and the 600S is £1,025. Highgate Acoustics, 184/188 Great Portland Street, London, W.1.

WW 304 for further details





Matrix Programming

A double sided version of Veroboard with a larger hole matrix is introduced by Vero Electronics Ltd., Industrial Estate, Chandler's Ford, Hants, for programming purposes. Described as the D.E.B. matrix programming system, it is intended for such applications as sequence controllers, analyser data loggers, machine tool control, process controllers, multiplexers, and automatic test equipment. This board is flush mounting, plug-in, and cordless. There are screw-in component holders, and shorting pins.

WW 305 for further details

POWDERED RESINS

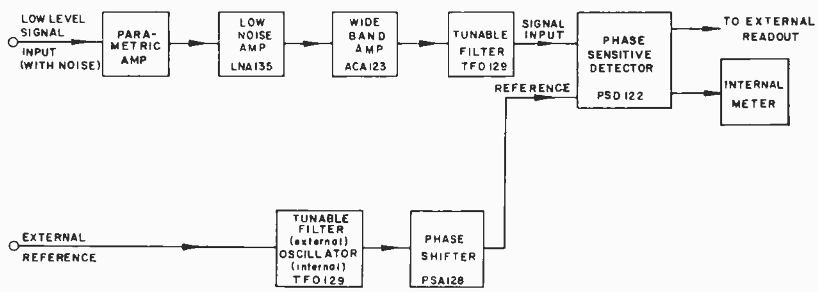
IMPREGNATING and encapsulating coils and windings is now possible by resistance heating of powdered epoxy resins. "Scotchcast" powdered resin 265, is a clear, low melt viscosity, one-part, rapid curing material. It is intended to replace conventional varnish and trickle impregnation resins for wire anchoring and holding, even at high revolutions. The heat generated when a current is passed through the coil is used to melt and cure the powdered resin. No masking of parts is required, since only the copper coil is heated and this retains the resin. Residual powder is blown off the unheated parts. Current requirements of 5 to 40 A for 15 to 20 seconds are said to be all that is needed for most small coils and windings. Minnesota Mining and Manufacturing Co. Ltd., Wigmore Street, London, W.1.

WW 306 for further details

Coherent Detector

SYSTEM 5.1 has been designed to recover signals of known frequency from a background of interference such as noise. Of modular construction, it is based on a coherent detector which is extremely sensitive—three nanovolts at 10 kc/s or 500 pV when a parametric amplifier is used. The modular design of this equipment, by AIM Electronics, 8 North Street, Cambridge, allows the user to build units of required sensitivity. In some applications it is also described as a lock-in amplifier, and phase-correlating voltmeter. System 5.1 has been used for infra-red spectroscopy, radio astronomy, measurement of motor winding temperatures, and Hall-effect measurements in liquid metals. The six basic modules may also be connected for use in harmonic analysis, plotting Nyquist diagrams, or as an a.c. servo test set. In this coherent detection system, an input waveform is amplified, and mixed with a reference signal of the same frequency, which chops both incoming signal and noise. Following an integration period, a d.c. output signal is obtained whose amplitude is proportional to the basic signal. A phase shifter is used to produce a maximum output—this occurring when input signal and reference are exactly in phase. The AIM instrument has a special tunable filter—for the input signal—which rejects most of the incoming noise, and thereby prevents the mixer from being overloaded.

WW 307 for further details



Digital Integrated Circuits

FIVE digital integrated circuits are available from Texas Instruments as additions to the Series 74 range of t.t.l. circuits, for digital computer and data handling systems. The SN7482N dual adder carries out the addition of two 2-bit binary numbers, and it is intended for high-speed, multiple-bit, parallel/add serial-carry applications. The speed is 15 ns carry time through two stages, and add time is 35 ns. The SN7483N quadruple full adder performs the addition of four 2-bit numbers. Carry time is 30 ns through four bits, and add-time is 35 ns. For frequency division by 12, 6, 3, or 2 there is the monolithic SN7492N divide-by-12 counter. Operating frequency is 15 Mc/s for this unit which consists of four master/slave flip-flops internally connected to provide a divide-by-six counter section, and a divide-by-two counter section. The two sections can operate independently and simultaneously. A four-bit binary ripple-through counter is another high-speed monolithic unit (SN7493N) consisting of four master/slave flip-flops, this time interconnected so that it functions as a divide-by-eight counter, and a divide-by-two counter. When externally connected, the two sections provide ripple through operation by dividing the output of each flip-flop by two, thus achieving division of the input frequency by 2, 4, 8, and 16. The operating frequency is 15 Mc/s. For driving gas-filled readout tubes, the SN7441N is a b.c.d. to decimal decoder. It consists of t.t.l. gate circuits which select one of the ten decimal output drivers. The ten high-breakdown (65 V) output transistors satisfy the collector-emitter characteristics specified by indicator tube makers. Texas Instruments Inc.

WW 309 for further details

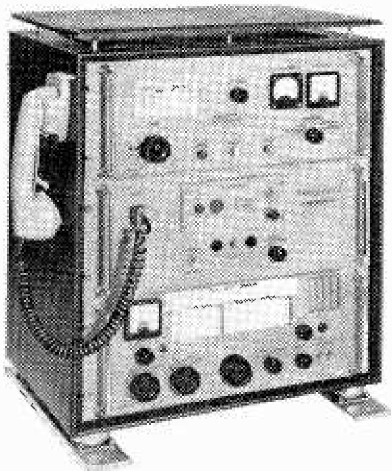
Oscilloscope Converter

THE Aerometrics Series 300 Sampler is stated to convert any low-speed oscilloscope into a high resolution sampling system. Made in the U.S.A., this instrument is a solid-state, dual-channel, pulse sampler, with rise times of less than two nanoseconds and a sensitivity of two millivolts to 500 millivolts in eight ranges. Horizontal sweep speeds are 10 ns to 5 μs. By using this sampler with an oscilloscope it is said that rise times, peak voltage, and decay times of transistors and diodes can be tested and measured, and the automatic testing of memory cores carried out as well. Low sweep speeds provide an output for driving an XY plotter. Wessex Electronics Ltd., Royal London Bldgs., Baldwin St., Bristol 1.

WW 308 for further details

Marine Radiotelephones

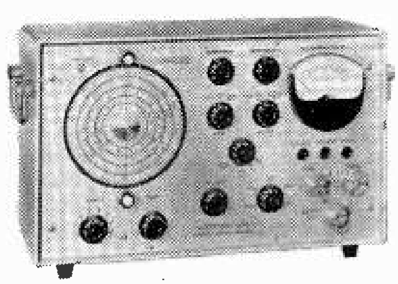
FULLY type tested and approved (G.P.O.) radiotelephone equipments for operation in the 1600 to 3800 kc/s frequency range are offered by Coastal Radio Ltd. Two transmitters and two receivers for a.c. or d.c. operation are the Cresta and Altair and Vega and Centaur respectively. The Cresta transmitter (illustrated) has a 100 W output into the aerial, with low power 10 W for local working. Except for the power amplifier and master oscillator valve stages, the unit uses solid-state devices. Simplex or duplex working is available, and there are 11 crystal-controlled transmitting channels. When the transmitter frequency is changed, the output tuning is carried out automatically. An automatic two-tone distress sender is incorporated. Heat dissipation is by natural ventilation. The Vega receiver (illustrated) is completely solid-state, and has 22 preselected crystal controlled spot frequencies in the R/T band. Two-speed manual tuning can be carried out on the telephony, navigational, long- and medium-wave broadcast bands. There is a b.f.o. for c.w. and beacon reception, and there are D/F facilities. Gain control of a.f. and r.f. is provided, and a wide range a.g.c. can be switched in when required. The Altair transmitter has a 75 W output, is fully solid-state, and has ten crystal controlled channels. The Centaur receiver has five pre-selected crystal controlled spot frequencies in the R/T band, and D/F facilities. It is stated that this range of equipment is suitable for the two classes of vessel aboard which the installation of R/T equipment is or will be mandatory by May 1967 and May 1968. Coastal Radio Ltd., Fleets Lane, Poole, Dorset.



WW 310 for further details

A.M. Standard Signal Generator

PRINTED circuits and solid-state devices have been used throughout the Westechno Model 28 laboratory-standard signal generator. R.F. ranges are 30 kc/s to 30 Mc/s in eight overlapped bands on a directly calibrated dial. The r.f. output is 0·1 μV to 100 mV attenuated, output impedance is 75 Ω at r.f. socket, and the output voltage accuracy is ±2dB. Frequency accuracy is ±1% without crystal check and the s.w.r. is less than 1·5-1 overall. The crystal oscillator provides check points over two range settings; every 10 kc/s throughout the entire range, and every 100 kc/s on ranges 2 to 8 (100 kc/s to 30 Mc/s), with an accuracy of ±0·01% or better. There is also provision on the front panel for calibrating the crystal against external frequency standards to 0·0001%. Critical cursor adjustment for crystal checks has a cursor lock. Internal a.m. is achieved by an a.f. oscillator producing a continuously variable 400 to 800 c/s (sinusoidal). Frequency accuracy is ±10% of scale indication. Internal modulation is variable from 0 to 100% and indicated on a panel meter; there are also provisions for external modulation monitored by a meter. Internal power supplies are mercury cell batteries in series (6·75 V and 5.40 V), with a working life of about 80 hours. External supplies required are 12 V d.c. Stabilization of voltage sensitive circuits is by Zener diodes. Protection is provided against wrong polarity connections of internal or external supplies. Price is £125. Westechno Ltd., Instruments Division, Estuary House, Camperdown Terrace, Exmouth, Devon.



WW 311 for further details

TRIMMERS

HELICAL trimming potentiometer 79P, ¾in long, is from Beckman Instruments Ltd. This 15-turn unit is fully sealed to meet immersion tests of MIL-R-22097C. Available from stock in standard resistances from 10 Ω to 12 MΩ, it has a total resistance tolerance of ±10% in most values. The power rating is 0.75 W at 25°C. The Cermet element permits setting to within ±0.05% of a required value. It will withstand power surges five times its rated wattage. The 79P measures 0.75×0.19×0.36 in. It has moulded plastic housing and cover, stainless steel adjustment screw, and gold plated terminal pins. Beckman Instruments Ltd., Queensway, Glenrothes, Fife, Scotland.

WW 312 for further details

U.H.F. SWITCHED ATTENUATOR

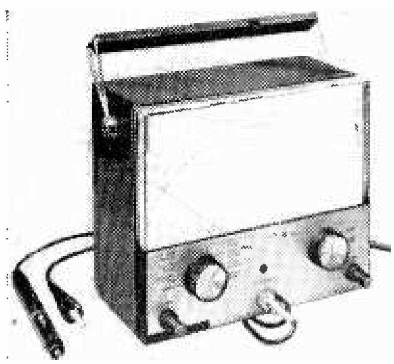
IT IS stated that the accuracy and fredom from standing waves to be found in the u.h.f. attenuator TF 2163 are the results of employing special resistive pads that are selected by cam-operated microswitches. This step attenuator by Marconi Instruments Ltd., operates over the frequency range d.c. to 1 Gc/s. The characteristic impedance is 50 Ω, and it has an attenuation range of 142 dB variable in 1 dB steps. This selection is achieved by two rotary step controls, one covering 120 dB in 20 dB steps, and the other covering 22 dB in 1 dB steps. The incremental attenuation is given by the sum of the two readings. The physical construction has been designed to offer a v.s.w.r. better than 1.1:1 up to 200 Mc/s; 1.25 up to 500 Mc/s and 1.5 up to 1 Gc/s. The price is £148. Marconi Instruments Ltd., St. Albans, Hertfordshire.

WW 313 for further details

LOW COST V.V.M.

WITH a seven-inch scale the tracking error of the Bach-Simpson valve voltmeter 312 is stated to be less than 1% and the accuracy on all a.c. and d.c. ranges to be ±3%. Input impedance is 16 MΩ, and the frequency response is extended to 250 Mc/s when the r.f. probe is used. A low voltage d.c. range permits precise measurements below 0.5 V levels. Both a.c. and d.c. voltage measurements can be made up to 1.5 kV, and with d.c. the range can be extended to 30 kV by using an external high voltage probe. Price is £35. Bach-Simpson Ltd., 19 Nortoft Road, Chalfont St. Peter, Bucks.

WW 314 for further details

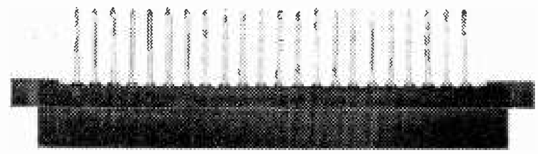


Strobing Voltmeter

FOR making measurements on fast waveforms at precisely located points on the time axis the EH153 strobing voltmeter is offered by E. H. Research Laboratories, U.S.A. Single-shot measurements of better than 1% accuracy can be made in a four nanosecond-wide window. Equivalent bandwidth is better than 100 Mc/s, and the noise level is less than 0.5 mV peak-to-peak. Delays may be programmed manually, or by external analogue voltage or by digital input. This instrument can convert a low frequency oscilloscope into a sampling 'scope. Capable of 1000 measurements a second, this essentially single-shot device has an output voltage available for a period of 1 ms, which was deliberately chosen to allow use of relatively slow data-handling equipment to accept measurement information. The amplifier-memory chain is d.c. coupled for stability. Marketed in U.K. by Instrument Division, Livingston Laboratories Ltd., Livingston House, Greycaine Road, North Watford, Herts.

WW 315 for further details

Wire-wrap Edge Connectors



BELLOWS-FORM contacts, which provide a firm and even grip over the entire contact area and appreciably reduce the wear on precious metal due to the low force required for insertion and withdrawal, are fitted to the Ultra series 864 precision-made connectors. And, because of the low spring rate employed, they can "ride" severe climatic, shock or vibration conditions—even when used with uneven or bowed printed circuit boards. All mouldings are manufactured from glass-filled diallyl phthalate (DAP) and the contacts, made from spring-tempered phosphor bronze, can be plated gold flash or heavy gold. Series 864 wire-wrap edge connectors are designed on 0.200 in (5.08 mm) contact centres and can be supplied for single or double-sided application. There is the added facility of easily removable/replaceable contact/termination assemblies. Standard number of ways available: 8, 16, 24 and 32, but non-standard types can be supplied by special request. Nylon end fixing feet are available for attaching connectors to a chassis or racking system, except for the 32-way which incorporates end fixing. Ultra Electronics (Components) Ltd., 419 Bridport Road, Greenford, Middx.

WW 316 for further details

YUGOSLAV MULTIMETER

A MULTIMETER of Yugoslav manufacture is being marketed in this country by Guest Electronics Ltd., 78-86 Brigstock Road, Thornton Heath, Surrey. This instrument, the S.6A, uses a series of input terminals for selecting the various measuring ranges rather than a rotary switch. Accuracy is said to be ±2.5% f.s.d. Sensitivity on d.c. ranges is 20 kΩ/V, and on a.c. ranges it is 4 kΩ/V. There are seven d.c. voltage ranges from 100 mV to 1 kV, and six current ranges from 50 μA to 5 A. A.C. voltage measurements cover 2 V to 1 kV in five ranges, and there is one current range 0 to 250 μA. Resistance can be measured up to 100 MΩ in six ranges. Capacitance values from 100 pF to 150 μF can be read off in four ranges. The requency ranges are 0 to 50 c/s, to 500 c/s and to 5 kc/s (with 5,000 pF capacitor). Meter movement protection against overload is provided by two silicon diodes. The price is £7 15s.

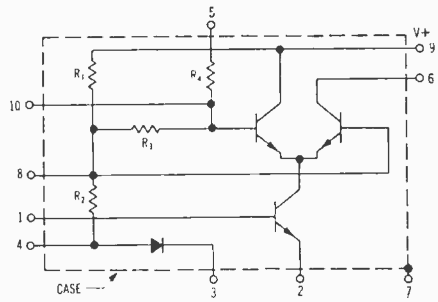
WW 317 for further details

Digital Tachometer

AVAILABLE with four or five digit display, the rack mounting versions of the Digicron 668 series of electronic digital tachometers can be supplied with either an oven temperature-controlled crystal timebase or a mains frequency derived timebase. There are four gate times of 1, 6, 10 and 60 seconds, permitting a direct r.p.m. indication with 1, 6 10, or 60 pulses per revolution. Alternative gate times are available to meet special applications and a wide input range between 100 mV and 300 V allows them to be used with most types of pick-up transducers. The display consists of Mullard 13 mm numerical tubes, and the actual display time (automatic) can be varied from 0.5 to 10 seconds, or it can be controlled manually by front-panel push-buttons. The accuracy of the timebase when mains derived is usually within 0·25%; when crystal controlled it is ±0·001% at 25°C. Darang Electronics Ltd., Restmor Way, Hackbridge Road, Hackbridge, Surrey.

WW 328 for further details

Consumer Linear I.C.



A MULTI-PURPOSE linear integrated circuit costing 26s for 100-up is available from Motorola distributors. The circuit (shown) has numerous applications at audio, video, intermediate and radio frequencies, e.g. as a tuned amplifier, modulator or mixer and oscillator. Power gain is 30 dB at 60 Mc/s with a noise figure of only 5 dB. Good stability can be had due to the low internal feedback.

First impressions of the circuit can be misleading. The lower transistor is not intended as a constant-current source but as a common emitter amplifier, this feeding a common base amplifier (right), thus forming a cascode arrangement. The additional transistor (left) is intended for automatic gain control, shunting the signal path while maintaining the input transistor at its operating point. This keeps the input impedance constant over the a.g.c. range and so does not affect tuning greatly. For this kind of operation pins 4, 8 and 10 should be by-passed to earth, the input signal applied to pin 1 (<100Ω source resistance), a.g.c. input applied to pin 5, the output taken from pin 6 and pins 2, 3, 7 and 9 earthed.

The amplifier is designated MC1550G. Motorola Semiconductor Products, York House, Empire Way, Wembley, Middx.

WW 319 for further details

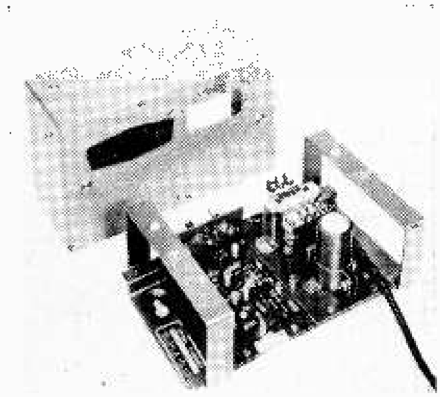
X-Y Data Reader

DIRECT digital read-out of x and y co-ordinates of charts, graphs and film records is provided by the P.C.D. Ltd., X-Y Digital Data Reader. Intended for graphical analyses, the output from this instrument may be read from a digital voltmeter or a computer. Continuous records of up to 200 ft long may be loaded on to the instrument for analysis, and charts, graphs or maps up to 30 × 50 cm ($11\frac{1}{4} \times 19\frac{1}{2}$ in approx.) can be accommodated. A Z-axis provides three-dimensional volumetric measurement facilities, and allows multi-channel records to be referred to a common base line. P.C.D. Ltd., 219 Sycamore Road, Farnborough, Hants.

WW 320 for further details

Broad-band Aerial Amplifier

INTENDED for small aerial distribution systems the L1812 is a self-contained v.h.f. amplifier with its own power pack, and covers continuously the frequency range 40-220 Mc/s. (Bands I, II and III.) The basic amplifier is broad band, but facilities are provided to enable the installer to use the equipment either as a single input, broad-band amplifier, or as a three-band amplifier. The change is made to the input circuit by plugging the appropriate filter network into the connector provided. Two filter networks are available, the first provides the single input signal for wide-band facilities, the second is a triplexer to allow the use of separate aerials for Bands I and II and III. The power pack is constructed on a separate printed circuit board mounted alongside the main amplifier. The metal-cased unit is wall mounted by means of two keyholed slots in the base of the amplifier. The gain is 24 dB measured in a 75 Ω system, and the output is not less than 100 mV r.m.s (peak white). Power requirements are 210-250 V, 50 c/s, 5 VA. Belling & Lee, Ltd., Great Cambridge Road, Enfield, Middlesex.

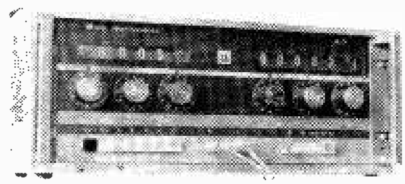


WW 321 for further details

Automatic Bridge

MEASUREMENT of capacitance, conductance, and dissipation factor can be undertaken with the three-terminal automatic bridge K1 from Mark Instruments Ltd., Portsmouth Road, Esher, Surrey. Constructed with extensive use of integrated circuits (both logic and linear types), the instrument has a capacitance range of 0.001 pF to 1.1 μF. Conductance ranges are from 10 pmho to 100 mmho, and the dissipation factor can be measured from 0.00001 to 1.00000. The accuracy of the conductance ranges is 0.025% and for dissipation it is 0.25%. A test frequency of 1 kc/s and d.c. bias are both applicable to the unknown, and are variable so that semiconductors can be measured. The price is £950.

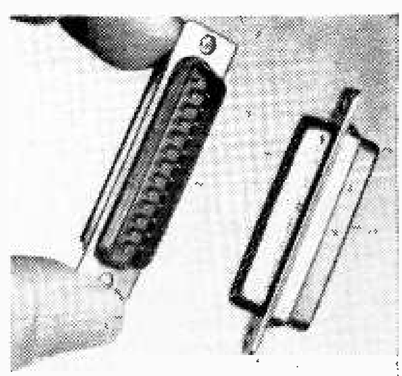
WW 322 for further details

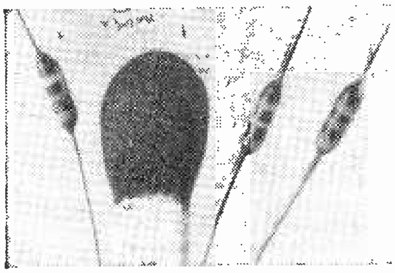


MINIATURE CONNECTOR

CONTACTS are removable crimp type and rated at 5 A on the Souriau type 8630 miniature rectangular connectors made in France. There are five shell sizes available, with 9, 15, 25, 37, and 50 contacts respectively. Voltage rating is 1 kV test at sea level, with insulation in diallyl phtalate. Operating temperature range is −55°C to +125°C. The contacts can be inserted and removed from the rear by a plastic tool. Agents are Lectropon Ltd., Kinbex House, Wellington Street, Slough, Bucks.

WW 323 for further details





Microminiature Resistors

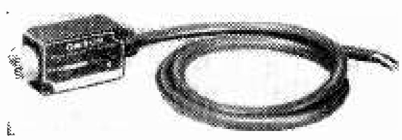
HIGH-stability film resistors type Rsx 00 are available in values up to 4.7MΩ from G. A. Stanley Palmer Ltd., West Molesey Trading Estate, Surrey. Only 2.5 m long and 0.8 m diameter, they are produced in I.E.C. preferred ohmic values series E6 (20%), and series E12 (10%). These microminiature resistors are rated at 0.05 W at 40°C, derating to 0.02 W at 70°C. The noise figure for all values is said to be better than 3 μV/V.

WW 324 for further details

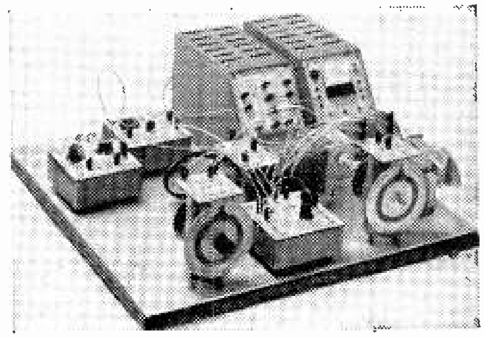
PROXIMITY SWITCH

INDEPENDENT of external power supplies, the YL2GPA proximity head operates directly from 230 V 50 c/s mains supply. The solid state circuitry is contained in a sealed case, and is splash and dust proof. It is capable of sensing ferrous metals up to 5.5mm away, and will provide an output sufficient to drive a relay. Omron Precision Controls Ltd., 313 Edgware Road, London, W.2.

WW 325 for further details



SERVO SYSTEM



WITH the MS 150 modular servo system, the student or trainee can check each of the discrete components before assembling them into a complete system. Made by Feedback Ltd., Crowborough, Sussex, each of the modules is labelled with a mimic diagram, indicating its function, such as amplification, summing, attenuation, etc. Self contained, except for low voltage power supplies, these modules emloy transistors for active circuit elements. Theory and experimental work, leading to comprehension of feedback techniques, system compensation, and performance analysis, are said to be possible with the MS 150. Basically it is a d.c. system intended to demonstrate speed and position control. It will also be possible to exchange the d.c. error channel for synchro elements and a demodulator, giving an a.c. (50 c/s carrier) error channel. With certain auxiliary apparatus, frequency and transient response studies can be carried out. The price is £275.

WW 326 for further details

T.W.T. Low-Power Supply

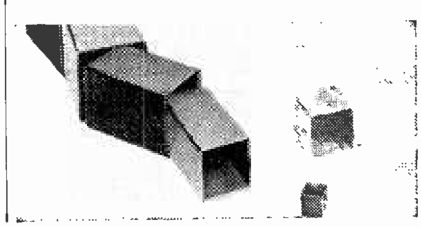
FOR travelling wave tubes of the low noise, low power type, Microtest Ltd. have designed a bench power supply. Type 615 provides continuously variable, regulated 600 V to 1 kV collector (beam) voltage and a variable helix voltage from 300 V to 800 V. Variable supply voltages for grids 1, 2, 3 and 4 are 0 to −20 V, 20 to 150 V, 50 to 300 V, and 0 to −650 V. The solenoid (electromagnetic focusing) current supply is ten amps maximum at 12 to 26 V. Correct switch-on sequence is ensured by special interlocking devices, preventing misapplication of voltages to the collector and other electrodes unless the solenoid current is flowing. Voltages are read on one meter and collector, helix, and solenoid currents are read on another, internal meter. Microtest Ltd., 9 Old Bridge Street, Kingston upon Thames, Surrey.

WW 327 for further details

Encapsulating Shells

ENCAPSULATING shells for components and sub-assemblies by Milton Ross Co. Ltd. cover sizes from $\frac{3}{16}$ to $2\frac{5}{8}$in square in increments of $\frac{1}{16}$in, and in heights from $\frac{3}{16}$ to $2\frac{1}{2}$in. Featuring a recessed shoulder which accepts a firm mating header, the units are moulded in diallyl phthalate (DAP). This material offers good adhesion with any encapsulant, has high volume resistivity, good dielectric strength, and can operate continuously at temperatures up to 210°C. All requirements of MIL specification M14F, type SGD are met using these shells. The thickness of the shell wall provides the encapsulated component with complete moisture protection. A wide range of colours is available. Milton Ross Co., Ltd., 14 New Road, Watford, Herts.

WW 329 for further details



SPECTRUM ANALYSER MODULE

MODEL VR-4 module for spectrum analysis covers 1 kc/s to 27.5 Mc/s and has applications in the field of c.w., a.m., f.m., pulse and random waveform analysis. It is particularly useful for telephone carrier work, where missing channels, carrier or signal leaks, and other faults can be detected at a glance. Available from Livingston Laboratories Ltd., Greycaine Road, North Watford, Herts., this instrument covers 0 to 25 Mc/s in a single pre-set sweep with continuously adjustable sweep widths from 50 kc/s to 5 Mc/s. Other outstanding features include a resolution better than 200 c/s; sensitivity of 30 μV for full-scale linear deflection; response ±1dB from 1 kc/s to 25 Mc/s and a 50dB linear dynamic range. The VR-4 is the fourth module in the Panoramic range and is, as with the others, interchangeable without the need for extensive calibration and adjustments when adapting a c.r.o. for spectrum anlysis.

WW 318 for further details

NEWS FROM INDUSTRY

NEW RESEARCH FACILITIES AT S.T.L.

THE Postmaster General recently opened a 22,000 square foot extension to the Standard Telecommunication Laboratories at Harlow, Essex. The extension houses an I.B.M. 360/30F computer that will provide S.T.L. with a scientific computing service and will also be employed for direct circuit design. Research is at present being carried out at S.T.L. in a wide range of fields, including an investigation into the use of a dielectric fibre as a waveguide for the transmission of energy at optical frequencies. A fibre with a refractive index higher than its surrounding medium is used. This form of structure guides the electromagnetic waves along the definable boundary between the regions of different refractive index. With materials at present available signals suffer a very high attenuation, for instance best quality optical glass has a bulk loss of some 200 dB/kM, this being largely due to the presence of impurities. It is thought that by employing semiconductor refining techniques the impurity content could be reduced to such a level that a loss of less than 45 dB/kM would occur, rendering the technique an economical proposition.

Work is being carried out on resistive thin film devices with potential applications as an analogue or a digital memory element. Its operation depends on the fact that when an insulator is heavily doped with a metal, such as gold, it develops an appreciable conductivity which is voltage dependent. If a voltage of more than four and less than eight volts is applied to the device and removed in a period not exceeding 100 μs the a.c. conduction state corresponding to this voltage remains in an induced or memory state indefinitely. The state of the device can be read non-destructively by applying not more than three volts.

The major cost of a small digital computer is in the core store, because of this S.T.L. have developed a computer (primarily intended as a control unit for navigational aids) which does not employ a core store. Typical problems that can be solved and tasks that can be carried out by the computer are:—the digital generation of sweep signals for a p.p.i. display, the generation of vector lines on a radar display with numerical read out of range and bearing, and the generation of an apparent or "ghost" beacon for air navigation. This last device enables new flight paths to be defined, without the expense of setting up more beacons, by changing the apparent location of a beacon received through the VOR/DME system.

Marconi International Marine Co. Ltd. have, under the terms of an agreement with the Canadian Marconi Company, acquired exclusive distributing rights in all countries (with the exception of North America) for the Voyageur transistor marine radar equipment. The low-power consumption of about 160 watts and the small size of Voyageur make it suitable for installation in small craft. The 6-kW transmitter has a pulse length of 0.2 μs and a p.r.f. of 1.5 kc/s. The display unit incorporates a 10-inch diameter tube which can be switched to cover 1, 4, 8 or 16 nautical miles, markers can be positioned at 0.25, 1 or 2-mile intervals. Voyageur can be powered from either 13, 26 or 36 volts d.c. or, with the addition of a power unit, 115 or 240 V, 50/60 c/s a.c.

G.E.C. (Electronics) Ltd. have been awarded a contract by the G.P.O. to supply, install and test ancillary equipment to be linked with the second aerial now in course of construction at the Goonhilly Communications Satellite Station. The equipment will enable Goonhilly to handle traffic to and from several overseas stations simultaneously. Hitherto, Goonhilly has only only been able to work one destination at a time via "Early Bird."

It has been announced by the boards of the **English Electric Company Ltd.** and **Elliott-Automation Ltd.** that a complete merger between the two companies is planned. This is felt by both companies to be the best method of co-operation for the purpose of exploiting and accelerating the very large potential growth in the field of industrial control, automation, computers and electronics in general.

Standard Telephones and Cables Ltd. have started producing a range of high reliability semiconductors for use in submarine cable repeaters. Initial quantities will be used in the supervisory circuits of the South Africa-Portugal cable and will be the first time that transistors have been used in a long-distance deep-water submarine cable. Because of the loss of revenue and the expense of raising a submarine cable for repair when a fault exists all components must exhibit extreme reliability, for example the characteristics of these semiconductors must not deteriorate by more than 5% over a twenty-year period of continuous operation. The necessary development work has been carried out in collaboration with the G.P.O. and has been partly supported by the Ministry of Defence.

The Marine Division of **Redifon** have introduced two new electronic systems for use on board ship, these were both announced at the recent World Fishing Exhibition at Olympia. The first is a public address system that allows speech or music to be broadcast either over the whole network or selectively to individual loudspeakers or groups of loudspeakers. Two-way communication can be established between any particular loudspeaker and the control point by using the remote speaker as a microphone. Also shown was a ship's communal aerial amplifier (designated type MCU.5) which can supply up to 100 cabins with long-, medium- and short-wave signals for use on private radio receivers. It eliminates the necessity for unsightly makeshift aerials that can cause errors in the vessel's direction-finding equipment. Another advantage is that it reduces break-through from the ship's transmitters by introducing heavy attenuation in certain frequency bands.

The Instrument Division of **Livingston Laboratories** have supplied the first installation of a new equipment known as HYTRESS (High Test Recorder and Simulator System) to Vauxhall Motors. The equipment basically consists of two separate units, the first of which is built round a Nagra tape recorder. A test vehicle is fitted with a resistance-to-frequency transducer unit which senses the various engine parameters, oil pressure, r.p.m., load, etc., and records them on tape using f.m. techniques. The tape, so prepared, is played back on the second part of the equipment and the various signals operate the controls of a dynamometer which loads an engine to be tested. In this way the engine being tested is subjected to exactly the same conditions of load and engine speed as the engine in the test vehicle encountered. At the same time the performance of the engine under the simulated conditions is monitored and recorded.

The Marconi Company have received an order through the World Meteorological Organization in Geneva to supply a self-tuning transmitting station to be installed in Brazilia. The station will consist of two 7.5 kW self-tuning transmitters, double diversity single sideband receivers, voice frequency telegraph terminals and the necessary aerial equipment. The station will act as a communications hub for the World Weather Watch effort in the region and it will be employed to relay information from regional centres to the World Meteorological Centre in Washington.

The M-O Valve Company are setting up a coast-to-coast marketing organization in the U.S.A. for the sale of industrial tubes to equipment manufacturers. The company, which sells under the Genalex trade mark in the U.S.A., have appointed the Metropolitan Overseas Supply Corporation of New York to handle their products in this rapidly expanding market.

The **Decca Navigator Company** have received another substantial order from the U.S.A. for their Decca/Omnitrac navigation system. The order is from the Flying Tiger Line and will equip their new fleet of cargo carrying DC8-63F aircraft with a duplicated Decca system.

General Telephone and Electronics International Inc. who have their European head office in Geneva, Switzerland, have started constructing a large colour television picture tube manufacturing plant in Belgium. The 80,000 sq ft plant located at Tienen, about 30 miles east of Brussels, is scheduled to begin production of Sylvania Color Bright 85 tubes in 1968.

Leevers-Rich Equipment Ltd. have dispatched the first of a new high-speed tape duplicating system to Sweden for installation in the linguistics department of the University of Sweden. Duplication of tapes may be carried out at single, double or quadruple speed giving the equipment the capability of copying up to 24 tapes, each of 600 ft, in one hour.

For some time equipment designed and manufactured by **Audio Associates Ltd.** has been marketed under the brand names of other manufacturers. Following a move to new premises at Wareham, Dorset, the company has decided that as a matter of future sales policy the majority of products, which includes a range of audio equipment, will be sold under the brand name Audac.

The Solartron Electronic Group Ltd., Farnborough, Hants, have been awarded a contract for the supply of two radar simulators to the United States Maritime Administration. The simulators, which will be situated in San Francisco and New Orleans, will be used to train merchant marine officers and captains in navigation techniques.

Electrosil have announced price cuts of up to 60% on the range of American Signetics integrated circuits marketed by them. Price for a dual in line gate is now down to about £1 and binary elements cost in the region of £3 for small quantities.

Electroniques, the S.T.C. component and equipment suppliers of Edinburgh Way, Harlow, Essex, have been appointed exclusive U.K. agents for the **Hallicrafters** range of amateur and professional communications equipment.

Rank Bush Murphy have announced that they have received orders for £3M worth of colour television sets and that production is geared to exceed 1,000 sets per week by the Autumn.

Hirel Electronic Developments Ltd. have acquired the **John Compton Organ Company Ltd.** and have formed a new company, Compton Organs Ltd.

In this section last month it was erroneously stated that **Wayne Kerr** had received a contract worth £1M from the Ministry of Technology, this should have been £100,000.

WORLD OF AMATEUR RADIO

Wireless Telegraph Bill, 1967

ANY fears that the Wireless Telegraphy Bill now before Parliament would enable the Postmaster General to interfere with the legitimate activities of radio amateurs were dispelled during the third reading when he gave an assurance that the Radio Society of Great Britain would be consulted before any steps are taken to ban the manufacture of certain types of communication equipment.

Some weeks earlier, Mr. R. F. Stevens (G2BVN) Immediate Past President of the R.S.G.B., had attended a meeting of the Conservative Broadcasting Committee at the House of Commons to explain the Society's views.

During the third reading the Postmaster General explained that the purpose of the clause (which had led to fears being expressed by the R.S.G.B. and other bodies) was for consumer protection and was intended to keep the radio market clear of apparatus that interfered with licensed radio. He emphasized that amateurs would benefit from the inclusion of this clause as would other licensed users of radio apparatus and equipment.

Amateurs in Region II meet in Caracas

THE second triennial conference of the Union Interamericana de Radioficionados Union—Region II of the International Amateur Radio Union—was held in Caracas, Venezuela, in May. Ten countries were officially represented—Argentina, Bolivia, Canada, Colombia, Jamaica, Mexico, Panama, Peru, U.S.A. and Venezuela—and five others were represented by proxy—Bermuda, Chile, Ecuador, Guatemala and El Salvador. The five days of meetings and informal discussions ended in unanimity of agreement on mutual objectives and plans for the amateur radio service in the American hemisphere.

Reports on the status of amateur affairs in each country were too voluminous to digest at the time and are to be published separately. Plans for an expanded network for emergency communication were endorsed and each society was urged to set up an "intruder watch" to report on the continued presence of non-amateur stations in "exclusive" amateur bands. The Conference approved the concept of a special advisory committee of one person from each of the three I.T.U. world regions to work with the president of the I.A.R.U. in instances, such as at an international radio conference, where rapid decisions may be required on the I.A.R.U. attitude towards problematical proposals.

The Conference decided that more attention should be paid in the American hemisphere to the development of v.h.f. and u.h.f. operation. The next Conference is to be held in Jamaica during 1970.

Prophecy.—The Smithsonian Astrophysical Observatory, Cambridge, Massachusetts, predicted that the interception on June 6th at 1000 G.M.T. between the Earth and a recently discovered comet would result in a major high intensity, short duration, meteor shower, perhaps equal to, or better than, the Leonids showers of 1965 and 1966. Reports on meteor scatter propagation around June 6th and 7th should be sent to A.R.R.L. Headquarters, 225 Main Street, Newington, Connecticut 06111, U.S.A.

Amateur Radio in Rhodesia.—The latest edition of the Call Book of the Radio Society of Rhodesia (formerly the Radio Society of Southern Rhodesia) records details of the 178 call signs, names and addresses current as at May 6, 1967. The secretary of the Council is Mrs. Mollie Henderson, ZE1JE, P.O. Box 2377, Salisbury, the chairman is Vincent Hustler, ZE4JE and the treasurer, Mal Geddes, ZE3JO.

Knokke Convention.—An International Amateur Radio Convention, the third of a series, is to be held in Knokke, Belgium, during the period September 15th-17th. The Convention is being organized jointly by the Oosthoek Section of the Belgium National Society (U.B.A.) and a local committee. Vistors who travel by road and have mobile equipment installed in their cars will be able to compete for the Grand Prix du Casino de Knokke. Rules and special log sheets are available from the organizers. F. J. H. Charman, G6CJ, will demonstrate miniature aerials at a lecture due on the second day, and this will be followed by a car rally for the Martini trophy. The programme will include a wide range of other events, full details of which can be obtained from Lucien Vervarke (ON4LV), Lippenslaan 284, Knokke 1, Belgium.

R.S.G.B. Exhibition.—The 1967 International Radio Engineering and Communications Exhibition organized by the R.S.G.B. will be held in the New Hall of the Royal Horticultural Society, London, S.W.1, from September 27th to 30th. Dr. John Saxton, Director of the Radio and Space Research Station, Slough, Bucks, will open the Exhibition and during the evening of Friday, September 29th, the Council of the Society is to give a reception to visitors from overseas. Further information about the Exhibition may be obtained from R. F. Stevens, G2BVN, c/o R.S.G.B. Headquarters, 28 Little Russell Street, London, W.C.1.

Welcome to Yugoslavia.—The Yugoslav national amateur radio society, Savez Radio Amatera Jugoslavije (S.R.J.) is organizing a summer camp for radio amateurs at Bauske Vode, near Makarska, on the Adriatic coast from August 21st to September 5th. An amateur radio station will be in operation. The full pension rate for radio amateurs and their families has been fixed at the low rate of $2.50 each per day. Full details can be obtained from S.R.J. Headquarters, P.O. Box 48, Belgrade.

Reciprocal Licensing Information.—The Amateur Radio Mobile Society and the Radio Society of Great Britain have completed, jointly, a booklet of information setting-out details of reciprocal licensing in Austria, Belgium, Finland, Germany, Luxembourg, Morocco, Netherlands, Poland, Portugal, South Africa, United States of America and Yugoslavia. Copies of the booklet can be obtained by sending 9d in stamps to Norman Fitch (Hon. Sec., A.R.M.S.), 79 Murchison Road, London, E.10.

Amateurs and Tasmanian Bush Fires.—Tasmanian radio amateurs played a major role in providing communications during the disastrous bush fires which swept the island recently. The Australian press praised them for remaining at their stations and reporting danger areas and new fire outbreaks, relaying instructions to otherwise isolated fire fighters and handling welfare traffic. Earlier this year Australian radio amateurs provided emergency communications during widespread floods in Queensland.

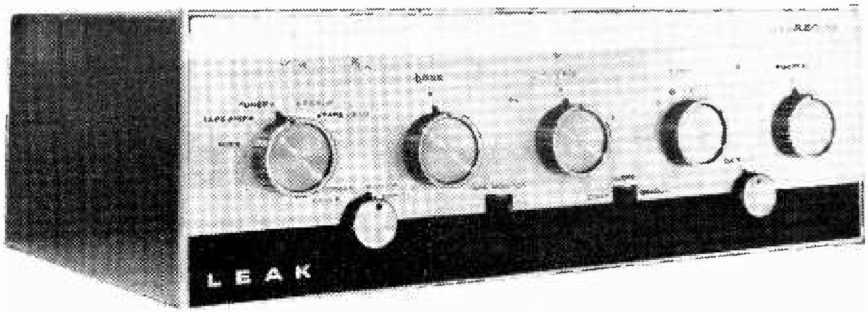
New Suffix.—The Post Office has decided that holders of a U.K. Maritime Mobile licence shall use the suffix /MA (signifying Maritime Alternative) when in port.

Faröes Award.—To commemorate the formation of a national Amateur Radio Society in the Faröes Islands a new certificate WAOY (Worked all OY) is being offered for contacts made with the Faröes after April 11th, 1965. Rules for the Award can be obtained from F.R.A., Box 184, Thorshaven.

Canadian Amateur Licences.—Five-year station licences have replaced annual licences previously issued to Canadian amateurs.

JOHN CLARRICOATS, G6CL

Britain's best Hi-Fi Equipment



STEREO 30 Transistorised integrated amplifier . . .

Superb styling, faultless performance and a breakthrough in price and reliability. Price £49.10.0

Designed to give optimum musical results from records, tape, radio and microphone, the Stereo 30 has received enthusiastic acceptance throughout the world

Reprinted from "WIRELESS WORLD" Editorial, May, 1963

" . . . Last autumn, during his presidential address to the British Sound Recording Association, H. J. Leak demonstrated a prototype high-quality transistor amplifier which gave results indistinguishable from those of his valve amplifiers . . . "

" People sometimes ask why there is any necessity to change to transistors. The elimination of the output transformer is, in our view, sufficient reason now that solutions of the problem of linearity in the response of the rest of the transistor circuit have been found. As additional bonuses we get smaller size, cooler running and the prospect of longer life."

F. M. TUNERS

MONO

Trough Line 3 **Price: £27.0.0 plus P.T. £5.3.11**

Styled to match the Stereo 30 amplifier, this unit has gained acceptance as Britain's best FM Tuner.

Designed to obtain the highest quality reception of FM transmissions, the "Trough Line 3" is the successor of the "Trough-Line II" tuner.

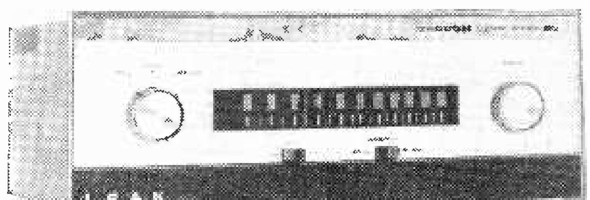
Engineering review of the Trough-Line II

" . . . Audio bandwidth is within 1 dB from 18 c/s to over 21 kc/s and distortion at 1.5V was impossible to measure on either Marconi or Heathkit Distortion-Factor meters . . . To sum up, the Leak Trough-Line II belongs to the very limited class of aristocrats in the tuner world. It has a positively astounding RF performance, impeccable limiting, ditto audio quality, and effective, but almost superfluous AFC . . . The price may appear substantial, but this is deceptive—you get a lot for your money. This tuner is definitely for the " carriage trade."

A. W. Wayne, " **HI-FI NEWS** "

STEREO

Trough Line Stereo **Price: £39.10.0 plus P.T. £7.12.1**



A stereo version of our " Trough Line 3 " tuner has been available for a number of years in the U.S. and Canada, where stereo broadcasting has been in existence since 1961. European Broadcasting Authorities are now using the same G.E.-Zenith System of stereo transmissions, and to enable listeners to obtain first class reception of these broadcasts we have developed the " Trough Line Stereo " Tuner. The dimensions and basic design are those of the " Trough Line 3 " with the addition of a transistorised decoder built into the chassis and with manual switching for mono or stereo operation.

LEAK THE FIRST NAME IN HIGH FIDELITY SINCE 1934.

Ask your Dealer for a demonstration, or post the coupon now for fully illustrated folders.

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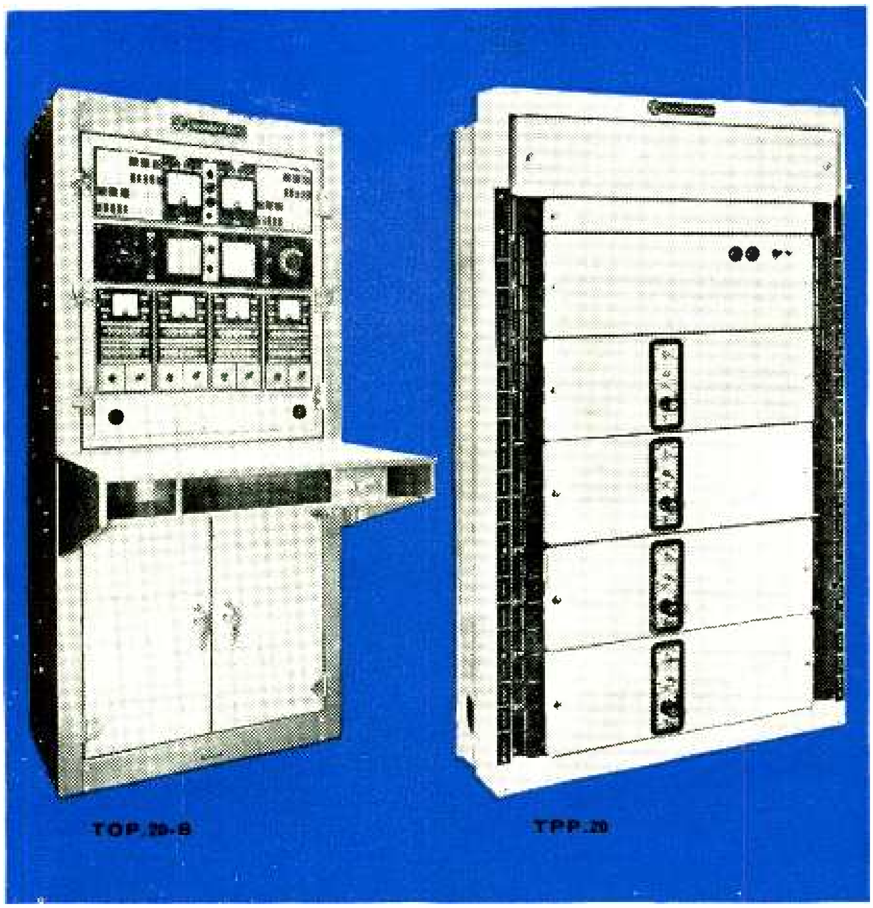
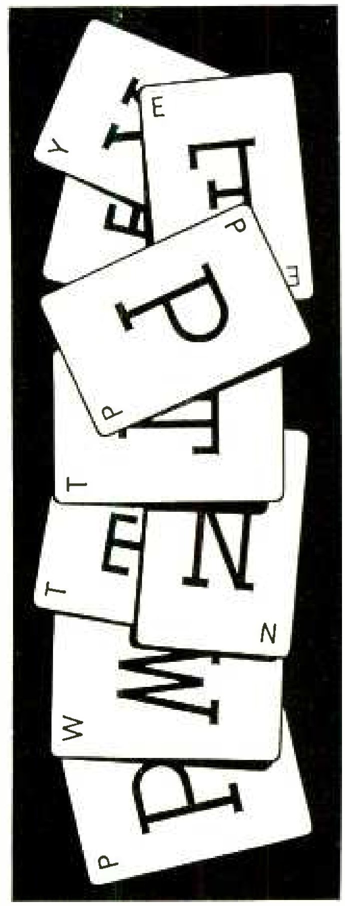
WW—090 FOR FURTHER DETAILS





Privacy in radio telephony

Whether it's a matter of discussing an important business deal or just saying hello to Mother, most international administrations operating into radio links consider privacy to be essential. STC provides this facility through the TPP.20 four-wire fully transistorized five-band speech scrambler equipment. This advanced equipment is designed to operate with h.f. radio link control terminals such as the STC TOP.20-B.

Both the TPP.20 and TOP.20-B handle four speech channels: both are extremely compact and embody modern techniques of transistorization and module construction

Five Band Speech Scrambler Type TPP.20
■ Four speech channels ■ Fully transistorized ■ 4-wire system (separate send and receive paths) ■ Remote selection of combinations ■ Compact yet extremely accessible ■ Cabinet 36 inches (91,4cm) high x $20\frac{1}{2}$ in (52cm) wide x $8\frac{3}{4}$ in (22cm) deep.

Radio Link Control Terminal Type TOP.20-B
■ Four speech channels ■ Fully automatic ■ Relay type VODAS switching ■ Built-in shifter and inverter for each circuit ■ Automatic station identification for each circuit ■ Console 75 in (190,5cm) high x 33 in (83,8cm) wide x 19 in (48,3cm) deep.

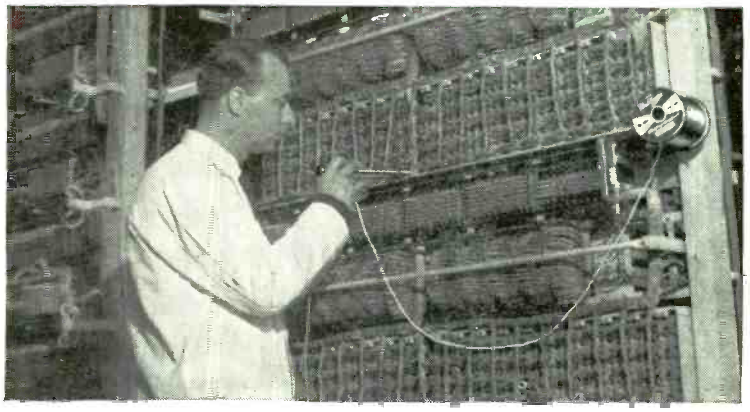
For further information please write, phone or telex Standard Telephones and Cables Limited, Radio Division, Oakleigh Road, New Southgate, London, N.11. Telephone: ENTerprise 1234. Telex: 261912.

world-wide telecommunications and electronics **STC**

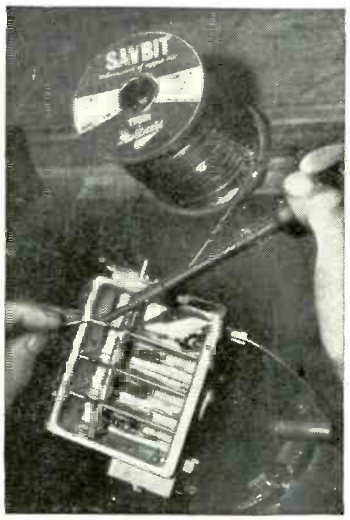
THE QUEEN'S AWARD TO INDUSTRY 1966

Why do leading electronic manufacturers use Ersin Multicore Solder?

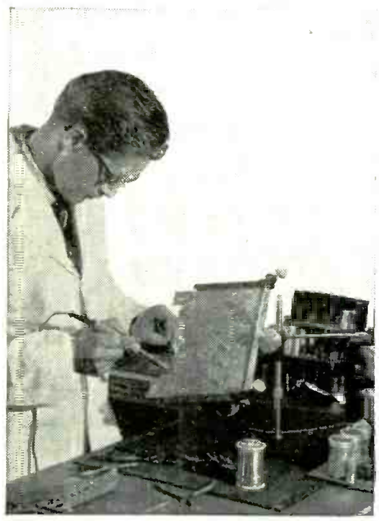
Because they realise that their reputation can rest upon the quality of the solder they use. For utmost reliability they use Ersin Multicore, the *only* solder containing the purest tin and lead, plus 5 cores of extra-active, non-corrosive Ersin flux. Whatever the application – the speedy soldering of miniature components or the individual production of large units – there is an Ersin Multicore Solder which is exactly right for the job.



Standard Telephones & Cables Ltd. The manufacture of Automatic Telephone Exchange Equipment involves many million soldered joints. The accessibility of many of the wires is restricted after they have been terminated, and it is therefore essential that only solder of proved reliability be used. Consequently Ersin Multicore Solder, 60/40 alloy, has for many years been preferred by Telephone Manufacturers in Britain and overseas. It is approved by the British G.P.O. and many foreign Telephone Authorities, and is shown here being used by Standard Telephones and Cables Limited in the manufacture of a G.P.O. Line Finder Rack.



British Radio Corporation Ltd. The critical characteristics of U.H F. tuners in television sets demand the highest quality of soldering. Ersin Multicore Savbit alloy containing five cores of extra fast Ersin 366 flux, is seen here being used by British Radio Corporation Ltd.



H. J. Leak & Co. Ltd. Ersin Multicore Savbit alloy is shown in use in the wiring construction of the Leak Integrated Stereo 30 Transistor Amplifier. In order to ensure the utmost reliability, the Leak High Fidelity equipment has been made with Ersin Multicore Solder for more than 20 years.

ERSIN Multicore 5 CORE SOLDER

reliability

Engineers and Technicians are invited to write on their Company's letter-heading for the new 7th Edition of Modern Solders which contains comprehensive information on the full range of Ersin Multicore Solders, Soldering Machines and Solderability Test Machine.

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