Wireless World

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100-watt quality amplifier

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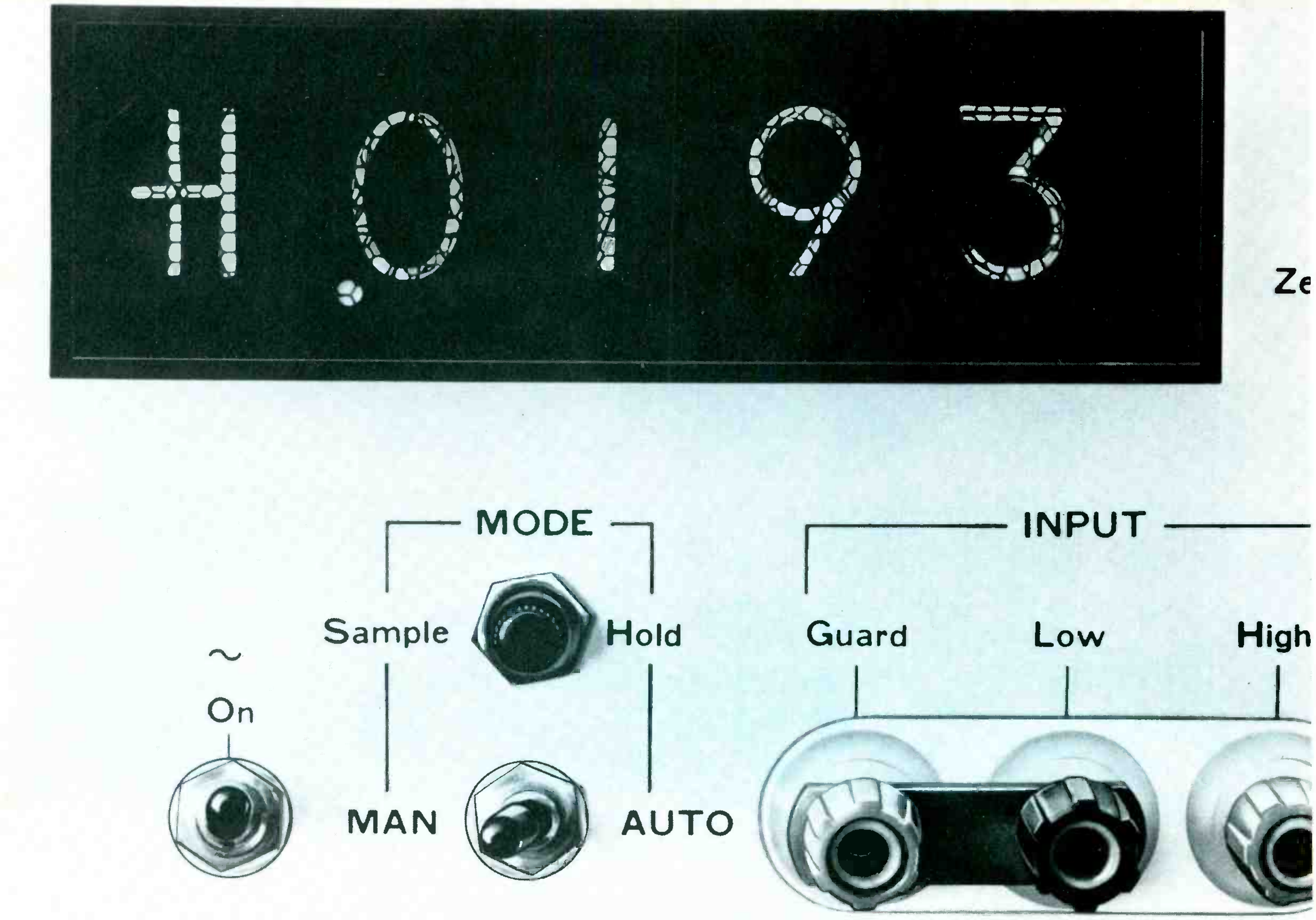

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Wireless World

Electronics, Television, Radio, Audio

Sixtieth year of publication

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Contents

This month's cover picture, showing the latest Burndept Electronics radiotelephone in use at an airport, draws attention to the continuing expansion in mobile radio communications, the technical problems of which are discussed in the report on page 494.

IN OUR NEXT ISSUE

Loudspeaker driver units. The fundamental parameters of moving-coil speakers will be examined and some new thinking introduced by Ted Jordan.

The first of a series of three articles on the design and construction of a **high-quality tape recorder.**

Details of an **extra-versatile tone control.**

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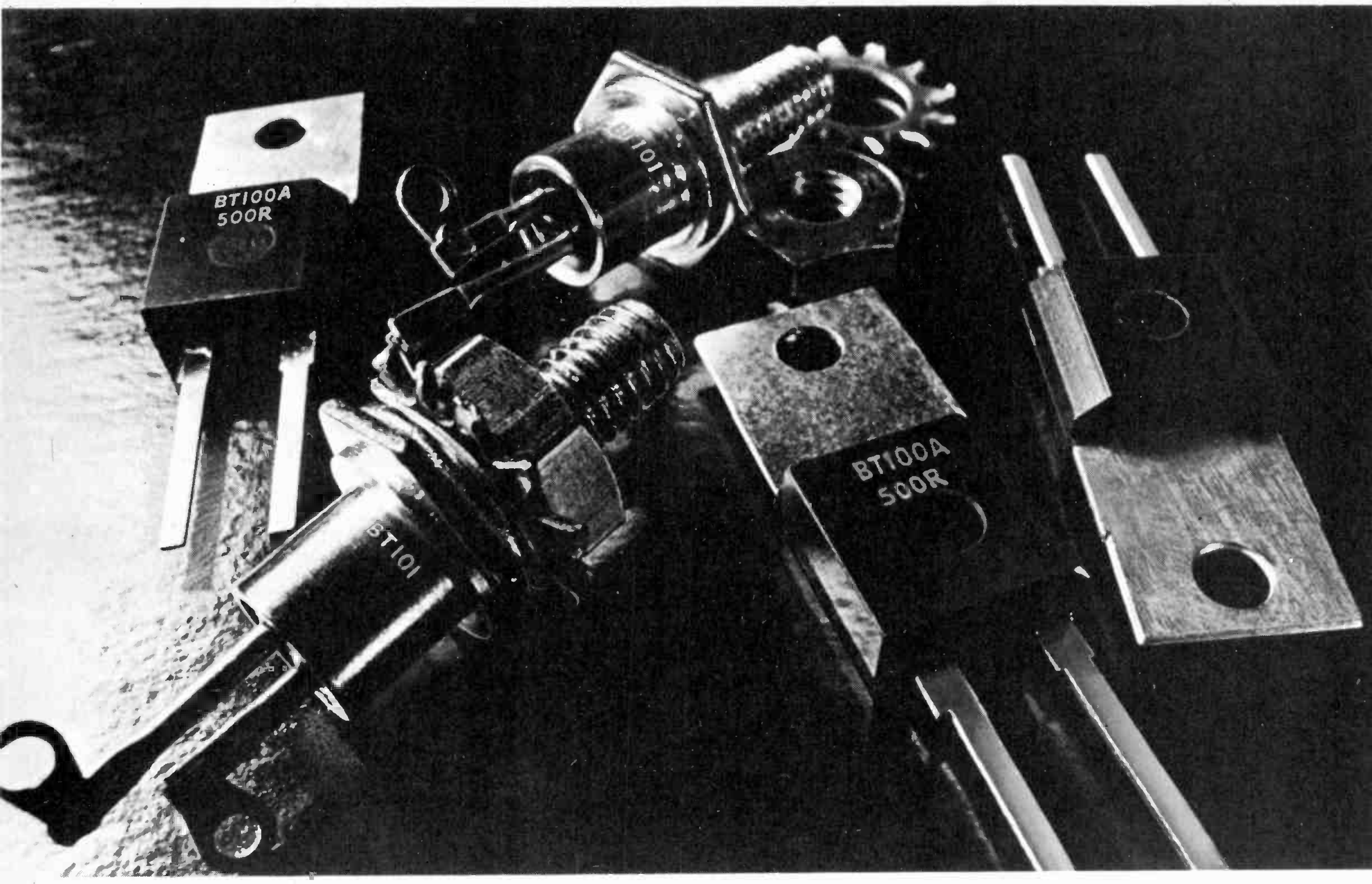

How we made thyristors a commercial proposition for consumer products

Three years ago a Mullard design team was given the problem of developing thyristors for motor speed control in washing machines and drills. Thyristors offered important advantages over conventional power control methods, but at that time, production was confined to relatively expensive industrial devices. The high unit cost was essentially due to specialist production techniques.

Two Requirements The Mullard team set about designing inexpensive thyristors, together with triggering devices, for use on domestic mains supplies. Two current handling capabilities were identified as being necessary to meet the range of applications—6.5A for washing machines and other heavy current loads, and 2A for drills and lighter loads.

Within six months two consumer type thyristors, BT101 and BT102, had been developed for 6.5A applications, and they were soon in mass production. Now these devices, in the TO-64 stud-mounted metal encapsulation, are well established.

Low-cost Plastic After further design work, a new *plastic* device, the BT100A, was introduced to meet the lower current requirements. Plastic power device technology is highly specialised, and only intensive effort over many years has resulted in the highly automated manufacturing techniques which ensure extremely good reliability.

Computer Testing To cope with the necessary high rate of production, computer techniques were introduced to record test results and to allow automatic grading. The testing cycle was significantly shortened by the use of high-current pulses for directly heating the thyristor crystal. This is one of the best automated methods of testing breakdown voltages at the highest junction temperatures.

The result? A range of thyristors capable of meeting all the consumer-appliance manufacturers' current needs, and of improving both the efficiency of power-control and the usefulness of the units controlled. They offer consumer product manufacturers smooth, continuous and efficient power control.

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Wireless World

The Dehumanization of Broadcasting

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Karl Marx, were he alive today, might have written that famous aphorism of his as: "Television . . . is the opium of the people". The fact that television is an opiate is generally blamed on the programme producers: the broadcasting engineers are inclined to take a superior, professional attitude and disclaim any responsibility for the material that is passed through their channels. Indeed some engineers, particularly in America, treat the programmes with open contempt. (The slang word "canned" used for recorded programmes is an implicit grouping of the creations of art and intellect with beer, soup, baby-food and other mass-produced commodities of a sloppy nature.) But it is too easy for the engineers to shelter in their sectarianism. They must become more aware of the effects of their work, their aims and attitudes, on the quality of what is broadcast. These effects may be indirect but are nonetheless real.

Anyone who has worked in mass-production industries will know that there is a strong pressure on designers to fashion their goods not so much for the convenience of the human customers as for the convenience of the production machinery. This, of course, increases manufacturing efficiency. If the resulting goods are not exactly what the customer likes he can always be persuaded to think that they are, by clever advertising. Equally it is well known that people must be continuously pressurized into buying goods which they may not really need, in order to keep the manufacturing plant fully loaded and hence economic—and, incidentally, to avoid unemployment and loss of consumers. The engineer, of course, has a vested interest in the design, manufacture and operation of the production machinery.

A similar situation is now developing in broadcasting. The mental and spiritual "goods" are becoming subservient to the broadcasting machinery, which is designed and operated by engineers whose main purpose is to achieve the most efficient distribution of canned, pre-digested, brightly packaged, expertly timed, second-hand experience to a mass audience.

At all costs the broadcasting machinery must be made efficient—and this seemed to be the cry at the recent International Broadcasting Convention in London (see p.484), where one of the main themes was the application of automation to broadcasting. One American contributor proudly remarked that the on-line computer programme for his computer-controlled broadcasting system had been "developed and documented to the point where it can be used as a general purpose program for any industry such as planning the production of manufactured goods."

Automation is fine, in so far as it relieves people of monotonous work, reduces operating costs and improves the technical quality of the broadcast programmes. But experience from industry shows that it is also a means by which the human being is even more completely enmeshed and demoralized by the production machinery. He is not driving but being driven, as Charlie Chaplin showed in "Modern Times", and now his brain is involved as well as his reflexes.

Nevertheless, one speaker at the I.B.C., on being asked why he had expressed disquiet about too much automation in broadcasting, said "because it can be destructive of human initiative and art". This speaker was an engineer. So there is at least one man who understands that the true work of engineers is not simply the perpetuation of engineering.

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Elements of Linear Microcircuits

1. What a linear microcircuit is, how it is made and packaged

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

What are these microcircuits that have revolutionized circuit design? In widest terms, they are a sort of supercomponent consisting of a number of circuit elements inseparably associated in a small package. In the ultimate they reduce the equipment designer's job to just fitting together a few prefabricated circuit blocks instead of designing a large complex of separate discrete components. No longer need the electronics experimenter puzzle out the design of a two-transistor 'Ridler' d.c. coupled pair for his tape replay amplifier; he just buys a ready built microcircuit.

Linear and digital: the difference

There are two classes of microcircuits (often called integrated circuits or just i.cs): digital and linear. Digitals are designed for on/off switching applications and provide the equipment designer with a range of complete logic elements, such as AND, OR, NAND, NOR gates, flip-flops and some extremely complex computer-type sub-systems. Linear i.cs are for applications where the output is in some way proportional to the input, and they provide the designer with ready-made d.c., a.f., r.f. and wideband amplifiers.

Digital microcircuits became generally commercially available in the mid-1960s and have been exhaustively discussed in the technical press since then. Linear i.cs did not become readily available until much later, and are only now finding wide use. This series of articles is aimed at the newcomer to the linear field.

Linear microcircuits can be 'multiple-purpose' or 'single-purpose'. Multiple-purpose units are gain-blocks which can be externally pin-programmed to perform a large variety of different circuit functions (usually by fitting different feedback networks). The archetype of these is the operational amplifier † or op. amp. This is a very high gain d.c.-coupled amplifier with a response which is completely defined by feedback. It was the earliest linear microcircuit to become generally available and is the best known.

Most circuit designers prefer single-purpose microcircuits, which are complete in themselves and do not need additional circuits designed around before they can be used (as is the case with multiple-purpose linear i.cs). Fortunately, more and more single-purpose linear i.cs are coming on the market, ranging from a simple package of a matched pair of transistors up to a complete 100-W audio power amplifier.

Early developments

Before we look at current methods of microcircuit manufacture, it is of interest to look back over the past three decades at the landmarks in their evolution. Up until World War II, the normal methods of assembling electronic equipment was to mount all the heavier components on some form of chassis, and then interconnect them with point-to-point wiring, either directly or via tag boards.

The first major move towards present-day microcircuits began with the miniature proximity fuses developed for the nose-caps of artillery shells in World War II. These radio-controlled fuses were closely packed assemblies using special valves, but the technique never spread into large scale commercial use because of the bulky valve needed for amplification.

The development of microelectronics really started with the invention of the transistor in 1948. This got rid of the large wasted vacuum space inside the valve, its inefficient heater and the need for a high anode voltage. Assemblies could now be much smaller, but they were still only scaled-down versions of the old point-to-point inter-wiring of discrete components.

In the late 1940s, Sargrove in England started a move away from point-to-point wiring. He pioneered a development in which a radio receiver was built on a ceramic substrate‡ on which resistors and intercomponent wirings were printed and fired rather than separately mounted. This was one of the earlier experiments in the integration of circuitry, but in spite of it, point-to-point wiring continued unopposed.

Later printed circuit boards (more correctly printed wiring boards) became a commercial reality, and this form of integrated wiring provided another big step towards commercial microelectronics.

The 1950s also saw many different approaches to miniaturization, apart from assembling conventional miniature components on a small printed circuit board. They gave rise to names like 'Cordwood', 'Tinkertoy', 'Micromodule' and '2D', which are now largely of historical interest only. Details can be found in books such as 'Microelectronics' by E. Keonjian. (McGraw Hill, 1963.)

All these forms of microminiature assemblies developed in the 1950s were more expensive than the standard printed circuit board and they were only used in equipment where cost was not the governing element, military equipment for instance.

Today's microcircuits

In 1958 a development occurred which changed the whole face of things. Kilby of Texas Instruments came up with an interconnected assembly of resistors and transistors made by diffusion in tiny silicon chips. The true monolithic silicon integrated circuit (s.i.c.) had been born. The first of these was a mesa-type r.t.l. (resistor transistor logic) bistable and it used only two chips interconnected by bonding wires in a single package.

In 1960 the celebrated Fairchild planar process for manufacturing transistors was developed which gave a strong impetus to

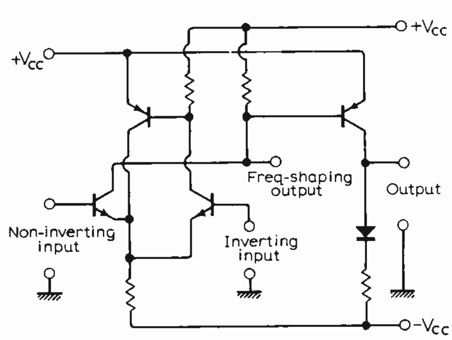

Fig. 1. Circuit of 1962 linear silicon integrated circuit (Texas Instruments SN521 operational amplifier).

*Newmarket Transistors Ltd.

†Operational amplifiers were dealt with at some length by G B. Clayton in a series of articles which appeared in the February to December issues of *Wireless World.*

‡Some details of this development are given in an article "Automatic Receiver Production" which appeared in the April 1947 issue of *Wireless World.*

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the production of monolithic s.i.cs. Early units were the easiest-to-make digital types such as the 1961 Texas Instruments Series-51 r.t.l. But linear s.i.cs were not long in arriving. By 1962, the Texas Series-52 linear i.cs were on the market. Typical of these was the SN521 general-purpose single-chip 62dB differential amplifier the circuit of which is given in Fig. 1.

In the linear field, however, the general use of linear s.i.cs can be said to have started with the now well known Fairchild μA709 op. amp. which came on the market in quantity in 1965. Since then there has been a proliferation of multiple-purpose linear amplifiers, particularly of the operational amplifier type.

But the linear s.i.cs available by the mid-1960s did not get the immediate wide usage that their high technical specifications invited. This was partly because production was small and the cost was much higher than a designer could achieve by using conventional component circuitry. Also, run-of-the-mill circuit designers were not skilled at using wideband, high-gain operational amplifier blocks for general purpose circuitry. They would rather have had low-cost single-purpose units.

In this climate, significant developments began along different lines. Techniques developed for producing prefabricated assemblies of resistors, capacitors and interconnections by printing on ceramics (thick film) or vacuum evaporation on glass (thin film) were married to special miniature semiconductor devices suitable for attaching to such substrates. Out of this marriage came the hybrid active linear microcircuit, which had advantages over monolithics in some areas. The two main ones were that the hybrid could be fabricated economically in small batches and that single-purpose units could be made up readily.

By the end of the 1960s, many semiconductor manufacturers had gone into monolithic s.i.cs. Cheap standard multiple-purpose linear i.cs had become widely available, but there were not many standard single-purpose units around. You could get a special s.i.c. custom built, but you would have to use very large quantities for it to be economic. As a result, many custom hybrid houses had sprung up, using thick and thin film techniques, to serve the smaller-run equipment manufacturer who could not use the existing standard monoliths and was too small to have a special monolith built for him. Almost as a by-product, these hybrid houses also put on the market standard commercial single-purpose linears.

It is anybody's guess how the demand for linear microcircuits will divide itself up in the future between standard single-purpose, standard multiple-purpose, and custom-built units. One estimate is that ih the 1970s linear applications will be met 50% by off-the-shelf single-purpose standards, 25% by multiple-purpose standards, and 25% by custom specials. As to how far the units will be monolithic and how far hybrid, again there is much doubt. The chances are that most multiple-purpose standards will be monoliths, most custom units hybrid, and single-purpose units a mixture of monolithic and hybrid.

No reference has been made so far to m.o.s.t. (metal oxide semiconductor technique§) microcircuits which use f.e.ts (field effect transistors) instead of bipolar transistors as the basic circuit elements diffused into silicon chips. They are cheaper to produce than bipolar monoliths, and have already found wide use in low-cost digital applications. However, they are not as yet well suited directly to linear applications. and will not be discussed further here.

As to hybrid technologies, thin film is gradually being phased out for cost and technical reasons, and most hybrids are now thick film.

Monolithic silicon circuit manufacture

Manufacturing monolithic s.i.cs is a highly complex business and many books have been produced on the subject. If you are seeking detailed information, you should consult one of the standard texts, such as Motorola's 'Integrated Circuits—Design Principle and Fabrication', edited by M. Warner (McGraw Hill, 1965). In this article we will give only a sketchy outline of how s.i.cs are made.

The process starts with an ingot, usually about 250mm (10in) long and 25mm (1in) diameter, of highly refined single-crystal silicon, shown in Fig.2(a). The ingot is sawn up into thin slices of which one is shown at Fig.2(b), and the s.i.cs are made in these slices.

As shown in the enlarged view of a single slice in Fig.2(c), a large number of identical circuits are formed in a regular pattern. Various techniques are used, such as high-temperature diffusion of impurity gases into the slice, selective surface etching of photoresist masking, formation of protective 'glass' (silicon oxide) surface layers, and deposition of metallic interconnections and lead bonding pads on the surface by vacuum evaporation (thin film) techniques. Depending on the area of the individual circuit in the pattern, a single slice typically produces anything from 200 and 2000 identical integrated circuits at the one time.

The slice is next scribed along the dividing lines between the circuits and broken up into individual units. A single circuit then finally appears as at Fig.2(d)—enlarged—in the shape of a square chip between 0.5mm (0.020in) and 1.25mm (0.050in) across with visible metallization on the surface.

This chip is packaged by bonding it face-up on a support such as the multi-lead TO-5 header shown in Fig.2(e), with the metallized bonding pads visible on the face. Connections are then made from the header leads to the pads by gold or aluminium wire about 0.025mm (0.001in) diameter. After being tested, the package is sealed. In the example shown, a metal top cap is fitted by welding round the rim.

§We normally regard m.o.s.t. as meaning metal oxide silicon transistor because this describes the structure of the device. That is a metal gate (aluminium) insulated from the drain/source silicon by a layer of silicon oxide. Ed.

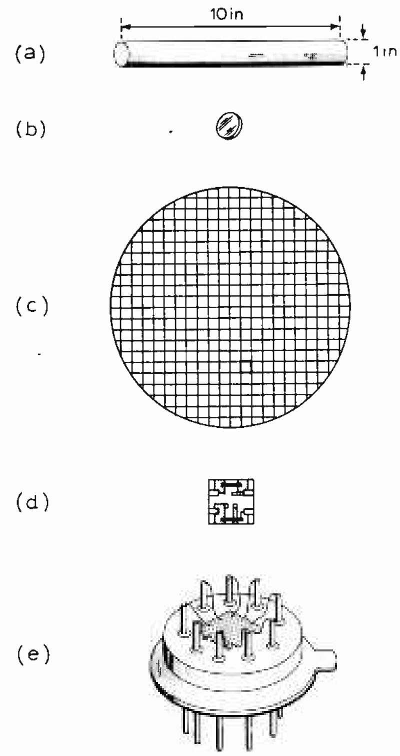

Fig. 2. Construction of silicon integrated circuits: (a) basic silicon crystal ingot from which process starts; (b) thin slice (wafer) cut from ingot; (c) enlarged view of slice after processing produces large number of identical circuits inside the silicon formed by repeated diffusion, oxidation, and selective etching with final evaporation of surface metallization for interconnections and bonding pads; (d) one of many single-circuit chips obtained by scribing and cracking the complete slice, and (e) individual circuit chip mounted on header with connections to header leads.

From this necessarily brief summary, it should be evident that the basic element in a monolithic s.i.c. is a very small processed thin chip of silicon about the area of a grain of sugar. This makes it clear why high-power dissipation presents a major problem in s.i.cs, because of the difficulty in getting the heat away from the tiny chip. Normally, temperatures inside the chip must be kept below about 150 to 180°C and because of its small size it is hard to dissipate much power without exceeding this limit. This also explains why most of the commonly available s.i.cs have a power rating somewhere round 100 to 500mW (very much the same as a single transistor), and also why most of the high power linear microcircuits on the market tend to use the hybrid fabrication to be described below.

Thick film hybrid fabrication

The assembly of a thick film hybrid starts with a smooth ceramic (aluminium oxide) blank substrate, typically about 25mm

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(1in) square and 0.375 to 0.875mm (0.015 to 0.035in) thick, as shown in Fig.3(a).

On to this ceramic (which is an insulator) a matrix of passive circuit elements is screen printed and fired, just like the decorations on a piece of pottery, as shown in Fig.3(b). This produces an identical pattern of resistors, capacitors, insulating layers and metal interconnection runs and bonding pads (the last for attaching discrete components and external leads) in each cell of the matrix.

The large substrate is then scribed along the cell dividing lines and cracked up into individual small circuit substrates. One of these is shown enlarged in Fig.3(c). The next step is to attach any subminiature discrete components required, such as the transistor shown in Fig.3(d), and the final substrate preparation is the attaching of external leads shown in Fig.3(e).

After being tested, the hybrid circuit is encapsulated in some form of protective package, as shown in Fig. 3(f). It can be seen that the dissipating semiconductors can be dispersed over a relatively wider area than is possible in an s.i.c. chip so

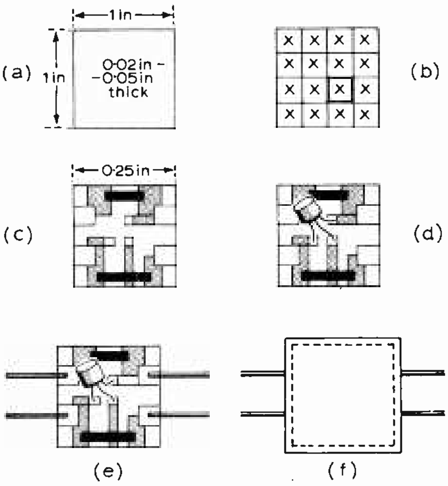

Fig. 3. Construction of thick-film hybrid microcircuits: (a) Starting ceramic substrate; (b) number of identical R, C and conductor networks printed and fired on substrate; (c) single circuit substrate scribed and cracked from complete multiple-unit substrate; (d) discrete components such as transistors attached to substrate. (e) Leadout wires attached; (f) circuit encapsulated in protective package.

higher power dissipation is possible. On the other hand, it must also be clear that the overall package size will tend to be larger for hybrids.

The packaging of a microcircuit is of importance not only because it is all that the user sees of the device, but also because it has such an important bearing on cost and reliability. For this reason, the rest of this article will be devoted to packaging aspects.

Microcircuit packaging; the problems

Just as with transistors, microcircuit packages are of two basic types, 'hermetic' (metal or ceramic and glass) and 'non-hermetic' (plastic). Non-hermetic are much cheaper than hermetic, but have not yet reached the stage where they can be regarded as satisfactory in extremes of temperature and humidity. Thus in high reliability applications, hermetic packages are the rule. Initially only hermetic packages were accepted for professional use, but recently plastics have improved so much that they are creeping in for the less demanding applications.

In commercial linear microcircuits, you will therefore find three grades in the market: (a) *Entertainment* or *Consumer*, suitable for use from 0 to 70°C and in low humidity environments (and almost always non-hermetic), (b) *Industrial*, suitable for use from −20 to +100°C and in medium high humidity (mostly hermetic), and (c) *Military* for −55 to +125°C and high humidity environments (until now always hermetic).

Unfortunately for the user, package standardization for microcircuits is a long way off. We are not yet in the comforting climate of transistors where you can take the same JEDEC standard TO-5 outline device from several different manufacturers and find that the case sizes varied by only a few thousandths of an inch and that the standard emitter-base-collector numbering of leads round the can obtained in every case.

For linear microcircuits at the time of writing there are over 700 different shapes, sizes and lead configurations available. In this chaos of packages offered, however, some trends are beginning to make themselves clear.

Monolithic s.i.c. packages show more standardization than hybrids because they have been around longer. But packages which have reached some acceptance for monoliths have had to be severely modified to encompass the generally larger hybrid element.

With regard to outlines, packages fall into four main classes: (1) low-power packages with leads to be inserted through circuit boards and soldered on the copper side; (2) low-power packages designed to be mounted directly on the copper side with leads attached flat to the metallization by soldering or welding; (3) medium-power packages with integral heat sinks for printed circuit board mounting; and (4) high-power packages designed for attachment to substantial metal chassis or heat sinks.

Low-power through-board-mounting packages

The most common low-power through-board-mounting microcircuit package is the *dual-in-line*, abbreviated to d.i.l. for the hermetic version and d.i.p. for the non-hermetic or plastic version b. Fig.4(a) shows the main dimensions of the most

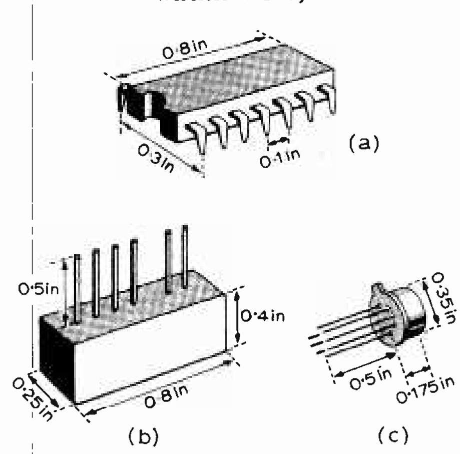

Fig. 4. Typical common low-power linear microcircuit packages for attachment to non-copperside of printed circuit board: (a) Dual-in-line; (b) single-in-line; (c) multilead TO-5.

common package, the 14-lead dual-in-line. Variants of the package may have anything from 4 to 24 leads or more. The interlead spacing of 2.54mm (0.1in) in the line of leads is standard (to allow conductor runs between the lead lands on the board). The inter-row spacing of 7.62mm (0.3in) is standard for monolithic s.i.cs, but, for the generally larger hybrid, other spacings such as 15mm (0.6in) are common.

Dual-in-lines are not easy to unsolder from circuit boards for servicing, and there is growing up another package style for through-board mounting which is easier to unsolder. This is the *single-in-line*, of which an example will be found in Fig.4(b). This s.i.l. can be thought of as half of a dual-in-line with the leads straightened into the plane of the device. It too, like the d.i.l., tends to use lead spacings of 2.54mm (0.1in), and is more common in hybrids than monoliths.

Historically the earliest monoliths were developed by semiconductor manufacturers and it was natural that they should package them in modified transistor cases. Fig.4(c) gives an illustration of the multilead TO-5 which may have anything from six to twelve leads. Because of the close lead spacing and the difficulty of removal from a printed circuit board, it is now not very popular with designers.

Low-power copper-side mounting packages

In these days of double-sided printed circuit boards and the demand for space

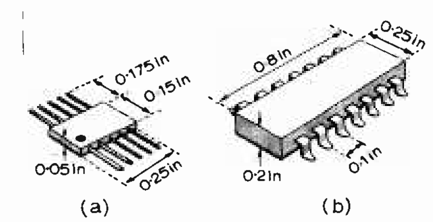

Fig. 5. Typical low-power linear microcircuit packages for attachment to copper side of printed circuit board: (a) Flat pack; (b) ready-to-reflow dual-in-line with preformed leads.

b Although d.i.p. always means a plastic package, we often find that in the various manufacturer's data sheets we receive, d.i.l. is also used to describe plastic packages. If your application needs a hermetically sealed microcircuit always check the data sheet for the type of package material used. Ed.

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saving, several packages have been developed for attaching to the copper side of the board. Two are fairly standard.

The *flat pack* shown in Fig.5(a) was developed by Texas Instruments for their early s.i.cs. With the general adoption of the dual-in-line package described earlier, users who wanted to mount them on the copper side of the board dressed their leads out flat as shown in Fig.5(b). This gave rise to the *readv-to-reflow* dual-in-line modification which manufacturers are now prepared to supply.

Medium-power microcircuit packages

The packages so far discussed usually cannot dissipate more than a few hundred milliwatts. Other packages had to be developed for higher powers, particularly in the linear field. These tend to fall into two main groups: (a) items designed for powers up to about 5W without any special substantial external heat sinking, and (b) high-power packages capable of dissipating up to 50 or 100W.

In the first, medium-power, category, several packages will be met with. Semiconductor manufacturers, accustomed to standard two-pin TO-3 outline power transistors, developed a multi-pin version of this outline, of which Fig.6(a) shows a typical example. On a printed circuit board this can dissipate up to about 2W (and on a substantial heatsink 10W).

Another transistor case used for medium-power microcircuits is the multilead TO-8 transistor package, of which Fig.6(b) is an example. This has twelve pins arranged in a square, but a sixteen-pin version is also available. This package can dissipate up to about 1W in free air and about 2.5W clipped to a substantial heat sink.

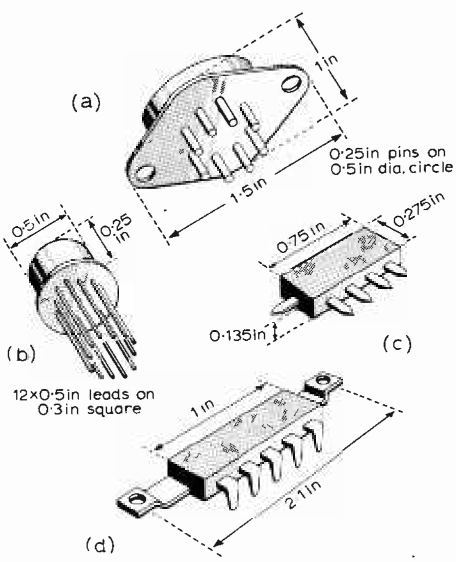

Fig. 6. Examples of medium-power linear microcircuit packages; (a) Modified multilead TO-3 power transistor package (2-5W); (b) modified multilead TO-8 intermediate power transistor package (1-2W); (c) modified dual-in-line with integral tongued heat sink (1W); (d) modified dual-in-line with integral heat sink for bolting to chassis (3W).

A different approach to a medium-power package is the integral *heat sink.* Fairly typical of this is the package sketched in Fig.6(c). This is really a dual-in-line with the leads dressed for reflow soldering and with a strip of metal inside extending from one end for better removal of heat from the chip Permissible power dissipation can be increased by soldering the metal tongue to board metallization or some area of metal. Packages like these are typically capable of dissipations up to 1W.

The power dissipation capability of integral heat sink package can be extended by making provisions for bolting to a metal heat sink. One well known example of this is the *chassis-mounting integral heat sinc* given at Fig.6(D). This package is used for a linear monolithic amplifier with a power output capability of 3W audio, which has been widely marketed in the United Kingdom. #

High-power microcircuit packages

Package design becomes a critical problem when we come to linear i.c.s. capable of handling more than a few watts, whether they be monolithic or hybrid. At the time of writing no standard packages have been evolved, but the main features to be expected in such packages can be seen in the illustrative example of Fig.7. This is a 50-W high-quality audio amplifier. In the casing outline at Fig.7(a), you can see that it is a fairly substantial package, 100 × 50 × 25mm (4 × 2 × 1in), with flanges for bolting to a chassis or heat sink. The terminals are stout pins issuing from one side of the package, to which connections can be made by soldering or by crimped-tag flying leads. The amplifier is of the quasi-complementary class-B type and the circuit used is shown in Fig.7(b). The main power-dissipating elements are the two output transistors. Some expertise is required to mount these in the package to ensure the most efficient removal of the heat—quite a problem when you realise that the power transistor chips are asked to dissipate internally 30W apiece.

Sockets for microcircuits

In the early days of transistors, when designers were a little uncertain how to handle them, it was common to fit sockets for them on printed circuit boards. We see the same development with microcircuits, and there is a lot to be said for it.

Sockets for standard dual-in-line, single-in-line and multilead TO-5 packages are nowadays fairly readily available from electronics distributors. Apart from distributors, some firms specialize in the supply of microcircuit sockets, such as Jermyn Industries in the U.K. and Augat or Barnes in the U.S.A. (with agents in the U.K.).

If you are buying microcircuit sockets, remember that they come in two types: "test" sockets specially designed for

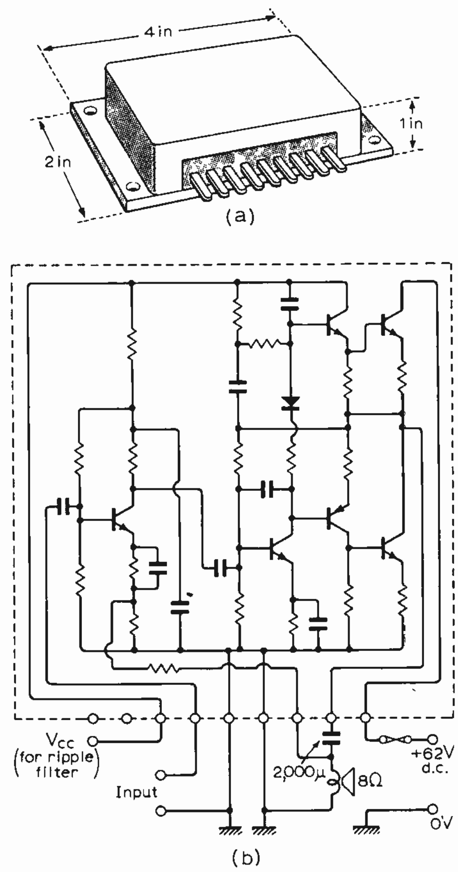

Fig. 7. Typical commercial example of high-power linear microcircuit (Sanken Electric SI-1050A, 50Watt Hi-fi power amplifier). (a) Package; (b) hybrid circuit.

continuous repetitive use, and "production" sockets for once (or occasionally twice) use in equipment. Obviously production sockets will be cheaper than the test type.

Availability of Linear Microcircuits

From this preliminary look at the linear microcircuit field, it should be evident that nowadays the designer can look to a large commercial armoury of such circuits around which to build his equipment. As later articles in this series will show, he will find linear i.c.s. for applications in frequencies from d.c. to 1,000MHz, powers from 0.1mW to 100W and gains from 0dB(×1) to 120dB (×1,000,000). The article next month will deal with what is commercially available in the way of linear i.c.s. and how to set about finding them.

Editor's Note. The arrival of the first integrated circuit from Texas Instruments in 1958, referred to by the author on p. 472, was foreshadowed by a report in *Wireless World* in 1957 (November issue, p. 516) dealing with early British work by the Royal Radar Establishment and Plessey. The title of our article, "Solid Circuits", later involved our reporter in some legal business with Texas, as this company subsequently used these words as a trade name for their devices.

#Plessey type SL403A.

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Circuit Ideas

Loudspeaker transmit/receive switch

In low-power radio transceivers it is customary to use the receiver output stage as the modulator when transmitting, a switch being used to disconnect the loudspeaker. The following circuitry allows this switching to be accomplished remotely without the use of a mechanical relay, and has been used in a four-metre portable transceiver in which all send/receive switching is accomplished electronically. The basic circuit is shown in Fig. 1 and has a very much lower insertion loss than any diode or other system that was investigated. The power loss is only slight with a 3Ω loudspeaker and is unnoticeable with a 15Ω load. Almost any transistors can be used but silicon planar devices give the greatest attenuation when the gate is off. Because of the fairly low reverse emitter-base breakdown voltage ratings of such transistors (about 5 to 6 V), some breakthrough occurs if the input voltage swing is greater than 12 V pk-pk. When one side of the input or output is referred to one pole of the supply battery the circuit of Fig. 2 can be used. If only one switched h.t. supply is available, for example that for the transmitter, the modification shown in Fig. 3 can be used. The diode ensures that Tr_4 does not conduct when any audio peaks cause reverse breakdown current to flow into the base of Tr_2.

D. A. TONG,
Kirkintilloch,
Dunbartonshire.

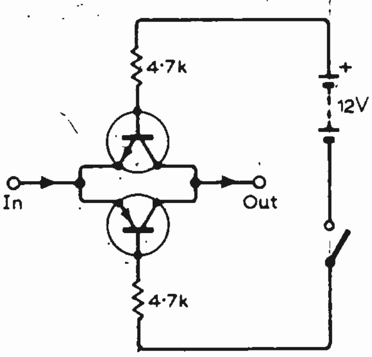
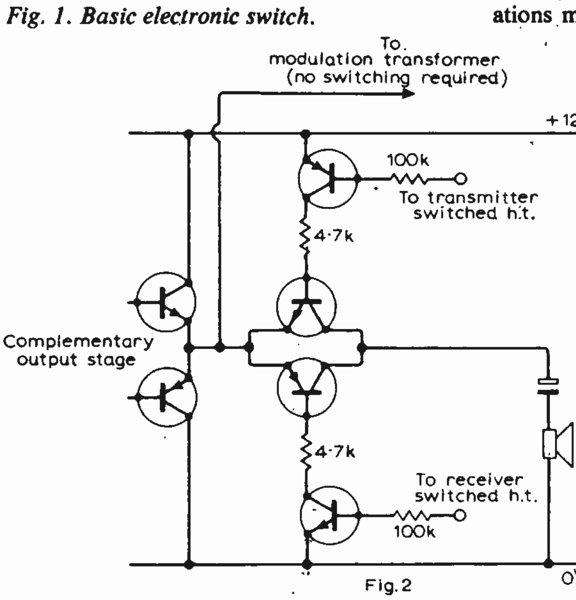


Fig. 1. Basic electronic switch.



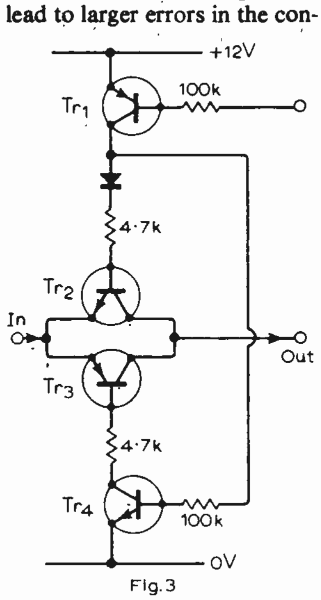
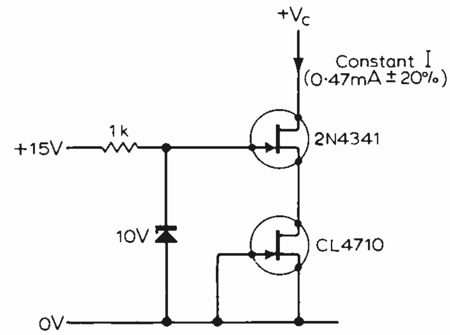


Fig. 2. Circuit for connection to two switched h.t. supplies. Fig. 3. Circuit arrangement for use with only one switched supply.

High-stability constant-current source

A convenient two-terminal constant-current source is a field-effect diode. Devices may be selected to have a temperature coefficient of less than 0.0005%/°C. However, the voltage coefficient of such diodes is typically 0.05%/V. This means that voltage variations may lead to larger errors in the constant current than temperature changes. A way of isolating the diode from voltage changes using an f.e.t. with its gate held at a constant voltage is shown. The f.e.t should be chosen to have a low V_p consistent with an I_{DSS} which exceeds the current passed by the current limiting diode. A 2N4 340 or 2N4 341 is suitable for use with the CL4 710. The gate source voltage of a 2N4 341 changed less than 10% over a 30V change in V_c giving a current stability of 0.0005%/V and 0.0005%/°C.

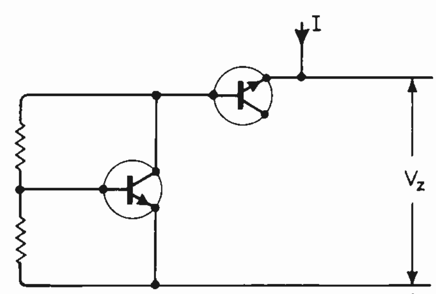


F.e.t. current source.

J. A. ROBERTS and J. R. JONES,
Swansea.

Cheap voltage reference

The base-emitter diode of most silicon transistors can be reverse-biased and made to operate as a zener diode. At currents of about 1mA a positive coefficient of voltage with temperature is usually found which is greater in magnitude than the negative temperature coefficient of the same diode when forward-biased. By operating a transistor in an 'amplified diode' arrangement the temperature coefficients of this diode and the zener can be made equal and opposite. With the two in series a very cheap temperature compensated zener diode can be produced which is the equal of most of the reference diodes available commercially for quite high prices. Using two 2N2484 transistors in a common heatsink a temperature coefficient of less than 10 p.p.m. over a ten degree temperature range can be achieved for 2mA current and 12V output. Resistors of the order of 10k have been used.



Temperature compensated 'zener' diode.

E. R. RUMBO,
Chippendale,
New South Wales,
Australia.

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High-power Amplifier

A design with a bridge output stage delivering 100W into 8Ω

by Ian Hardcastle, M.A., & Basil Lane*

Regular readers of *Wireless World* will recollect an earlier article of ours describing a simple 15-W power amplifier†. This article takes the design several stages further to provide an output power of 100W.

In attempting to upgrade the original design (reproduced in Fig. 1) two alternatives could be adopted, either to use 8Ω speakers and raise the supply voltage or to use 4Ω speakers and accept a very high output current.

Table 1 lists the requirements in detail and from the data sheets of the output transistor range (TIP29A to TIP36A) it can be seen that the supply voltage equals the breakdown voltage in the case of the 4Ω version, and for the 8Ω version the supply rail is well above breakdown. In addition if we are to retain the same pre-driver the supply voltage is also well above the breakdown of the BC212L used. Any alternative device used as a pre-driver would represent a rise in cost as would be the case for the output transistors in the 4Ω version. Here the large peak current demands would necessitate the selection

* Texas Instruments Ltd.
† "Low-cost 15-W Amplifier", *Wireless World*, October 1969.

TABLE 1. Voltage and current requirements for 100 watt amplifier

load impedance	r.m.s. voltage	peak voltage	pk-pk voltage	r.m.s. current	peak current	power supply
8	28.3V	40.1V	80.2V	3.53A	5.0A	90V
4	20V	28.3V	56.6V	5A	7.06A	64V

of TIP35A/36A preceded by TIP29A/30A to give the base drive required by the output pair under peak output-current conditions. In turn, base drive for these devices, if derived from the existing pre-driver, would set a collector current requirement for this stage of around 20mA giving a power dissipation beyond the capability of the low cost small signal devices. In addition the stage gain would be reduced and by a chain reaction reduce the ratio of open loop to closed loop gain giving a high level of distortion. A way round this problem is to insert another set of drivers using a further pair of TIP29A/30A as shown in Fig. 2.

To ensure sufficient overall current gain around the quiescent point, resistors R_1 to R_4 are added to the configuration to increase the running current of the output stage to a point where the devices have developed useful gain. Since a 4Ω load is being employed the load coupling capacitor must be large and its current rating must be greater than 5A.

Finally the use of the 64-V rail means that the transistors are operating at and beyond the limits of their voltage ratings making the production of such an amplifier a risky business. Furthermore, the unregulated supply previously specified would be unsuitable since its off-load voltage could well rise too far. A fresh look at the design is obviously required.

Argument for the bridge

At the sort of power levels under consideration we are stuck with the output current and voltage requirement already mentioned and any attempt at reducing this (with say an 8Ω load) will only result in a prohibitively high rail voltage to provide the large voltage swings required. The reason for this is that the use of the capacitor coupling arrangement to the load limits these voltages to peak values of + or − half the supply voltage. What is required is an arrangement that eliminates the coupling

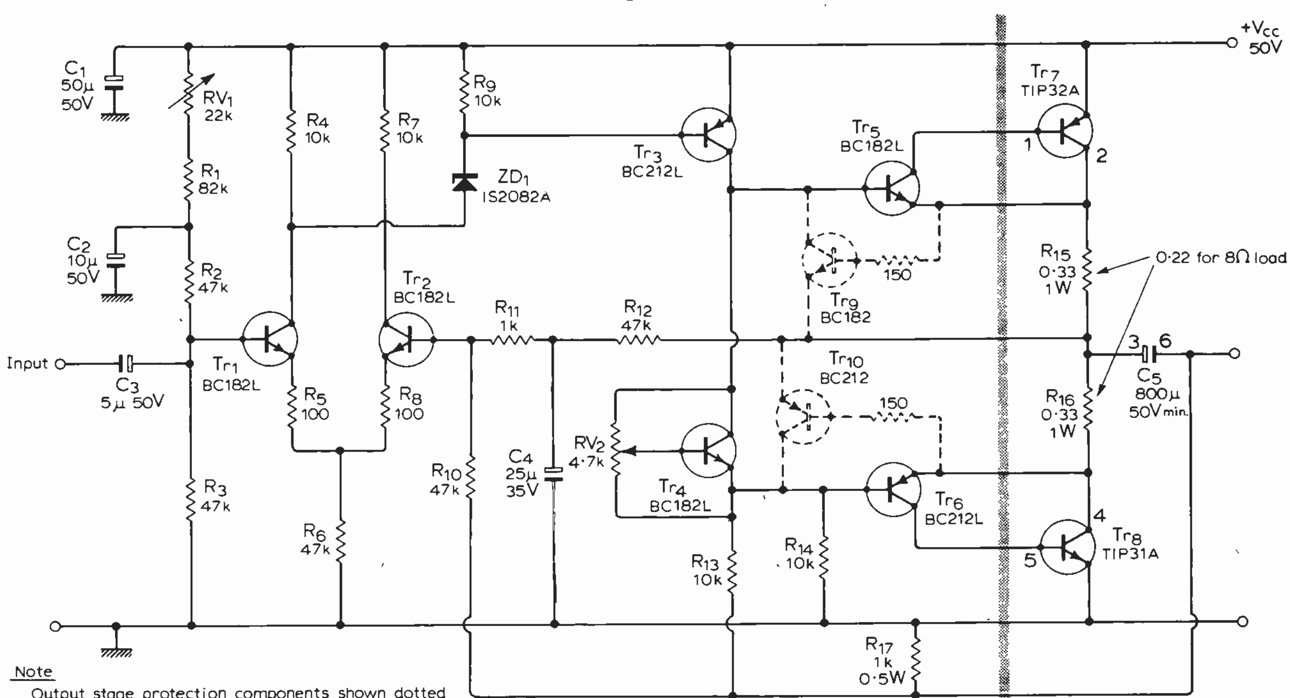

Fig. 1. Circuit of amplifier described in October 1969.

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capacitor without incurring the penalty of d.c. flowing in the loads.

Fig. 3 shows such an arrangement, where four output transistors are arranged as a bridge with the load connected across the d.c. null. If an anti-phase signal is applied to this configuration, on peak positive signals Tr_1 and Tr_4 are turned hard on, Tr_2 and Tr_3 are hard off, causing the full supply voltage to appear across R_L. With a peak negative input, Tr_2 and Tr_3 are turned hard on, causing the full supply voltage to again appear across the load but in the opposite direction. Taking another look at Table 1, we see that an r.m.s. voltage of 28.3V is required to produce 100W in 8Ω.

With the arrangement shown in Fig. 3, each half of the output stage only has to swing half of the required voltage. In this instance 14.15V r.m.s., 20.05V peak, or 40.1V peak to peak. This now allows the rail voltage to be lowered to 50V enabling the same transistor families to be used in the output stage as were employed in the original 15-W design. In addition, if at any time the power output of the amplifier needs to be raised, the rail voltage can be raised to achieve it. For example an 8Ω load with 60-V supply would provide an output power of up to 175W. Alternatively for a 4Ω load power in excess of 350W should be obtained.

Fig. 4 shows a block schematic of the arrangement to be used in this instance together with a full circuit diagram in Fig. 5.

Considering one half

Referring to Fig. 5, the left-hand power amplifier will now be described, similar arguments being applicable to the right-hand amplifier. Readers will recognize the configuration of the output stage, which is similar to that of the early 15-W design. The transistors TIP35A and TIP36A would be suitable for use in this 100-W version, but at the moment they are considerably more expensive than the metal can versions, the 2N3715 and 2N3791. It is possible that at some time in the future the plastic encapsulated versions will become cheaper,

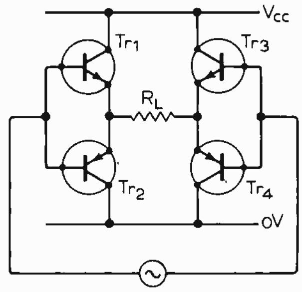

Fig. 3. Bridge output stage with the load across the d.c. null.

in which case they can be substituted without any modification. The latter transistors have a minimum current gain of 20 at 5A and hence a peak base current of 250mA is required. This is above the power dissipation capabilities of the smaller plastic encapsulated transistors which were previously used but the substitution of a pair of TIP29A and TIP30A will serve the same purpose without the need for heat sinks, since the demand is well within their free-air current and power ratings. The current gain at their peak collector current is at least 40, the peak base current required being about 5mA.

This particular type of transistor, when operated at low quiescent collector currents has a somewhat low current gain, and to obviate any difficulties likely to be encountered by variations in gain in such a situation, resistors R_{21} and R_{23} are used to shunt the base-emitter diode of the output transistors. Under quiescent conditions, the collector current of the drivers is now set at about 4.5mA giving them a current gain of around 100. A point not mentioned in the previous article dealing with the 15-W version, was the necessity for such resistors to be included to provide a path for the output transistors' collector-to-base leakage current.

Readers who have built the 15-W version, will find it advantageous to insert resistors in a similar position, of value 1kΩ.

Two additional functions are performed by these resistors in both designs; they ensure thermal stability of the output stage and by providing a path for collector-base

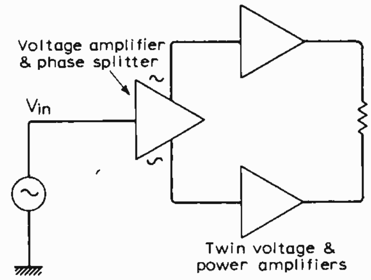

Fig. 4. Block diagram of circuit employed.

leakage, decrease the turn-off time of the transistors thus reducing high-frequency cross-over distortion and dissipation.

The familiar transistor-potentiometer biasing arrangement has been used again to supply the necessary inter-base voltage to the drivers, thus causing the output stage to operate in class AB, giving a nominal output quiescent current of 20 to 50mA. The pre-driver stage operates at a quiescent current of 10mA, this current being well within the current and power handling capabilities of the devices specified in the 15-W version. Referring to Fig. 1, the reader will see that a bootstrap resistor has been used as a constant-current sink, this resistor being connected between C_5 and Tr_4. In this instance, at the cross-over point, the input impedance of the resistor tends to fall because the voltage on the collector of the driver transistor Tr_3 changes more than the output voltage, thus a voltage imbalance occurs at either end of the resistor and the bootstrapping fails. An alternative providing a more satisfactory constant-current sink, is to replace the resistor by an additional small signal transistor suitably biased from a constant-current source. Such an arrangement is shown in Fig. 6 where the transistor Tr_9 has a constant voltage at its base provided by a potential divider chain formed by the resistors R_{11} and R_{12} with diode D_1. The voltage developed by D_1 approximately matches the base-emitter voltage of Tr_9 over a range of ambient temperatures and thus ensures that the voltage across resistor R_{19} and in turn the current sink provided by transistor Tr_9, remains constant with changes in temperature. An improvement in the cross-over distortion figures (still present in spite of the output stage operating in class AB) is obtained, since the constant-current sink is no longer dependent upon feedback from the output mid point.

This arrangement has been used for the 100-W amplifier for the reasons discussed above, and also because there is no output capacitor to transfer the output voltage from the output mid-point down to the earth rail. To provide such a point suitable for the connection of a bootstrap resistor, an extra capacitor and resistor would have to be inserted. All in all a comparison of costs between providing the additional transistor, or providing the capacitor-resistor arrangement shows that they are about equal, but the improvement in performance from using the transistor constant-current sink more than sub-

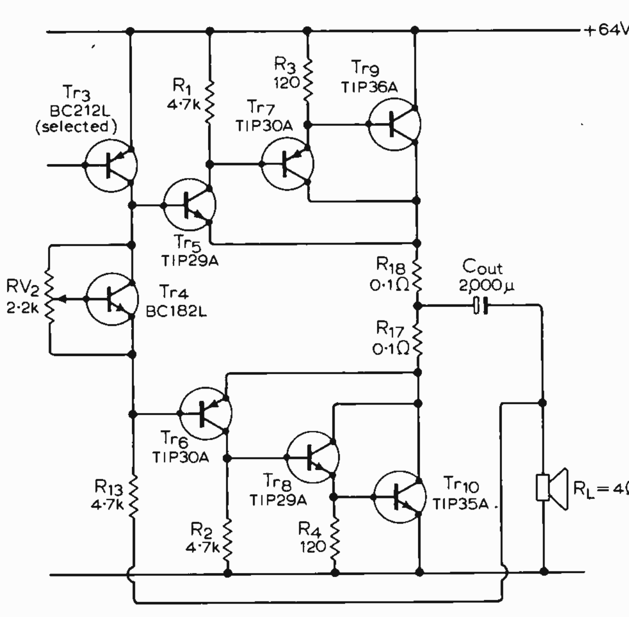

Fig. 2. Possible modification of Fig. 1 to obtain greater power.

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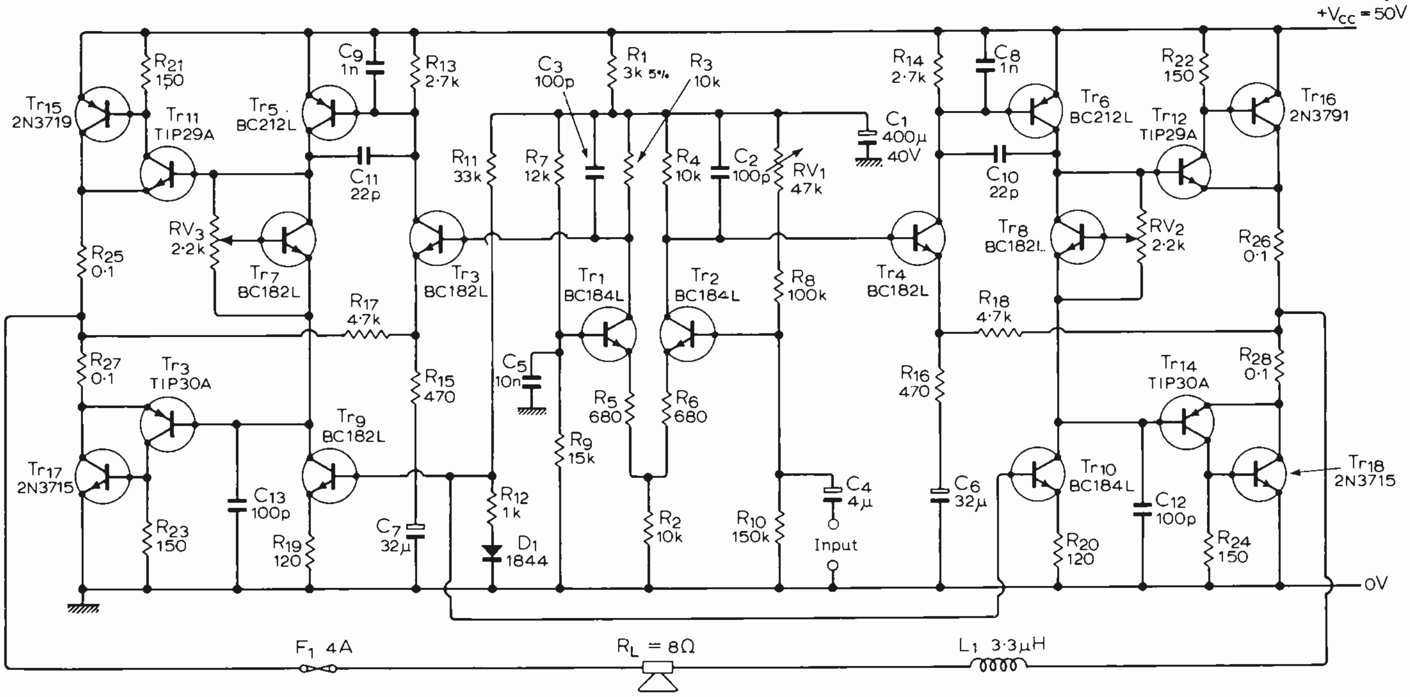

Fig. 5. Full circuit of 100-W amplifier. Resistors can be 10% types except R_1.

stantiates any reason for its selection. Additionally, small signal transistors are far more readily available than the electrolytic capacitors at the time of writing and in the foreseeable future. Since our 100-W Figuration employs two identical power amplifiers, the cost of providing the constant potential divider chain for this additional transistor has been reduced by making use of the same bias chain for both amplifiers.

Input stage

To complete the power amplifiers, a different input stage is used from the version shown in Fig. 1. The long-tailed pair has been replaced by a single transistor with its collector connected directly to the base of Tr_5, the base-emitter junction of which is partially shunted by R_{13} to reduce variation in the collector current of Tr_3 due to variations in the current gain of the driver transistor. Resistors R_{17} and R_{15}, together with capacitor C_7 provide an a.c. feedback path to the emitter of Tr_3 and set the somewhat low a.c. closed loop gain of 11. In our original article several benefits were claimed for using the long-tailed pair as an input stage. These included good stability of d.c. level of output mid-point, and high input impedance to the input and feedback circuits.

In the 100-W amplifier, the stability of the d.c. level of the two output mid-points is certainly worse than that of the 15-W amplifier. The reason for this is that changes of both the base-emitter voltage to Tr_3 and the current gain in Tr_5 will affect the d.c. output level. Under normal conditions the results of these changes due to temperature variations would be expected to be small, for example, a 20°C change will cause an output level shift of 40 mV and a change in gain of Tr_5 and Tr_6 from 60 to 300 will cause the output level shift of 400 mV. However, this is not really important since what we are interested in is any changes of voltage across the load creating a d.c. unbalance in the bridge, because at all costs large direct currents must be prevented from flowing in the load. However such currents are unlikely to occur since any temperature change affecting the left hand power amplifier will similarly affect the right hand power amplifier. Such changes in both amplifiers would simply result in a similar shift of output mid point voltage at both ends of the load, resulting in a cancellation of effect and the preservation of bridge balance.

The high impedance feature of the feedback circuits in the original design has become unnecessary in the 100-W version, since the closed loop gain has been made small.

The advantage of having low value decoupling capacitors is retained since resistor R_{15} is relatively large (470Ω) and needs only a 32μF decoupling capacitor to give a low-frequency −3dB point at 10Hz—once again comparing favourably with the 15-W version. If this arrangement were used in a simple amplifier with a gain

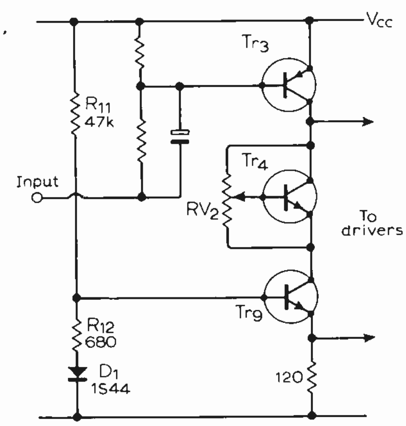

Fig. 6. Replacement of bootstrap resistor by transistor.

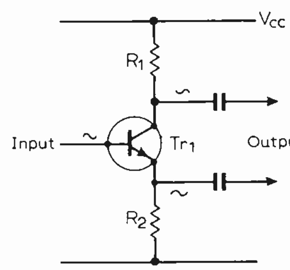

Fig. 7. Single-stage phase splitter.

of, say 100 where the feedback loop encompassed the whole amplifier, a capacitor of ten times this value would be required to give the same —3dB frequency. Despite the change of configuration from long tailed pair to single transistor the input impedance looking in at the base of Tr_3 is still high, due to the presence of the large in-phase feedback signal at the emitter. As the entire amplifier is d.c. coupled, the stability of the output mid-point voltage needs to be ensured by a d.c. reference provided at the base of Tr_3. Two alternatives present themselves at this point.

(a) To provide an independent potential divider bias chain for Tr_3 and Tr_4, the doubling up being necessary to allow compensation for individual variations in V_{BE} of these two transistors together with the potential drop tolerance in R_{17} and R_{18}.

(b) To make use of the d.c. level present at the output of the phase splitter as a reference.

The latter alternative represents a considerable simplification and has therefore been chosen.

Phase-splitter stage

Readers will note that the phase splitter used consists of a long-tailed pair formed by the transistors Tr_1 and Tr_2. Other

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alternatives could have been used, such as a single transistor with equal collector and emitter resistors (see Fig. 7) one phase being taken from each output electrode. The preference for the phase splitter was set by the following considerations:

1. Voltage gain is obtainable from a long tailed pair allowing a reduced voltage gain in the power amplifier and making a simpler design. In addition, a greater ratio of open to closed loop gain is obtained, giving lower distortion.

2. In the case of the single transistor, the impedance seen by each of the identical power amplifiers would be different since the collector output impedance of the phase splitter is higher than that of the emitter. The long-tailed pair shows no such disadvantage and provides identical drive conditions for both power amplifiers.

3. A single transistor phase splitter presents different d.c. levels at emitter and collector thus complicating the problem of direct connection to the input of the power amplifiers. In the case of the phase splitter using a long-tailed pair, connection is made to collectors at the same d.c. level for both power amplifiers.

An examination of Fig. 5 reveals that the bias arrangements for the two halves of the phase splitter are different. Resistors R_8 and R_{10} are of an order of ten times the value of R_7 and R_9. One of the main reasons for this difference is that the input to the phase splitter is fed to the base of Tr_2 and it is necessary to preserve a high input impedance (to avoid loading the output of the pre-amplifier). The source impedance seen by transistor Tr_2 when taking noise into consideration will be low, since C_4, the input coupling capacitor and the impedance of the source acts as a de-coupling network across R_8 and R_{10}. Provision of a network of similar impedance at the base of Tr_1 would mean that the source impedance seen by Tr_1 would be too high for minimum noise output, and additionally, radiated interference would be easily picked up by this device. The values chosen for resistors R_7 and R_9 provide the optimum low-noise condition, capacitor C_5 being used to de-couple the bias chain at high frequencies to eliminate noise from radiation sources.

In most conventional long-tail pair arrangements efforts are made to see that both halves of the pair are balanced in their current, voltage, and impedance characteristics. In this way such an arrangement takes advantage of the inherent self-balancing of d.c. conditions available from such a configuration. Since in this version of the long-tailed pair we have an impedance imbalance at the base of each half of the long-tailed pair, gain changes in the transistors arising from temperature shifts is likely to result in a drift in the relative d.c. voltage levels of the two bases; this in turn appears at the output of the long-tailed pair and hence throws the power amplifiers into a state of imbalance. In practice, this effect is very small due to the selected high-gain characteristics of the transistors specified for Tr_1 and Tr_2. As an example, if a minimum gain transistor's gain doubles due to any changes in temperature, a differential change in the base voltage of Tr_1 and Tr_2 of about 120μV can be expected. The resulting differential change in the output voltage of the whole amplifier is likely to be about 1.7 mV.

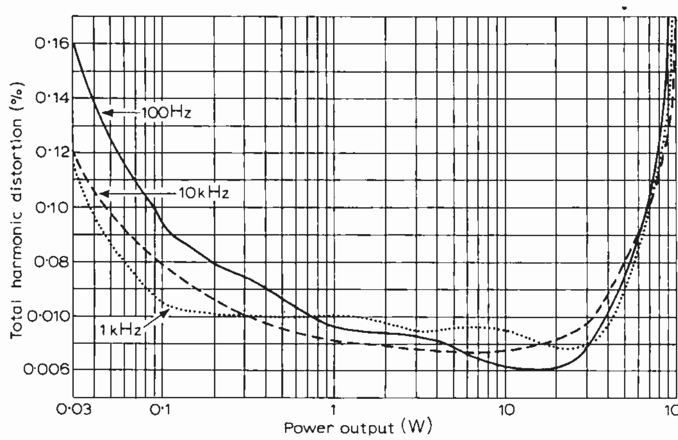

Fig. 8. Total harmonic distortion at different power levels.

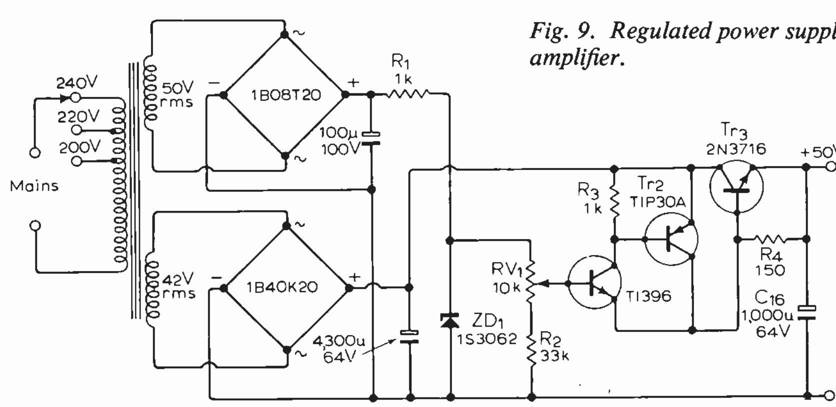

Fig. 9. Regulated power supply for amplifier.

As already mentioned, the input impedance of the power amplifier is high and thus though it shunts the collector resistors of each half of the long-tailed pair this need not be considered when calculating the gain of the phase-splitter stage. Gain is calculated by the ratio of the collector resistors R_3 and R_4 to the emitter resistors R_5 and R_6 and produces a value of 15.3. Resistor R_1 is used to connect the complete phase-splitter stage to the positive rail. Since it is decoupled by C_1, ripple and noise is prevented from appearing at the inputs of the power amplifiers. Since the power amplifiers themselves are fully hum-proofed, noise appearing at the output source is extremely low. One could argue that the de-coupling of R_1 is not strictly necessary, since any hum appearing at R_3 and R_4 will be in-phase and will thus appear at the output of each power amplifier still in phase, constituting a null in the load. This condition would only be true were the voltage gain of each power amplifier equal. In practice this is somewhat difficult to ensure, and anyway the cost of the additional decoupling is small ensuring low noise as well as incidentally improving the high-frequency stability of the amplifiers.

A further consideration in the selection of the value of the collector resistors in the phase splitter, was the requirement to provide the appropriate d.c. level to allow direct coupling into the power amplifiers. This voltage is 28.5V allowing the mean d.c. level of the power amplifier output mid-points to be set at approximately +26.2V. Since a 50V supply rail has been specified, one might expect this output mid-point voltage to be nearer 25V, however, the output voltage of the power amplifiers cannot swing as close to earth as it can to the positive rail because the emitters of the current-sink transistor Tr_5 and Tr_6 are set at 1.5V.

As has already been stated it is undesirable for differences in the d.c. mid-point voltage levels to occur because of the resultant large direct currents flowing in the load. Component tolerances tend to create such a situation which is to a large extent compensated for by adjustment of the base voltage of transistor Tr_2 using the potentiometer RV_1. This potentiometer allows the adjustment of the output mid-point voltage in the right hand power amplifier to the same level of the output mid-point voltage of the left hand power amplifier.

If low-cost wide-tolerance resistors are used throughout the system it may be found that the adjustment of RV_1 is insufficient to correct any imbalance in the output mid-point voltage. In such an instance resistor R_7 should be reduced to 10kΩ and a 4.4kΩ potentiometer connected in series with it giving an additional adjustment to the left hand amplifier's output mid-point voltage.

High-frequency stability

Due to the high cut-off frequency of the transistors used throughout this design the whole amplifier has gain in the mega-

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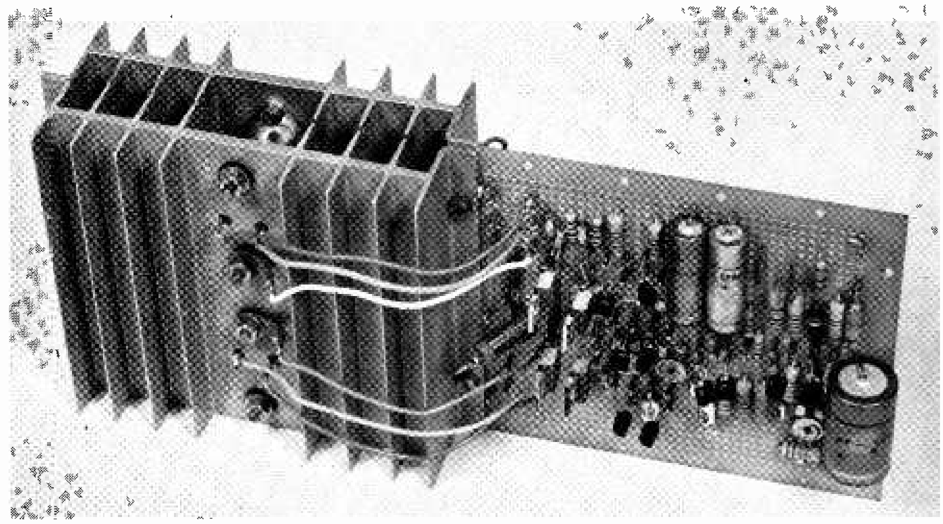
hertz region and is therefore unstable. Where straight resistive loads are to be connected stability may be ensured by connecting 100 pF capacitors between the collector and base of Tr_5 and Tr_6. However, if the load displays parallel capacitance the amplifier is again unstable and capacitors C_2 and C_3 have to be added and C_{10} and C_{11} increased to 330 pF. Unfortunately, although this stabilizes the amplifier, the maximum undistorted output frequency is limited to about 10kHz. Naturally this is undesirable and to alleviate the situation and extend the frequency response beyond 20kHz, capacitors C_8, C_9, C_{12} and C_{13} have to be added. In addition, a small inductor L_1 completes the stabilization arrangement. This inductor can be made by winding 25 turns of 26 s.w.g. enamelled copper wire in a single layer on a high value 1W resistor.

It may seem that rather a large number of components have been used to ensure the high frequency stability of the amplifier, but this situation has arisen as a result of the large open-loop gain of the amplifier and the large current and voltage swings involved. Distortion characteristics for the amplifier are shown as a graph in Fig. 8. The main performance figures are given in Table 2 and intermodulation distortion readings in Table 3.



A complete amplifier. The four electrolytic capacitors used in the circuit are small enough to be wired directly to the Lektrokit board.

Construction

As is the case for the 15W amplifier described in our previous article the layout of this amplifier is not very critical. However, some simple rules should be observed in the layout of the output stages. Transistors Tr_{11}, Tr_{12}, Tr_{13} and Tr_{14} should be mounted as close as possible to one another on the circuit board.

Short wires should be used to connect these stages to the output transistors. If possible separate leads from the emitters of the n-p-n output transistors should be taken directly to the reservoir capacitor. Alternatively if the power supply and reservoir capacitor are not close to the power amplifiers these leads could be connected to the circuit board at one point close to the driver stages with the negative lead from the power supply. A similar arrangement should be made for the emitters of the p-n-p output transistors.

As one may expect from an amplifier of such an output there is considerable heat dissipation in the output transistors. In fact this amounts to about 25W, so efficient heat sinking should be employed. One heat sink should be used for each output pair.

A method of component layout employed in the prototype was to make use of the plain perforated Lektrokit board.

TABLE 2. Performance Figures

power output	100W
power output (t.h.d. = 10%)	150W
frequency response (± 1dB)	10Hz—20kHz
signal-noise ratio (R_{source} = 600Ω)	89dB
signal-noise ratio (R_{source} = 10kΩ)	84dB
input impedance at 1kHz	56kΩ
output impedance at 1kHz	0.08Ω
nominal input voltage (P_{out} = 100w @ 1kHz)	180mV
power-supply current (at 100W & 1kHz)	3.5A r.m.s.

To avoid the inevitable rage and frustration in the event of turning your amplifier on for the first time and seeing large black clouds of smoke rising, the following setting up rules should be closely observed:—

(1) Terminate the input with a 10k resistor. Connect a voltmeter across the open circuit output terminals.

(2) Set RV_2 and RV_3 such that their wipers are at the end of their tracks connected to the collectors of Tr_7 and Tr_8.

(3) Set the wiper of RV_1 at its mid-point.

(4) Connect a 3W 5k rheostat and a 250mA meter in series with the positive supply lead. Set the rheostat to 5kΩ and connect the amplifier to its power supply.

(5) Turn the power supply on, and slowly reduce the resistance of the rheostat to zero. Should the current drawn exceed 150mA turn off and check the wiring.

(6) Connect a voltmeter across the output points and adjust RV_1 until the output voltage as read by the meter reads 0V (±50mV).

(7) Switch off. Disconnect the rheostat and current meter, and connect a suitable resistive load. Switch on and re-check the differential output voltage. Re-adjust as necessary and then switch off.

(8) Connect a 100mA current meter in series with the emitter of Tr_{15}. Switch on and adjust RV_3 until 50mA is indicated. Switch off.

(9) Move the meter connections to the emitter of Tr_{16} and switch on. Adjust RV_2 until the meter indicates 50mA. Switch off, and remove the meter.

(10) Switch on and re-check the d.c. differential output voltage. The amplifier is now ready to be used.

Overload protection

In an amplifier of this type, some form of protection is essential and the method shown in our earlier article is suitable for use here, though the number of items is doubled. In addition the fitting of fuse F_1 is regarded as imperative in view of the fact that the output stage can still overheat under long-term overload conditions. The value of this fuse should be 4A.

TABLE 3. Intermodulation Distortion

output frequencies & amplitudes				intermodulation distortion
Hz	V	Hz	V	(%)
1100	14.4	900	14.4	0.16
11k	14.4	9k	14.4	0.23
10k	24.0	1k	6.0	0.14
10k	6.0	1k	24.0	0.14
10k	24.0	120	6.0	0.15
10k	6.0	120	24.0	0.15
1k	24.0	120	6.0	0.135
1k	6.0	120	24.0	0.145

Power supply

It is just feasible that an unregulated power unit could be used to supply this amplifier, but it is important that under no-load conditions the voltage does not rise above 60V and at full power it does not drop below 50V. Large currents are drawn from the supply making these requirements difficult to obtain and most readers may favour a regulated unit. A suitable design is given in Fig. 9, where the 50V winding (rated at about 10mA) provides a supply which when rectified and smoothed acts as a constant voltage source. Resistor R_1 acts as a constant current source for the zener diode ZD_1 and the parallel arrangement of RV_1 and R_2 in series provide a small measure of adjustment. Transistors Tr_1 and Tr_2 provide the necessary current gain to minimize the effect of the base current swing of Tr_3 on zener diode ZD_1. The dissipation in Tr_{20} reaches a maximum of about 35W, and so it should be mounted on a heat sink having a thermal resistance of less than 3.5°C/W. The leads from C_{16} to the emitters of the four output transistors should be kept as short as possible.

The three electrolytic capacitors used in this power supply cannot be eliminated or safely reduced in size. Separate transformers can of course be used to provide the 50 and 42 volt secondaries.

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Avionics at Farnborough

Some impressions from the electronics show

Have you ever thought of all the things that occur when a modern aircraft is displaced slightly from its flight path by an air disturbance or for some other reason? As an airline passenger one is inclined to think only in terms of seating comfort and cabin service and to hope that the wings won't fall off or that engines won't pack up. It is interesting to look at the aircraft as a single unit, that is, to roll all the sub-systems into one.

The small disturbance mentioned above will cause a host of transducers of many types to give out a signal or to change their existing outputs. Innumerable gyros will measure the amount, rate and direction of the disturbance; accelerometers will measure the size and direction of the accelerations and decelerations involved in three dimensions; aneroid and other capsules will measure change in height, airspeed and rate of climb or dive; direction-finding aerials move slightly to stay aligned on the beacon they are tracking; doppler beats are generated proportional to the rate of the disturbance; flux valves measure the angle of the earth's magnetic field relative to the aircraft, and so on.

This flurry of signals is fed to a number of digital, analogue and hybrid computers. Trains of pulses are generated during the computations, analogue signals change in phase and amplitude in such a way as to describe the amount and direction of the displacement, scores of small motors turn to either drive pointers or to assist in the calculations by turning a shaft which mimics the aircraft's angle in a particular plane.

As a result of all these actions the gyro compass now indicates the new heading (true or magnetic) and updates all the navigation computers which will provide a read-out of position; the outputs of the accelerometers are changed into signals representing velocity and to distance flown in a given direction for the navigation system; the doppler radar will be giving an indication of ground speed and drift from which true track can be computed; and signals will be fed to limited authority motors in the flying control system to cause the hydraulics to apply control surface movement to limit the size of the displacement. With automatic control the automatic pilot will apply control surface movement, after taking into account the flying characteristics of the particular aircraft. in such a way as to correct the displacement practically before it occurs. During even a disturbance these and many other actions occur. For instance, signals from different sources, but describing the same thing, are compared and an error signal may be generated to correct one of the sources.

Apart from these systems there is the radar blind-landing and communication equipment as well as the electrical, electronic, hydraulic, pneumatic and mechanical systems needed to control and monitor cabin pressurization, fuel management, undercarriage, flying controls, flaps, dive brakes, engines, temperatures, etc., etc.

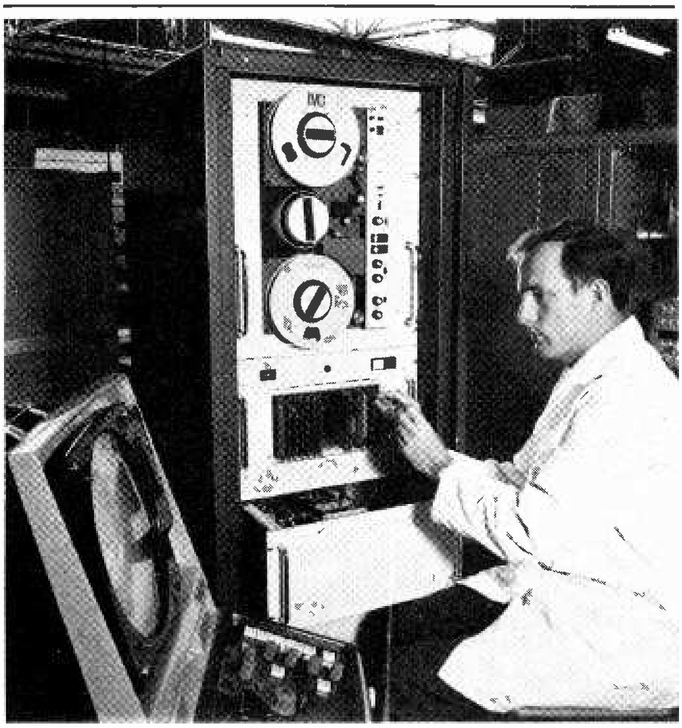
All these systems have to be powered and the prime movers for this are of course the engines. The outputs of the engine-driven alternators have to be controlled in frequency and amplitude and distributed throughout the aircraft so a great deal of electrical equipment is required.

The preceding few paragraphs were presented so that the reader who is not familiar with aircraft systems could get some idea of how complex a modern aircraft is. Not all systems were mentioned and if a military aircraft had been taken as an example. instead of a civil one, the various weapon aiming systems would have increased the complexity by a very large factor.

Britain's avionic industry can design, develop and build all the systems necessary including the engines and the airframe for any aircraft that is technically feasible by today's standards if it had the money to do so. The only other countries to have such a complete capability are Russia and America.

This lead cannot be maintained unless the industry designs and builds aircraft and at present its hopes are placed on perilously few products. Topping the list is of course the Anglo-French *Concorde* which stands to win £4,000M worth of export orders if it comes up to its design performance. There is also the Anglo-French *Jaguar* and the European *Multi role combat aircraft* (MRCA) controlled by an Anglo, German, Italian company called Panavia. There is the successor to the BAC one-eleven, the three-eleven, which is a short/medium haul jet airliner capable of carrying about 250 passengers which could be ready by mid-1975. About £200M is needed by BAC to get this project off the ground. Also awaiting money and policy decisions are two vertical take-off airliner projects the Westland WE.01/02 which employs a helicopter principle and the Hawker Siddeley H.S.141, a modified Trident with jet powered vertical take-off. If the British aircraft industry is to stay with the leaders it must have the money even if the costs are offset to some extent by Anglo-European co-operation. The benefits for the country in terms of exports could be enormous.

Not only can the British aircraft industry build aircraft but, as was



Designed primarily for air traffic control use by Solartron the interface equipment shown allows a standard television tape recorder to be used to record radar data for subsequent analysis on a radar display.

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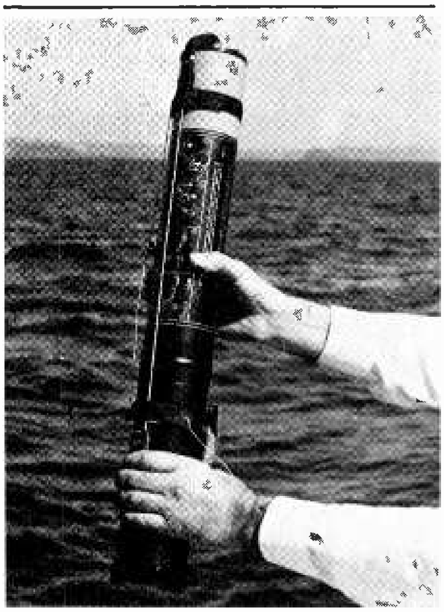
exemplified by the static exhibition of Farnborough, our own electronics industry is fully capable of meeting all the demands likely to be made upon it for the equipment so essential to today's aerospace technology.

At the show Marconi introduced a new computer called Myriad-3 which is designed around t.t.l. logic and is intended for handling radar data.

Myriad-3 is a parallel binary fixed-point machine with single address operation having a word length of 24 bits with facilities for double-length working. Extensive input/output facilities are available in the form of optional plug-in modules. Most single-length instructions take about 3.2μs and the more complicated instructions, such as multiply, take around 11μs. Storage capacity is up to 256,000 words and two types of store are available; a 32k word store with a 650ns cycle time and a 16k word store with a 1.6μs cycle time. Because the store timing is controlled from within the store unit, it is possible to have both types of store in a system, with the slower (and cheaper) store carrying less urgent information.

The basic machine comprises a number of modular 'building bricks'. The main units are the central processing unit, store units, power units, operators' control unit, programmers'/engineers' control unit and a highway extension unit. A master peripheral unit, designed to accept standard control modules, accommodates the input/output control circuits for peripheral devices.

Among the many demonstrations on the Min-tech stand was the use of glass fibre light guides in an experimental optical communications system employing a gallium arsenide diode as the light source and an exhibit which showed how a c.r.t. and associated keyboard could be used on board an aircraft to communicate with a navigational digital computer.



This floating beacon manufactured by Burndept, which has already captured orders worth £50,000 from Norway, operates on a primary battery and transmits on both the civil and military aviation distress frequencies (121.5 and 243MHz) for a minimum of 48hrs at 0°C. It measures 670 × 85mm and weighs 2.89kg.

A personal survival beacon, Sarbe 5, introduced by Burndept for military use. It allows two-way speech and a built-in test system.

Visitors to Farnborough were able to navigate their way round the airfield without moving from the Sperry pavilion. To do this, they went for a simulated run in a Land-Rover equipped with the Sperry Vehicle Navigator and Map Display which is currently being offered for airfield navigation duties.

The Land-Rover was driven in the normal way; the wheels revolving against a dynamometer to provide a speed input for a computer. Turning the steering wheel produced heading change inputs for the computer, in place of the gyro or flux-valve normally employed. A map display unit was mounted on the bonnet for the demonstration and the position of the vehicle was indicated by hair line cursors. A digital computer monitored the process.

The helicopter version of the Solartron Simfire seen at the exhibition is one of a family of direct fire weapon effects simulators developed as aids for combat training. The system has two basic roles. As a method of training helicopter pilots in making the best use of ground and cover to avoid ground fire; and as a means of assessing the vulnerability of helicopters to ground fire, for operational analysis purposes. A laser-beam projector is aimed by the operator at a helicopter target. When the operator 'fires', a stream of pulses is directed towards the target. Detectors on the helicopter receive the laser beam impulses and illuminate indicator lamps which show the pilot that he is under attack. The direction of the attack is indicated by a lamp to the pilot who can then take avoiding action. If the helicopter target is held in the centre of the beam for a period exceeding one second, the helicopter is 'killed'.

Texas Instruments, as part of their Klixon range, were showing what they describe as a solid-state vane switch—a contradiction of terms if ever there was one. The device is designed to sense when a supply of cooling air is stopped or reduced in a piece of equipment. The sensing element is a positive temperature coefficient thermistor built into a probe.

Sperry were showing a c.r.t. display intended to replace a number of instrument panel dials. The c.r.t displayed characters and symbols to give aircraft attitude, flight director information, altitude, airspeed and weapon aiming information either individually or as an integrated display.

Decca Navigator, who were rightly shouting about the recent order by America's Eastern Airlines for the Omnitrac Area Navigation System, were also displaying a new back-projection navigation display.

Elliott Automation Radar Systems introduced a new nose radar of advanced design capable of terrain following, mapping, ranging, and air-to-air operation in strike aircraft while Elliott Flight Automation had on display a digital automatic flight control system for civil use.

Ultra Electronics were displaying a personal radio location beacon designed to replace the larger beacon now in service with the armed forces. Called Sarah-B the beacon is capable of twin-frequency working and it is built using thin film circuit methods. It measures 110 × 75 × 25mm and weighs less than 20 oz. Sarah-B transmits a tone on the distress frequency of 243MHz and can be used for two-way communication over short ranges on 243 or 282MHz; battery life is in excess of 24hr.

That completes our quick glance at a small percentage of products that were on show at Britain's biennial avionics supermarket at Farnborough.

Among the new items introduced by E.M.I. was a video map generator (Type VM101) which is designed to be an inexpensive adjunct to radar simulators or for use with real time radar displays for simulating stationary or moving clutter. It will also produce a video map. Patterns are generated by a flying spot scanner.

Elliott's stand contained a vast assortment of equipment for both military and civil applications and the Digital Systems Department introduced an airborne display control unit. Facilities provided by this unit include alphanumerics, lines and circles with 10-bit angular and positional accuracy. Individual symbols can be made to flash or be made brighter, and several displays, each showing a different picture, can be driven simultaneously.

Plessey had an engine vibration monitor on show which was designed for *Concorde*. The system consists of eight piezo-electric transducers, one mounted on both the fore and aft bearings of each of the four engines. An eight-channel display of the flight deck gives immediate indication of increased vibration, often the first indication of serious engine trouble. Another item on the Plessey stand was a stall warning unit for use on the Hawker *Harrier*. Intended for use when the aircraft is hovering, the unit senses an incipient stall condition and warns the pilot by shaking the rudder bar.

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London Broadcasting Convention

Some developments in broadcasting seen at the September convention

With over 80 technical contributions the biennial International Broadcasting Convention in London in September attracted 1000 delegates. Most of the papers were about television—the big stride nearly a year ago when the U.K. passed out of the 405-line monochrome age into the 625-line colour age overshadowing developments in sound broadcasting.

Automation in broadcasting was one of the major themes of the convention and a good 25% of the papers dealt with some aspect of automatic control of equipment, ranging from closed loop servos for particular operations (e.g. automatic registration of colour-separation images in television cameras) to a vast data processing system using several digital computers that controls the whole of the NHK (Japan) broadcasting network and even encompasses the audience. All this is becoming necessary, apparently, because of the increasing scope and complexity of broadcasting operations and equipment —human control is beginning to prove inadequate—and, in turn, because of the excessive cost of training and using technicians for these operations. The subject has become so important, and the automation techniques are so varied and complicated, that we intend to make a special report on this aspect of the I.B.C., to be published in the November issue of *Wireless World.*

Satellite broadcasting has its problems, some of which have been discussed before. One, discussed by P. L. Mothersole, is that of mains isolation in television receivers; a problem when thinking in terms of adding a converter to receive satellite broadcasts (such an add-on unit would convert from f.m. to a.m. as well as frequency). But with a new receiver design a convenient way around this is to have an insulated winding on the line scanning transformer, at earth potential, providing the low-voltage supply that would be required for a converter. A possibility which Mr Mothersole foresaw is the use of a communal aerial and converter system. This would avoid mounting and aligning dish aerials in domestic installations, and could also help to keep cost down.

An interesting point touched on by J. Redmond, director of engineering, B.B.C., in his address was that of *sound* broadcasting from satellites in the h.f. band. A study has shown this to be technically feasible in the 21 and 26MHz bands. The satellite would probably be in a low-altitude orbit in such a scheme, because of the high power needed from a geo-stationary satellite—of the order of tens of kilowatts. A power in the region of 1kW is thought feasible for a low-orbiting satellite, and no doubt folding aerials would not present undue difficulties. Snag is the expense of maintaining orbiting satellites and with the limited number of sets able to receive 21 and 26MHz broadcasting authorities might feel there would not be much return for an outlay which could be as high as £10M p.a.

Speech clipping is used to increase the average modulation level in transmitters without overmodulation by reducing the peak-to-average energy ratio in speech waveforms. Normally this is done at speech frequency and results in trapezoidal waveforms. As well as imposing a linear phase response on the modulator this means a much wider bandwidth is needed and also power supply requirements would need to be increased. To avoid modifying older modulators to cope with the clipped speech, frequency translation can be used to ease filtering. The problem is of course that some harmonics of low-frequency speech signals in the clipped waveform are still within the speech spectrum and can not be filtered out. By modulating the speech signal on to, say, a 20kHz carrier (suppressed) and then clipping, filtering is made easy. This is because the highest upper-sideband frequency will be at, say, 25kHz—well away from the lowest first harmonic at about 40kHz. The signal is then demodulated in a product detector. With this kind of system, Radio Liberty (Federal Republic of Germany) has produced a clipper with harmonic distortion of less than 1%, even with an overload of 16dBm.

Studio acoustics. The well-known technique of making models to predict acoustic properties of concert halls has been applied to studio design and fault correction by the B.B.C. Pioneered by Spandöck at the University of Munich, this is the first time it's been used with a good signal-to-noise ratio—at least 50dB—allowing models to be listened to seriously. This method imposes severe constraints on experimental apparatus, severity depending on the scaling factor. Not least of these is the problem of air absorption, and to get the sound attenuation right a relative humidity of 3% is needed. The usual way of doing this is with silica gel but it takes several weeks to achieve the required dryness. In the B.B.C. experiments H. D. Harwood and A. N. Burd have reduced drying time to half a day for the first drying and to 15 minutes for subsequent drying by using an artificial zeolite.

Using a scale model of a studio reverberation time in the model agreed with that in the studio to within $\pm$ 10% throughout the whole frequency range, this difference being inaudible. The model was also a success subjectively in that addition of an 'orchestra', made from polystyrene backed with felt, to the model was clearly apparent in recordings made from the model. An unexpected change in *tonal quality* from the model occurred when this orchestra was added, as well as the expected change in reverberation time. On checking against recordings made in the real studio this change was also apparent and although there is no obvious theoretical explanation for this change the fact that the model predicted it is a very satisfying validation of the modelling technique.

Another interesting development in studio design is the realisation that reverberation time need not be independent of frequency. Absorbers needed to keep low-frequency reverberation time low (say 0.35s, the recommended value for talks studios) are expensive to make and install, bulky and make modifications difficult. Availability of a recently constructed studio with removable bass absorbers made experiments on how much bass r.t. could be increased possible. As a result of various subjective tests,‡ it's now recommended that r.t. can rise to 0.4s at 500Hz, 0.47s at 125Hz and 0.74s at 62.5Hz.

A discussion of new developments in television will be published in the November issue.

‡ Spring, N. F. & Randall, K. E. "Permissible bass rise in talks studios". *B.B.C. Engineering,* July 1970, pp.29-34.

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Domestic Receivers

New techniques seen at the London trade shows

Television

The design of domestic receivers is a compound of two elements: the available market and the available technology. To be more precise, the sales people are not interested in the electronic contents of the box provided the complete product is attractive for renting (and, to a smaller extent, owning); technical features, whether real or cooked up, are seen merely as fodder for sales promotion campaigns. The engineers have at their disposal new devices and techniques and would like to use them to the full in elegant, tidy circuits, but, being restricted by cost, usually end up with a design which to them is an unsatisfactory compromise.

What the engineers have in the back of their minds, as a sort of *Volksfernsehempfänger,* is a flat box producing coloured pictures and containing nothing more than a few integrated circuits. We are certainly on the way towards this ultimate, leaving behind as we go valves, round tubes, monochrome pictures, hybrid circuitry, dual standards and discrete components. Meanwhile, in our present confused state of transition, there exist side by side monochrome and colour, hybrid and semiconductor circuitry, dual-standard and single standard, picture tubes of different sizes and shapes, single chassis and modular construction, together with a variety of "furniture" styles and electronic circuit techniques which seem like tricks because they are individual to particular manufacturers and not yet standard. In this confused situation it is difficult to decide what is "typical".

To begin with the most obvious thing, the picture tube, the six main sizes available last year (19in, 20in, 22in, 23in, 24in and 25in) have now been supplemented by two more, 17in (monochrome) and 26in (colour). There are in addition the various portable battery sets with screen sizes of 12in and smaller. The range, however, is not as embarrassingly large as it seems, as the 19in and 23in monochrome tubes have been superseded by the 20in and 24in sizes respectively, and the 25in colour tube, though still being used in sets, is due to be replaced eventually by the 26in type. The 17in monochrome tube really takes the place of the 16in type that has been widely used in transportable and small "second sets". Thus one sees a general up-grading in size all round. Monochrome tubes have deflection angles of 110°, colour tubes 90°; but colour tubes with deflection angles of 110° and, consequently, a few inches off their length, are available on the Continent and it is only a matter of time before they appear in British sets.

The distribution of valves, transistors and i.cs in receiver circuitry varies considerably from maker to maker. British Radio Corporation and Rank Bush Murphy have all-transistor colour sets which are virtually the same as last year's designs, and a new transistor colour receiver has been introduced by Philips. In the hybrid receiver circuits, colour and monochrome, it is common practice to use valves for the line and field scanning oscillators, for the associated output stages and for the sound output, but there is some variability with other stages. In a Decca single-standard monochrome chassis, for example, valves are also used for the 1st and 2nd i.f. stages, video amplifier, sync separator and line pulse discriminator. This design is, however, very much up with the times in having an integrated circuit, the Motorola MC1351P, for the sound intercarrier i.f. amplifier and f.m. detector (followed by a transistor a.f. output stage). Some current monochrome sets are still all-valve designs except for the u.h.f. tuner. Generally speaking the U.K. manufacturers are not bothering to put design effort into producing all-transistor monochrome circuits, which will only be found in the small imported portable sets.

Integrated circuits, last year restricted to the intercarrier sound and colour decoding functions* have now blossomed out somewhat, and in the new Philips colour receiver mentioned above there are four: (1) intercarrier sound; (2) stabilization for the tuner; (3) video pre-amplification, line gated a.g.c., sync pulse separation, noise protection for a.g.c. and sync channels; and (4) PAL switch, identity circuit and colour-difference demodulators. These devices are available from Mullard under the general type-number prefix TAA. But the general run of set makers is still rather wary about using i.cs, mainly for the reason that, as distinct from the digital computer field, there is a lack of standardization and "second sourcing" in the manufacture and supply of devices.

*"Colour Receiver Integrated Circuit" *Wireless World*, August 1968, p.263.

Two noticeable trends in circuit design this year are the use of variable-capacitance diode tuning and the replacement of wound components by equivalent circuitry using transistors, resistors and capacitors. Variable-capacitance diode tuning on both u.h.f. and v.h.f. has been used fairly extensively for some time in Continental television tuners but this is the first time it has appeared in sets on the British market. Examples of the technique were seen in receivers by Philips, Ekco, ITT and Teleton. The reason why varicap diodes, as they are called, have not been used here is a technological one and not, as might be thought, because U.K. makers were slow to appreciate the advantage of diode tuning. It is to do with the way in which channels are allocated.

In the U.K., u.h.f. channels are assigned to transmitters in groups of four. In most cases the four channels are n, $n + 3$, $n + 6$ and $n + 10$, or n, $n + 4$, $n + 7$ and $n + 10$, n being the lowest channel number for each station. For example: the channels assigned to Emley Moor are 41 (n), 44 ($n+3$), 47 ($n+6$) and 51 ($n+10$). Those assigned to Winter Hill are 55 (n), 59 ($n+4$), 62 ($n+7$) and 65 ($n+10$).

It will be seen that in each case the highest channel is spaced 10 channels from the lowest, which means that with the 8-MHz channel spacing laid down by the European frequency plan the highest channel is 80MHz away from the lowest. Now, for a receiver tuned to the lowest channel, the highest channel becomes the image frequency. This is because the agreed vision i.f. is 39.5 MHz. If the oscillator is tuned "high" as is usually the case, the image frequency is 39.5 Hz "on the other side of" (above) the oscillator or 79 MHz above the wanted station.

This figure is only 1 MHz short of the 80 MHz separation between n and $n + 10$ and a signal radiating at $n + 10$ if allowed to enter the pass band of a receiver tuned to

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n channel would beat with the oscillator and cause severe interference at i.f., unless measures were taken to eliminate or greatly attenuate signals at image frequency. In the case of an oscillator tuned "low" the wanted channel and image channel would be in reverse, i.e. $n + 10$ and n respectively. The presence of interference channels at image frequency is unavoidable because four of the forty-four available channels have to be allocated to several hundred stations, which necessitates the same channels being shared by many stations. The two stations quoted already share their channels with up to six others.

It is arranged for stations sharing channels to be sited geographically remote from each other, and mutual interference can also be alleviated by the use of directional aerials. Despite this, u.h.f. tuners have to be designed with specially selective r.f. circuits to reject signals outside the pass band, particularly those at image frequencies. By arrangement between BREMA (the association of set manufacturers) and the broadcasting authorities the specification for the magnitude of rejection or attenuation of image frequency signals is fixed at 53 dB. All makers try to achieve this.

Using valves and three variable tuned r.f. circuits controlled by a ganged capacitor, 53 dB image rejection is relatively easy to obtain over the u.h.f. bands. However, because of production spreads it is difficult to track the capacitance of four varicap diodes (three r.f. tuning and one oscillator) over the required frequencies to obtain the specified selectivity.

On the Continent, in Germany for instance, transmitter density is much less and the receiver i.f. is different (38.9 MHz vision), therefore the problem does not arise. The German specification for image rejection is 40 dB. This less stringent requirement allows the designer to use one less r.f. tuned circuit, and the incorporation of three varicap diodes instead of four, making tracking relatively easier.

In the varicap-diode tuner designed by Philips (Fig 1) the basic problem has been overcome in two ways. First, matched sets of five diodes with similar voltage/capacitance characteristics are selected for individual tuners. The fifth diode is employed to equalize the oscillator voltage over the tuning range. Secondly, a special image frequency rejection tuned circuit is built in. This is coupled to the aerial input inductor in the emitter of the r.f. amplifier transistor. It is automatically tuned to the image frequency as the aerial input is tuned to the wanted station.

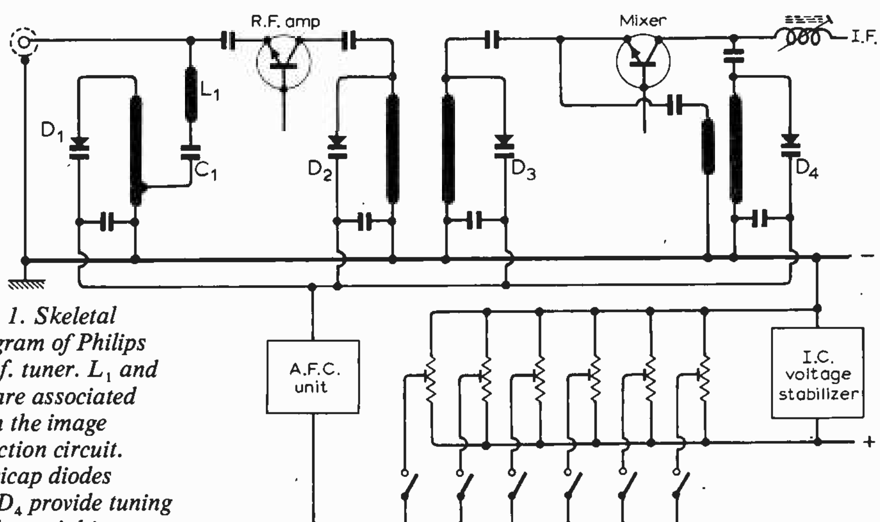
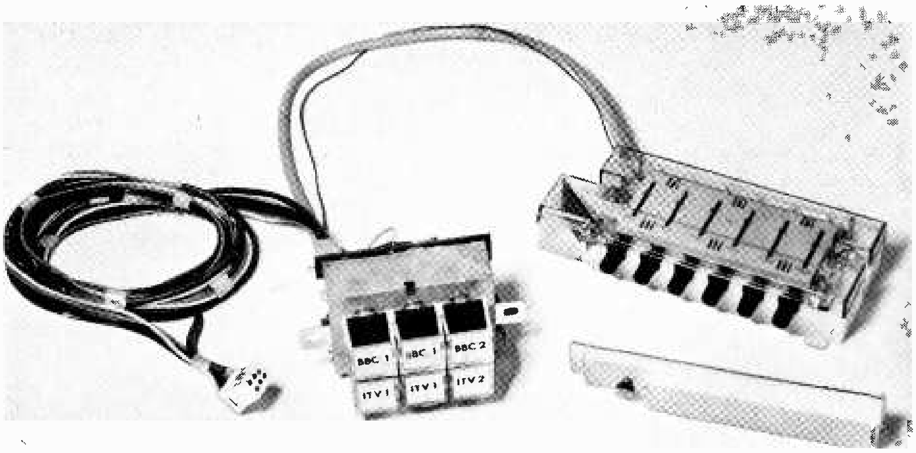


Fig. 1. Skeletal diagram of Philips u.h.f. tuner. L_1 and C_1 are associated with the image rejection circuit. Varicap diodes D_1-D_4 provide tuning of the aerial input, bandpass primary and secondary, and oscillator respectively.



Push-button channel selector (left) and pre-set potentiometers (right) of the Philips varicap-diode tuner (Fig. 1), as used in the Philips G8 colour television receiver.

The big advantage to be had from using varicap diodes is, of course, that tuning is possible by means of a single simple potentiometer. The Philips receiver uses six preset potentiometers, each associated with a press-button switch, and so provides a six-programme station selector.

Two additional bonuses in the hands of the set designer are the smaller size of the tuner unit itself and the realization that the unit can be mounted anywhere in the cabinet. The press-buttons need be only simple switches remotely connected to the tuner via low-voltage wiring. In the Philips receiver the tuner unit is actually mounted on the i.f. printed panel and it appears in the unusual position in the centre of the chassis. The effort required to operate the "programme-change" buttons, as Philips call them, is surprisingly light when compared with mechanical systems.

The varicap diodes, being voltage-dependent devices, are susceptible to variations in supply voltage. Since any variation in supply voltage would have a serious effect on tuning accuracy it has been found necessary to stabilize, not only the receiver power supply, but also the voltage line feeding the tuner itself. The tuner incorporates a simple i.c. stabilizer (see above).

In order to take station tuning out of the viewer's hands and make it into a preset adjustment, the six tuning potentiometers are concealed by a plastics cover at the rear of the receiver. Any tendency for the tuning to drift during reception is counteracted by an a.f.c. circuit which provides the large pull-in range of 1-2 MHz either side of the correct tuning point. To prevent the receiver being "captured" by a strong transmission during programme changes the a.f.c. is cut while a button is pressed.

Avoidance of wound components was illustrated by a technique used in an ITT colour receiver, type CVC5, which is based on a Schaub Lorenz design that has been available in Germany for about a year. The circuit has *R-Y* and *B-Y* synchronous demodulators that dispense with the usual transformers needed to drive the electronic switching circuits (see for example, *W.W.* May 1969, p.223). The arrangement used is shown in Fig. 2. The demodulators proper function in the manner of a symmetrical phase discriminator. To one side of this is applied the subcarrier reference oscillation (for *R-Y*, via the 62pF capacitors; for *B-Y*, via the 560Ω resistors), while to the other side is applied the signal voltage to be demodulated (at the common point of the 33pF capacitor in each case). At the common point of the 470Ω resistors in each demodulator the demodulated colour signal appears without the reference oscillation because the reference oscillation is balanced out in the bridge network. As can be seen the *R-Y* demodulator is preceded by a transformerless PAL reversing switch which operates on the chrominance signal from the delay line, not on the reference oscillation.

It has become customary with transistor colour receiver circuits to use *RGB* drive to the c.r.t. cathodes, as this requires lower

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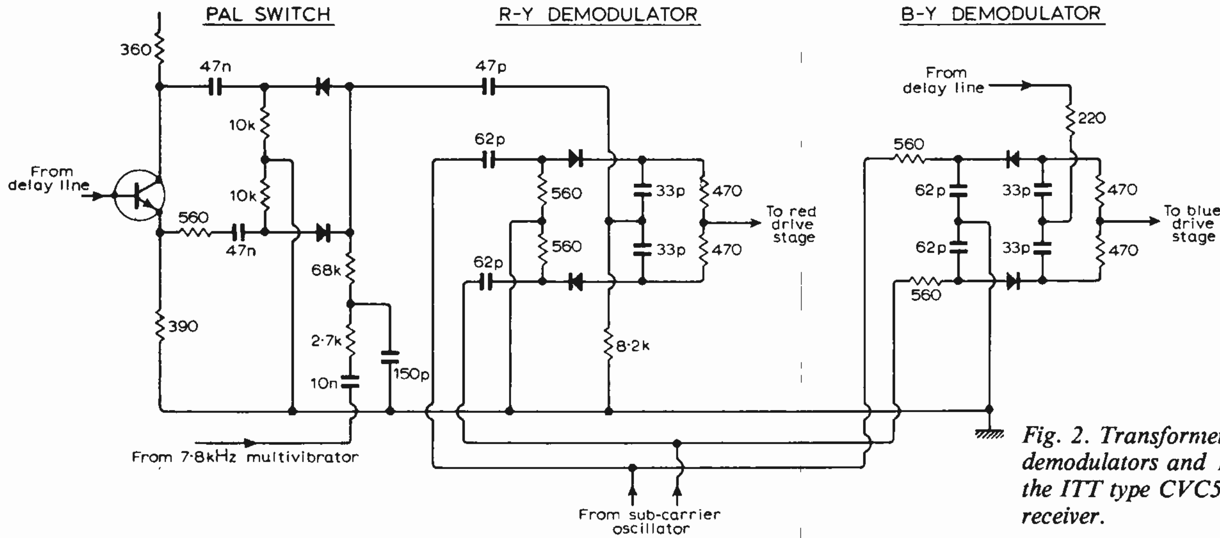


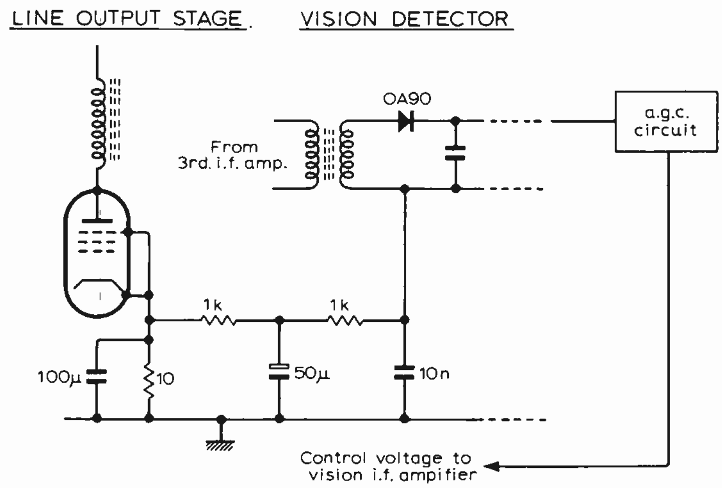
Fig. 2. Transformerless synchronous demodulators and PAL switch used in the ITT type CVC5 colour television receiver.

voltage swings than do colour-difference signals, which are matrixed with the luminance signal by the c.r.t. In the new Philips set referred to above, for example, matrixing is carried out in the decoder i.c. (4) at low level and the *RGB* outputs receive several stages of amplification through separate channels before being applied to the c.r.t. cathodes. This method relies on accurate setting of the grey scale for good monochrome reception and for this reason the colour drive adjustments are preset at the factory.

Colour television sets usually include a circuit for preventing the beam current of the colour c.r.t. from rising above a certain upper limit, the purpose being to save the tube's shadow mask from becoming overheated and consequently distorted in shape. In any case the shadow mask carries the major part of the beam current, and with certain types of picture content the electrical energy converted into heat could become excessive. Normally the beam current limiting is arranged as a form of automatic brightness control, but during operation this causes an error in colour saturation and also a loss of detail in the darker parts of the picture. To avoid this undesirable side effect, Decca, in their latest colour receiver chassis, have designed the beam limiter as a form of automatic contrast control. Fig. 3 shows how it is arranged. A potential proportional to beam current is derived from the cathode of the line output valve and this is applied through the vision detector and other circuitry (not shown) to a video emitter follower and then to the a.g.c. circuit. The direct voltage from the a.g.c. circuit is fed back to bias the first vision i.f. amplifier transistor and so control i.f. gain in the normal manner, but is further controlled by the beam limiting circuit. Thus if the current through the line output valve starts to become excessive the effect is to reduce the amplitude of the i.f. signal and, consequently, reduce the drive to the c.r.t. and hence the beam current.

Numerous small u.h.f. television receivers from Japan were shown at the various trade shows we visited. Three new models from Standard included an all-transistor 3in portable, the SRV 307, employing 29

Fig. 3. System used for c.r.t. beam current limiting in the latest Decca colour receiver.



transistors and 23 diodes. This works from an internal supply of nine U11 cells or from a car battery. A mains adaptor is available. The weight, with batteries, is 2.4kg and the price is £77 4s 6d. The other two sets were the same price and were 7in and 12in mains/battery portables. Sanyo showed a 5in portable set, the 5TC1 weighing only 4kg with built in cadmium/nickel cells (12.5V). Using the batteries 16 hours of reception is possible: charging takes 4 hours. This set uses 25 transistors and 22 diodes, and costs £69 15s including a mains adaptor. The Toshiba 11 TBB is an 11in receiver selling at £79 10s. All these models gave clear stable pictures in the London service area with modest loop or telescopic aerials. The continuous tuning employed on these sets was easy but inter-station noise was considerable.

Teleton (European distributors for the 26-company Mitsubishi combine) still haven't produced the 12in colour portable that was rumoured in last year's survey. **Other circuit features noted:** The B & O Beovision 3200 colour set has a circuit which automatically cuts out the colour sub-carrier notch filter in the luminance channel when the set is receiving monochrome transmissions. This colour set also has the unusual features of a separate high-frequency loudspeaker, and bass and treble controls. In the ITT CVC5 colour receiver the saturation and contrast controls are mechanically ganged, and there are separate diode detectors for vision and intercarrier sound. The e.h.t. pulse winding on the line output transformer of the latest Decca colour chassis is underneath the primary winding instead of on top—a cost saving feature. Automatic line hold, dispensing with the usual manual control, is employed in the Philips and ITT colour receivers.

Sound receivers and reproducers

The mere appearance of integrated circuits in receivers is no longer a matter for surprise, but as the art of using i.cs progresses makers may find alternative or novel uses. A case in point is the use of the TAD100 circuit in the new Dynatron amplifier and a.m./f.m. tuner (HFC91), announced earlier this year. This i.c. is intended as an i.f. amplifier, detector, local oscillator and a.f. preamplifier, but in the Dynatron circuit the stage normally used for audio amplification is used instead as a d.c. amplifier feeding an a.m. moving-coil tuning meter. (Same meter acts as tuning indicator on f.m. straight from the discriminator.)

This tuner incorporates diode tuning on v.h.f. for four circuits, two back-to-back diodes each, with three pre-set potentiometers and another linked to the tuning scale. The scale covers 88-100MHz, giving greater spread than with 108MHz top

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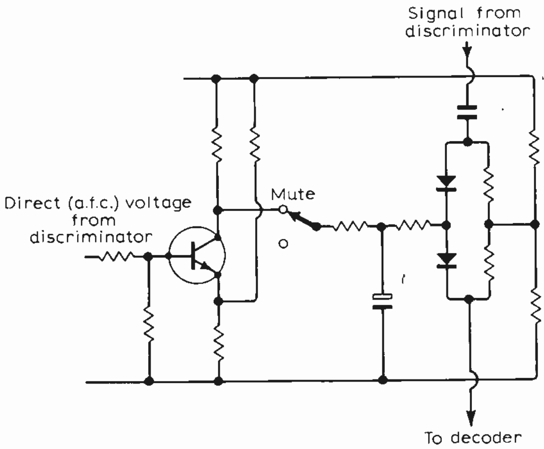

Fig. 4 (left). Interstation muting circuit in Dynatron tuner-amplifier. Tuner uses diode tuning, five integrated circuits and dual-gate f.e.ts.

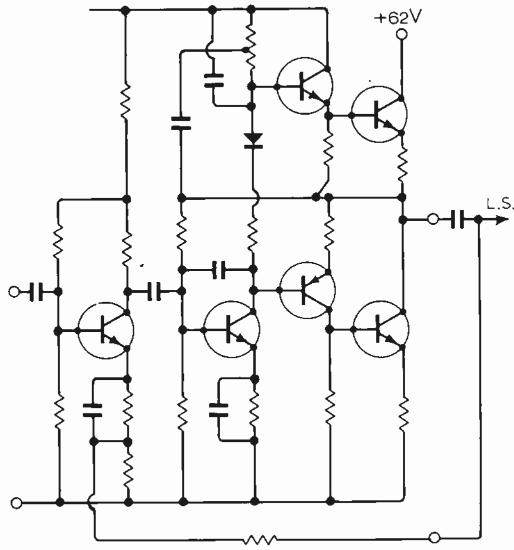

Fig. 5 (right). Hybrid thick-film high-power amplifier module made by Bendix is used in new Kellar amplifiers. Circuit shown gives r.m.s. output of 50 watts.

frequency, but the pre-set potentiometers extend to 108MHz, in anticipation of this model being sold in the U.S.A. and Canada. Supply voltage for the tuning circuits is regulated with a TAA550 i.c. between supply rail and ground. The three pre-set controls together with the displayed range allows a local station to be pre-set as well as the three national services, but we have seen some sets with four pre-set controls. (Incidentally, some local stations will be using 45° slant polarization with consequent benefits to car and 'picnic' radio reception—see p. 492.)

Interstation muting is achieved by switching back-to-back diodes in the signal path to the decoder from a d.c. amplifier following the discriminator (Fig. 4). Other features of this design are dual-gate f.e.ts in the f.m. tuner, four-i.c. Görler f.m.-i.f. amplifier, ceramic filter unit with the TAD100 a.m. i.f. amplifier, twin-T high- and low-pass filters and negative feedback tone controls in the pre-amplifier. Extensive use is made of BC148 transistors (11) with BC149s as low-noise input transistors. The audio amplifier is a conventional quasi-complementary circuit giving 45 + 45 watts into 3 ohms, or reduced powers into higher impedance loads, at low distortion (0.2%). It is typical of modern designs using silicon transistors which allow a high voltage supply with consequent high-power output. A thyristor overload protection switch reduces output to a safe level when safe operating limits of the transistors are exceeded, lighting an indicator lamp and reducing the supply voltage to the pre-amp emitter follower. Price of this tuner amplifier is £165.

A notable trend in this kind of quality tuner is the omission of a ferrite aerial for a.m. reception—or at least provision of a separate coil pack for an exterior aerial. This was also noted on the new B & O Beomaster 1600 which, apart from the addition of f.e.ts and an f.m.-i.f. ceramic filter, is a re-styled version of the 1400. This new a.m./f.m. model, with a power output of 15 + 15 watts, and including two short-wave bands, costs £122. Another new a.m./f.m. tuner-amplifier from B & O is the Beomaster 1200. A new design, but along fairly conventional lines, it uses a cascode f.e.t. input circuit and diode tuning. Sensitivity is better than that of the 1600 (1μV for 26dB s/n ratio f.m., 7μV for 3dB s/n ratio a.m.) and power output is 15 + 15 watts. Price £108.

Latest Teleton tuner-amplifiers (made by Mitsubishi) acknowledge the difficulty in seeing vertical tuning scales in 'low-line' models by using a horizontal scale—in one case both types are used.

The Dynatron stereo decoder is fairly typical of current practice. Designers seem to have settled on the balanced switching demodulator, this only partially demodulating the *L-R* signal of course. Correction to give 40dB separation is made in a common-mode amplifier (one transistor in each channel with emitters bridged with a resistor). The Dynatron circuit has the added luxury of twin-T filters in each channel to reduce sub-carrier level.

First commercial amplifiers using a hybrid thick-film integrated circuit for the power amplifier are designed by Stanley Kelly, joint managing director (with Sidney Larholt) of Kellar Electronics Ltd (Maryland Works, 9 Brydges Road, London, E15 1NA). One of the series of amplifiers is used in the new Kellar cassette tape recorders using the Dolby 'B' system for noise reduction above 2kHz. The system gives a signal-to-noise ratio improvement of 10-15dB (see p. 519). As well as using this system with the 25 + 25 watt cassette tape recorder, it will be available separately for reel-to-reel tape recorders (to be shown at the Audio Fair). We expect the Dolby 'B' system to appear in equipment from various makers later this year—Decca, Rank and Metrosound are rumoured to be working on this.

The hybrid thick-film module used by Kellar—a Bendix circuit originally designed for industrial applications—has made possible a low-cost 15 + 15 watt amplifier. Marketed under the name Nova by L. L. Electronics at £33 and with low distortion (about 0.2%) this could be the best value for money in terms of watts per £ at this power level. The circuit uses a conventional quasi-complementary output stage and complementary driver with discrete transistors on the thick-film circuit. Module measures 20 × 45 × 80mm. A slightly larger 50-watt module is available, circuit is shown in Fig. 5, and will be used in an amplifier to be released shortly.

In designing the regulated power supplies for this amplifier, Stanley Kelly found that by using two series transistors (OC22s) each feeding one amplifier, instead of operating them in parallel, and arranging feedback from one side of a long-tailed pair circuit, he obtained a worthwhile increase in channel separation.

Another Nova Design, the 10 + 10 watt amplifier, currently using an AD161/2 output pair, is due for re-design around a Siemens TAA420 integrated circuit.

Another hybrid high-power amplifier is made by RCA (type TA7625) which operates from a split power supply and has a built-in limiting circuit.

Among the new tape recorders there is a profusion of cassette machines and medium quality reel-to-reel types. Three models from Sharp are of particular interest, having servo control of motor speed: a cassette recorder with slider controls for setting the recording level (model RD423, price £57 10s) and two 5in reel-to-reel recorders with switched speed change (model RD513, price £39, and its push-button-control counterpart, the RD514, price £42). In each of these recorders the d.c. drive motor is in the collector circuit of a control transistor, the base circuit of which is fed from a generator coupled to the motor shaft. Speed change from 3¾ to 1⅞ i.p.s. is accomplished by a switched resistor change in the bias circuit.

Sanyo now have thirteen cassette recorder models available ranging in price from £19 15s to £99 15s—some of these have an f.m. radio too.

Along with several new tape recorders, ranging in price from £55 15s 7d to £157 14s, Grundig displayed the TK3200 three-speed battery recorder with a stick microphone priced at £178 3s. At 7½ i.p.s. the frequency response is said to be 40Hz to 16kHz and wow and flutter as 0.1%. Signal-to-noise ratio is claimed to be better than 48dB at 1⅞ i.p.s. Outputs are 500mV (into 15kΩ) with provision for driving an external 4Ω speaker with 2W or the internal speaker with 0.8W. This is a single channel recorder.

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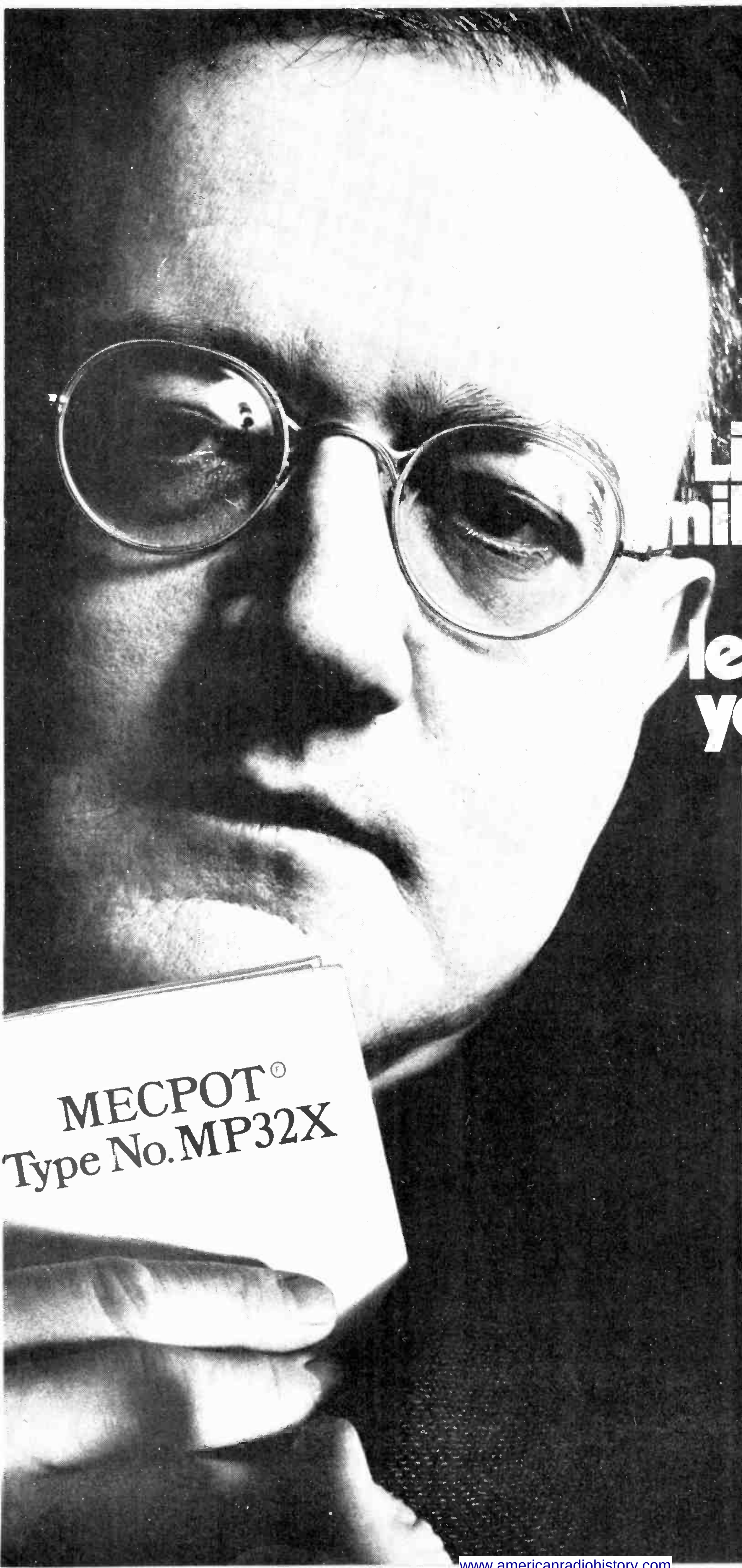

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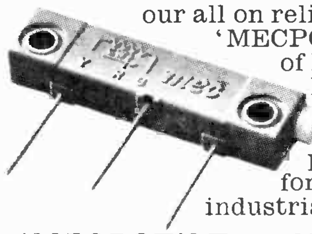

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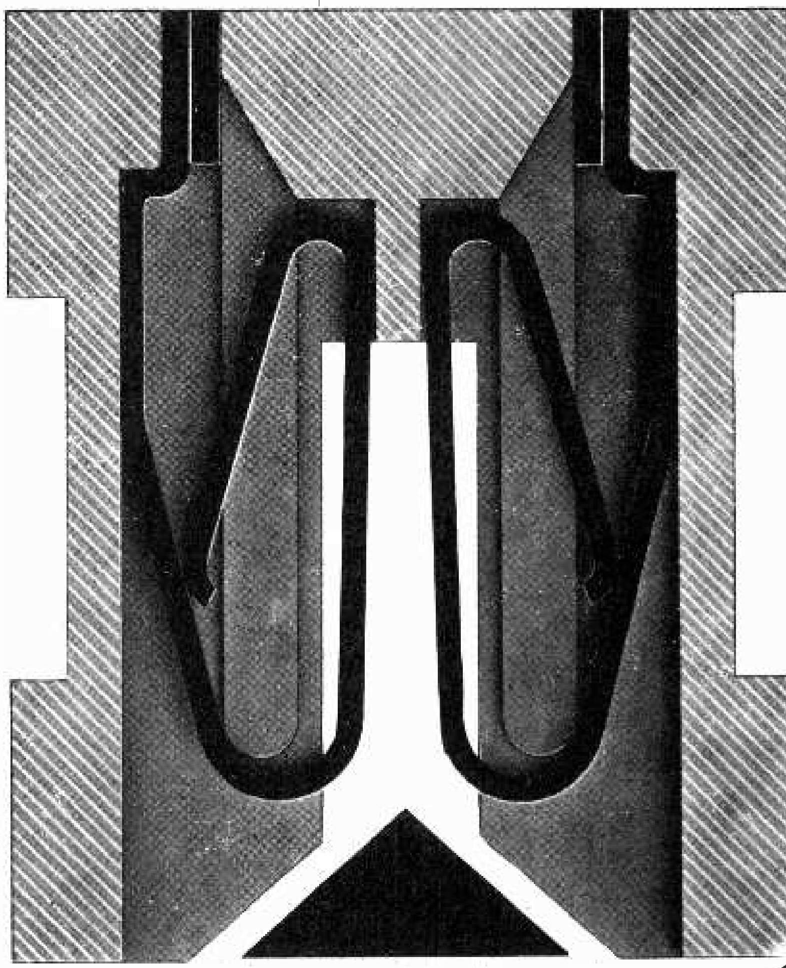

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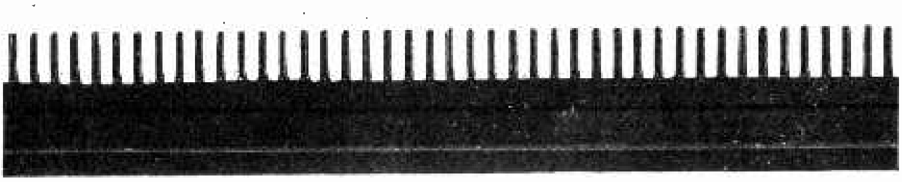

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Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Mobile radio and amateur bands

I refer to the letter under the above heading which appeared in your September issue.

The second paragraph implies that some unofficial arrangement exists concerning the 70-MHz amateur band. The Electronic Engineering Association seems unaware that the amateur service has been allocated frequencies between 70 and 71 MHz since 1956. Indeed, recent issues of your journal have carried news of propagation experiments by amateurs using these frequencies. It is implicit by the terms of the Radio Regulations that where the amateur service is the secondary user then this is subject to non-interference with the primary service.

The last sentence of the third paragraph of your correspondent's letter shows a lack of knowledge of the occupancy of the 420-450 MHz band. There are, of course, a number of amateur television stations using frequencies in this band, but, in addition, there are a far greater number of stations occupying the band for experimental work in connection with moonbounce, meteor scatter and satellite communication. Stations operating under the terms of the Amateur (Sound) Licence B are present in large numbers on both the 2-m and 70-cm bands.

The final sentence of the E.E.A. letter mentions "amateur associations". I would point out that the Radio Society of Great Britain is the organization recognized by the Ministry of Posts and Telecommunications as representing the stations of the amateur service in the United Kingdom.

There are regular discussions between the Ministry and the R.S.G.B. concerning licensing matters and frequency allocations and we have no doubt that at the correct time the Society will be approached by the licensing authority.

J. A. SAXTON,
President,
Radio Society of Great Britain.

Crisis in microelectronics

Your September editorial draws attention to the difficulties of the integrated circuit industry, but the conclusion one should draw depends rather on one's standpoint. The U.S. microelectronics industry obviously has excess production capacity for the foreseeable future, and the present marketing situation cannot be changed rapidly. Dr. F. E. Jones cites import duties as a reason for British companies not manufacturing in the Far East. From a quick enquiry I get the impression that duty on radio components entering Britain is at most 17% which is not decisive in comparison with price differentials quoted in your Editorial, and nil from Commonwealth territories. So why not use Hong Kong? In addition, I feel that more could be done to establish the facts about American manufacturing costs and selling prices, unless it is the desire of the British i.c. industry to befog the issue. If American i.cs which are packaged in the Far East are taken back into America on conditions which assume that the major part of the manufacture has taken place in America, then surely there is no doubt about the country of origin. But even if there is doubt about the country of origin in the manufacturing sense, it must be possible to establish the country in which the manufacturing company has its registered office. Before we worry too much about the difficulty of ascertaining the true country of origin, let us know at least the selling price in U.S.A.

If there is still a possibility of Britain joining the Common Market there might be a case for a European production facility. So far as concerns Britain as an isolated unit, however, the sensible thing might be to cut one's losses and abandon the whole of the British microelectronics industry as now understood. This sounds a rather staggering proposition, but surely the financial loss to be cut would be no greater than losses which have been cut on unsuccessful military development projects. When I say "the industry as now understood" I mean an industry producing logic units and simple linear amplifiers. These are the type of thing which can be bought cheaply from U.S.A., and so far as military strategy is concerned there could well be either a factory operating under licence or an assembly and encapsulation plant in Britain. But I doubt however whether this is a serious concern: one plane load of integrated circuits could be enough to last for some time—longer than a nuclear war!

On the other hand, I think that large scale integration should be pursued in this country because by their nature large-scale-integrated-circuit chips are usually tailor-made for a particular purpose. It is no novelty for an industry to cut down on the manufacture of a basic product in order to progress to more advanced products.

Another point is that we must not be mesmerized by the idea of integrated circuits using present techniques, which are essentially collections of junction semiconductor devices. There may be a big future for solid state devices which do not depend on junctions, and Britain might have the chance to take the lead in this field. The Gunn effect has already been put to extensive use. The DOFIC appears to have faded into obscurity after a brief appearance, but perhaps there may yet be possibilities of developing it further. There is also the acousto-electric amplifier based on the interaction of electrons and phonons in materials such as cadmium sulphide. These may well play a large part in the future of electronics and should not be overlooked in a moment of panic about the marketing of devices which although scientific marvels to the layman can now be manufactured as a matter of routine.

D. A. BELL,
The University of Hull.

Class AB amplifiers

I am grateful to Mr. Mitchell for his letter in the September issue concerning my class AB amplifier, but there are some points which he makes which, I feel, should not pass without challenge.

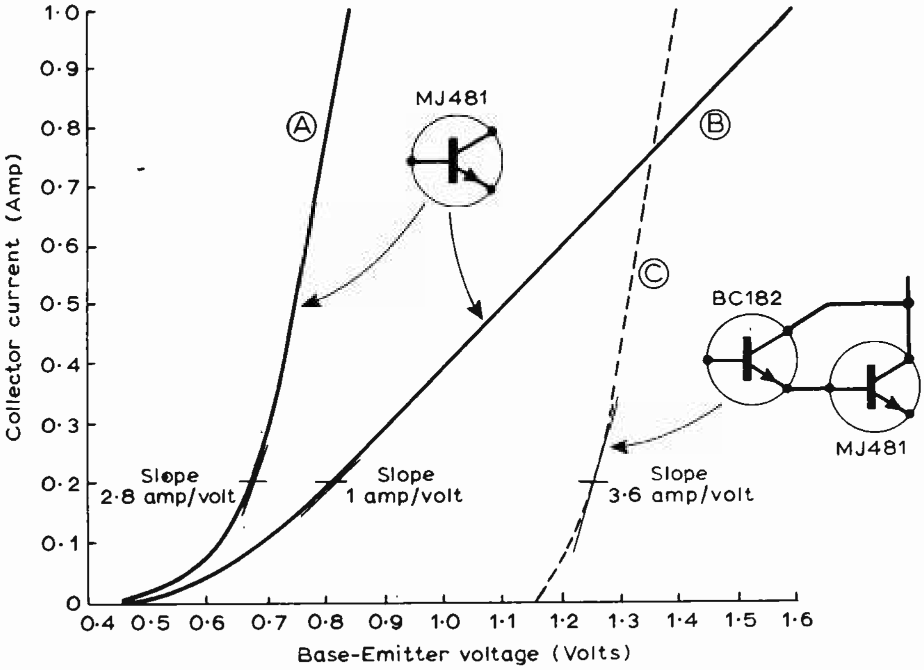
In particular he states that a Darlington pair output stage has a lower mutual conductance than the output transistor on its own. While, in theory, this could follow from the fact that the second transistor imposes an impedance in the emitter circuit of the first, this situation does not arise under any but near zero source impedance systems, as I have illustrated in the transfer characteristic graphs on the next page.

Curve A is the transfer characteristic of a simple (MJ481) transistor with a source (base input circuit) resistance of 10 ohms. Curve B shows the performance of the arrangement but with a source resistance of 100 ohms. Curve C is that of the same output transistor, but with an input Darlington configuration using a BC182 input transistor. There is no measurable difference in performance, in this configuration, with source resistances of 10, 100 or 1,000 ohms.

In the event, the slope of the Darlington pair, at 200 mA, which was my chosen quiescent current, is 3.6 amps/volt as compared with 2.8 amps/volt for the simple output transistor.

The presence of as little as 100 ohms input circuit resistance reduces this to 1 A/V, which confirms the point I made in my article, which was concerned, implicitly, with the circumstances which would exist in a practical design.

The second point on which I differ from Mr. Mitchell concerns the conditions of operation of a class A stage. I believe this classification should be restricted to systems in which each component of the output stage operates in its linear region over the



whole of its effective output swing. The mere fact that one or other of the output transistors is not completely cut off is not enough to satisfy this requirement.

Although I had not mentioned this point specifically in the article, the use of the amplifier in true class A does bring about a reduction in the distortion typically to below some 0.01%, at power levels below 15 watts, over the frequency range 100 Hz-5kHz, and the distortion content then decreases linearly with reduction in output signal magnitude.

My decision, in the design of the amplifier, to employ a variable resistor, as a source of bias, between the bases of the output transistors, rather than a more complex temperature compensation network was based partly on the convenience of adjustment of such a biasing system, as compared with, say, a string of diodes (two forward biased silicon diodes will, in fact, give almost the correct quiescent current, and this arrangement was used in some of the prototypes in use by friends) and partly on its lesser proneness to catastrophic failure than transistor "amplified diode" systems.

My curve B indicates the relative insensitivity of the single transistor output stage to variations in forward bias (and the choice of 200 mA quiescent current very much reduces thermal effects, even with an 8-ohm load!) as well as the excellent transfer linearity of such a system which contributes to the lower harmonic distortion figures obtainable with such an output stage in comparison with the more normal push-pull configurations.

Both Mr. Mitchell and Mr. Gibbs (letters, Aug. 1970) have taken me to task for my observation in the article that "the use of a complementary pair of emitter followers 'driven from a low source impedance' appeared to offer the best way of minimizing the several problems" described in the introduction.

The article in question was in fact written as one, rather lengthy, article which was divided in two for convenience of publication, and this division, coupled with some editorial deletions, resulted in the observation above being given an unexpected degree of prominence. Since I was, at this stage, reviewing the thought processes which had led to the choice of this output stage configuration, it would have been better if I had continued "and this type of stage was therefore chosen as the starting point for this design".

In the event, both the preliminary calculations and the initial experiments indicated that it was neither practicable nor desirable, from the point of view of linearity of operation, that the output stage should have a low source impedance and the solution suggested by Mr. Mitchell in his letter, that of a relatively high driver impedance with a low inter-base impedance, was the configuration which had been adopted in the final design.

In reply to the letter from Messrs Smith and Walker in the September issue I would point out that the total harmonic distortion was quoted at 1000 Hz, because this is the recommendation of the B.S. and DIN specifications. The t.h.d. figures, at full output, at 100Hz and 10kHz, are typically 0.04% and 0.06% respectively. At low frequencies the harmonic distortion is mainly influenced by the impedances of the power supply bypass capacitor and the decoupling and 'bootstrap' capacitors, and an improvement can be made, if necessary, by increasing the value of these.

At high frequencies, the distortion content is mainly determined by the deliberate and necessary reduction in the open-loop gain, and feedback factor, required to maintain good reactive load stability, although the circuit layout and stray capacitances have some effect.

I apologise for the omission of the bandwidth limits for the noise figure measurements. These were effectively those imposed by the amplifier gain/frequency characteristics, as would be measured by a very wide bandwidth millivoltmeter. The use of a more restricted bandwidth, say 20Hz—20kHz, would allow an apparent improvement in the specified noise figure. (It is, in fact, quite inaudible.) However, on looking through back numbers of *Wireless World* I find that other authors have been equally remiss in omitting measurement bandwidths when quoting noise levels. This point will, perhaps, be noted in the future.

I regret that the measurement parameter "square wave transfer distortion" was not accompanied by some further explanation. In practise, transfer distortion is measured by comparing electrically the waveforms at the input and output of the system under test, and then expressing the error arising in the transfer as a percentage of the input waveform, as measured on an r.m.s. calibrated voltmeter such as that used for conventional t.h.d. measurements. Any convenient waveform may be used for this purpose.

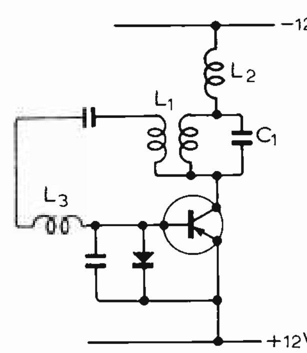
Typical values for transfer distortion with conventional audio amplifier designs using a 10kHz square wave and a resistive load range from 0.2% to 10%. Square-wave transfer errors as high as 30% are fairly common under reactive load conditions, and this, in conjuction with the relatively high distortion levels sometimes found at low volume levels, may account for much of the so-called 'transistor sound'. Unlike harmonic distortion, transfer distortion with reactive loads may worsen as the amount of negative feedback is increased.

J. L. LINSLEY HOOD,
Taunton,
Somerset.

Sine-wave power oscillator

The basic requirements for any circuit are that it should be comprehensible, designable and as simple as possible. The old rule, simplificate and add lightness, still applies. Mr. Armer's circuit (August p.402) and his explanation, do not seem to satisfy this.

It is rather easier to examine this circuit if the earth point is moved: indeed it is probably easier to use it if a second variation is adopted. Fig. A shows the circuit rearranged for conventional analysis, and Fig. B shows it with the collector of the transistor at an a.c. earth. In both circuits the tuning capacitor has been transferred to be in parallel with only the larger part-winding, simply to make the circuit look conventional.

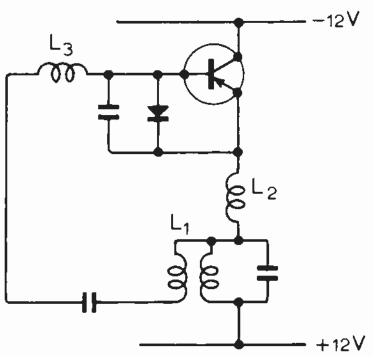


(a)

Before discussing how it works, let us look at the numbers. L_1 is tapped to give

400μH in the one part, 200μH in the other. The total inductance is thus 200 $\times$ $(1 + \sqrt{2})^2 \mu$H, or 1.16mH. This tunes with $C_1 = 0.05\mu$F to 21kHz. It would seem that the frequency is not determined, as Mr. Armer suggests, by L_1 and L_2 in parallel. They are, in fact, in series, to give the reactance plot shown in Fig. C.

The feedback path when neither the diode nor the base emitter junction are conducting very much is a low-pass half-section with a cut-off frequency of below 1kHz and an impedance of about 75Ω. When either diode is conducting it reduces to an induc-

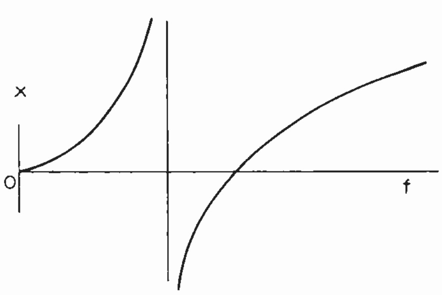


(b)

tance of about 180Ω, to give a 90° phase shift.

The load for 13W must be in the region of 10Ω. The reactance of C_1 is about 160Ω, so that the Q of the circuit is very small indeed. Furthermore the value of $I_{dc}(L)^{\frac{1}{2}}$ for the 400μH section would seem to require a ferrite core of some 30cm^3 volume.

In a push-pull circuit L_2 would be identified as a constant current supply choke. Here its value is just low enough to avoid really awkward high voltages as the transistor is cut off, provided that the cut-off is



(c)

slow, and thus lossy. What Mr. Armer means by the statement that the base current controls the collector current while the transistor is saturated I cannot understand. The only way this can be done is by limiting emitter current and drawing much of it off into the base. But 2A base current?

Mr. Unsworth ("1000:1 Attenuator"—same page), in offering us an output of 0.768V, appears to neglect the fact that the 8 should be significant. He obtains his 0.7-0.8V across an R which is loaded by roughly 100R. Thus if we look at the first step, nominally 0−0.1 we shall have, in fact, 0−0.099. Moreover, V_{out} must be loaded by something like 5000R—I have not worked it out—if the system is to be plausible.

T. RODDAM,
London.

R.S.G.B. Exhibition

Where have all the experimenters gone? Seduced by sales talk every one!

Not so long ago under the heading of "Components, Complaints and Complacency" we urged manufacturers to turn a sympathetic ear to the component needs of the home constructor. It would seem, however, that as far as the amateur radio enthusiast is concerned we should not have wasted the ink and the paper. There were only seven entries from the whole amateur radio fraternity for the home constructed equipment competition held at the recent R.S.G.B. Exhibition. A lamentable turnout.

Is it true that the ham of today does no more than gossip on the air while twiddling the spun aluminium knobs and peering at the gleaming Perspex dials of a shiny piece of commercially constructed equipment? If so the frequency bands allocated to amateurs are being wasted and should be re-allocated to a more deserving cause.

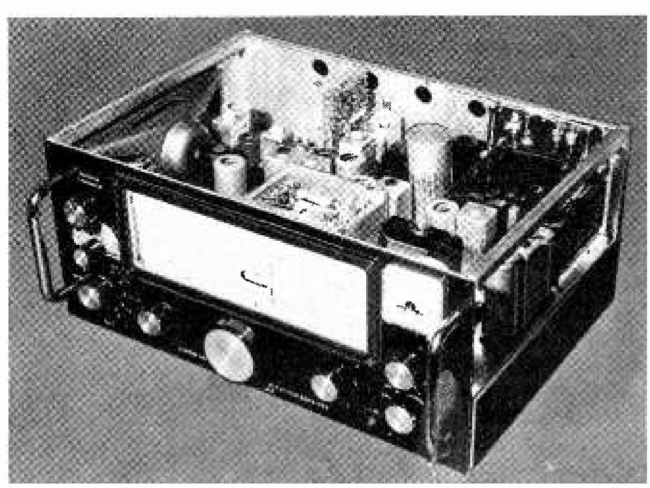
Today's home constructor has never had it so good. He can obtain at reasonable prices sophisticated components which were not even available to industry a few years ago. Is it that today's amateur does not want to risk his money on experiment? Or are the days of trial and error design over?

The winner of the Horace Freeman award for home constructed equipment this year was George Goldsmith, from Jersey, with a finely engineered communications receiver for the amateur bands. It employed both valves and semiconductors and incorporated three crystal filters. Sensitivity was 0.4μV for a 10dB signal-to-noise ratio. Other prizewinners were M. F. Taylor, of Reading, with a 2-m transceiver and J. R. Jessop, of Pinner, with a linear amplifier. This year the John Rouse Trophy was not awarded as there was not a single entry from the under-16 age group, a necessary qualification for this award.

The exhibition itself is still affectionately known as the R.S.G.B. Exhibition, rather than by its somewhat pretentious title (International Radio Engineering and Communications Exhibition). This year it sprouted the additional title Radiocom 70 as well. Attendance was just over 7000; about 800 less than last year. This may have been due to the show being held in August instead of later in the year as in the past. We do not question the value of the exhibition which, as a meeting place for the electronics-radio fraternity, is almost second to none. It is patronized by all types of people from home constructors and amateurs to engineers and academics. It is a pity that working equipment demonstrations were so thin on the ground this year. There were many bargains in surplus equipment to be found even though one exhibitor admitted that the exhibition gave an opportunity to get rid of a "pile of junk".

The Ministry of Posts and Telecommunications had a display which illustrated how interference can spread in large urban areas and the problems incurred in tracing, measuring and suppressing it. This year the Radio and Space Research Station had a stand which showed the work they are doing in the field of propagation of radio waves over a very wide band of frequencies.

The communications receiver built by George Goldsmith which this year won the Horace Freeman Trophy. It has a hybrid circuit and operates on the amateur bands.



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News of the Month

Ceramic stores for information displays

Some interesting work currently being carried out at Bell Labs in America has resulted in a sort of re-usable slide for projection purposes. Images stored on the slide can be erased and replaced with a new image at will. At the present stage of development the device, or 'ferpic' as Bell Labs call it, would only be suitable for use in displays of written text and figures which do not require to be quickly altered. Not a great deal of information has been released but it does not take a great deal of thought to see that if the speed can be increased and if the devices can be manufactured at reasonable cost the ferpic could have a bright future.

The material used in the construction of the ferpics was first announced by an American concern called the Sandia Corporation. The device is a sandwich. A very thin plate of fine-grained ferroelectric ceramic is coated with a photoconductive film which in turn has a system of transparent electrodes deposited on it. The whole is then bonded to a flexible transparent substrate.

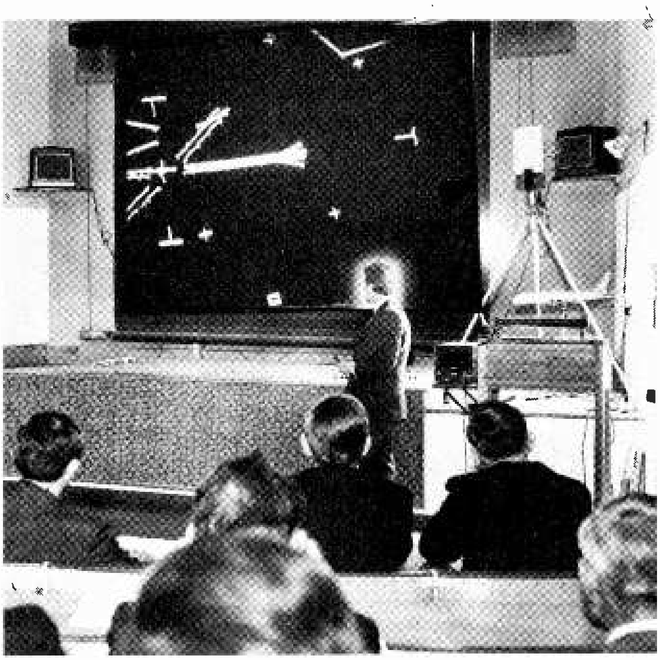
The release from America which describes the operation of the device is rather sketchy and it is not clear exactly how the device works. Quoting directly from the American statement; "To change the stored information in this simple structure, a new technique, called 'strain-biasing' was developed by Bell Lab scientists. The ferpic sandwich is flexed so as to stretch or strain the material.

"In an ordinary film slide an image is stored in the form of variations of transparency across the film. In a ferpic the image is stored as a variation of the 'Birefringence' of the ceramic plate, i.e., as a variation in the way the plate transmits polarized light.

"In one mode of operation, a scanned laser beam records an image on the photoconductive film—one picture element at a time as in a TV picture. A voltage applied to the transparent electrodes develops a field across the ceramic. When the field is removed, the image remains stored on the ceramic plate. The image stored in the ferpic device can be viewed by putting light polarizing sheets over it, or the image can be projected on a screen using polarized light in a conventional projection system.

"To erase the image, the entire structure is flooded with light in the presence of a reversed electric field; the plate is then ready to store another image."

Bell Labs is exploring this device in the hope of obtaining efficient, low-cost solid-state information displays with features that are difficult to obtain in present display systems. Because the image store in the ferpic device can be projected, very large displays can be obtained. Also ferpic slides can retain images for a long time without electrical power, fading only slowly.



Representatives of the Services, Ministry of Defence and industry watching a demonstration of the Eidophor large-screen display system at the College of Air Traffic Control, Hurn Airport. This equipment was used by N.A.S.A. during Apollo 11 and 12 missions, and its pictures were re-transmitted by practically every TV network throughout the world.

Slant polarization for local radio

Certain B.B.C. local radio stations will be using slant (45°) polarization instead of horizontal polarization. This follows successful tests by B.B.C. research department engineers in Kingswood and Nottingham areas†. Reception in cars and in the open air with vertical aerials will be improved through the effective increase in transmitter power equivalent to about 6—9dB, at least doubling signal strength on average. Hitherto, satisfactory reception with vertical aerials has been largely a matter of luck. As a consequence, horizontal roof aerials will give a reduced signal—about 70% or −3dB. If necessary, this could be recovered by slanting the aerial (clockwise by 45° when looking at the transmitter) but the national v.h.f. services, not using slant polarization, would suffer a loss.

Stations using slant polarization are: Manchester—95.1MHz; and early in 1971 Blackburn—96.4MHz; Derby—96.5MHz; Leicester—95.05—95.2MHz and Nottingham—94.8MHz.

B.B.C. make the point that some roof-level aerials may not be correctly oriented for the local station and some adjustment may be needed for best results.

Mullard's golden jubilee

As part of their Golden Jubilee celebrations Mullard Ltd. are to stage a 3-week public exhibition in the Electronics Centre of their London headquarters, Mullard House, Torrington Place, London, WC1E 7HD.

The exhibition will trace the history of electronics—linked with the company's own history—over the past 50 years.

The exhibition opens on October 5th, and will run until October 24th, opening every day (except Sundays) between 10.00 and 18.00 (21.00 on Thursdays). Admission will be free. Later, the exhibition, probably in a modified form, will tour Mullard establishments throughout the country.

The main window of the Electronics Centre will contain examples of vintage electronic equipment contrasted with their modern counterparts and supported by displays of Mullard components used in their manufacture.

Within the Centre one of the main attractions will be a radio transmitter built and operated by Mullard 'hams'. Some of the company's earliest valves will be used

†Spencer, J. G. Tests of mixed polarization for v.h.f. sound broadcasting. *B.B.C. Engineering*, July 1970, pp.4-12.

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in its construction. It is also hoped to show part of the original 2LO transmitter used by the B.B.C. for its first public broadcasts.

Another main feature will be entitled 'Mullard Through the Decades': This will show electronic products and their components of each era, beginning with the first Mullard high-power transmitting valve. Displays in the 'Today' section will feature many working exhibits including a microwave oven. Another section will provide a glimpse of some of the likely future applications of electronics. The exhibition will also include a section where visitors can test their skill and reflexes in various electronic games.

I.T.A. completes 405-line network

The coming into service of a local v.h.f. relay station at Newhaven, Sussex, on August 3rd marked the completion of the I.T.A. network of 405-line v.h.f. television transmitting stations. No further additions to this network, which operates solely in Band III, are currently envisaged. All I.T.A. station building plans are now concentrated on the rapid expansion of the new 625-line network of u.h.f. transmitters.

Since the original opening of the first I.T.A. Band III station at Croydon, south-east London, on September 22nd, 1955, the 405-line network has expanded steadily and now comprises 47 transmitting stations, of which 20 are manned, and 27 remotely controlled. The network provides television coverage for approximately 98.7% of the population of the United Kingdom.

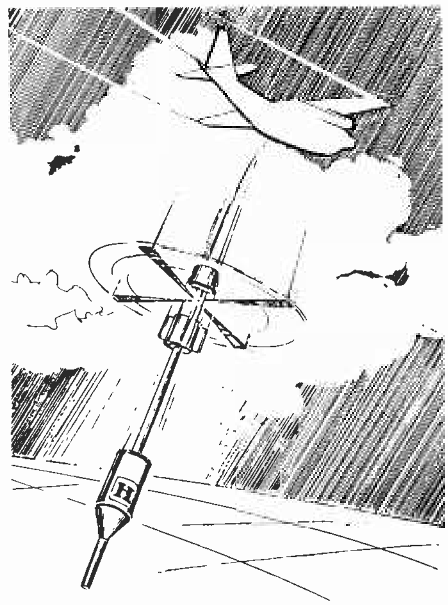
Want to use a satellite?

The National Aeronautics and Space Administration is offering the use of six satellites which have long since fulfilled their original task but are still operational.

The six spacecraft are: OGO-1 (Orbiting Geophysical Observatory), ten experiments still operational, highly elliptical orbit 111,000 by 39,000 miles, inclination 58°; OGO-3 similar orbit to above, 13 experiments operational; Explorer-31 (direct measurements explorer) with five operational experiments and Explorer-33 (Interplanetary monitoring probe) orbit 410,600 by 109,000 miles, orbit period 32 days; finally the orbiting solar observatory satellites OSO-3 and OSO-4 are also available.

Interested scientific bodies should contact Code SG N.A.S.A. Washington D.C. 20546.

As a matter of interest the oldest satellite still transmitting useful information is the Canadian Alouette-1 which was launched on September 29th 1962.



The sketch shows a device called EROWS (Expendable Remote Operating Weather Station). The eight-foot arrow-like structure implants itself in the ground after being dropped from an aircraft. The device is turned on remotely and it begins to transmit information on wind speed and direction, atmospheric pressure, temperature, humidity, precipitation and cloud cover. EROWS is being developed by Honeywell for the U.S. Air Force.

Apollo colour TV camera

Under a $196,500 contract R.C.A. will deliver two colour television cameras to the National Aeronautics and Space Administration which can be used under the extreme lighting conditions found on the moon. The cameras, which weigh ten pounds each, not only operate in dim lighting conditions but they can be pointed directly at the sun without sustaining any damage!

The cameras employ silicon intensifier tubes, the imaging surface being made up of almost 400,000 silicon diodes. The colour system (525 lines per frame and 60 fields per second) is as used in earlier Apollo missions with a colour filter wheel arranged so that monochrome pictures corresponding to the red, blue and green brightness levels are transmitted sequentially for recombination at the ground station.

A lamp, which simulates the brightness level of the sun, being pointed directly into the lens of the new Apollo TV camera.

European component standardization

A 'Harmonized System for Electronic Components' is being introduced in Western Europe in which a set of common specifications and quality assurances will remove all technical barriers to inter-country trading. The system is similar to, and fully compatible with, our own BS 9000.

Agreement has already been reached on all the major aspects of the Harmonized System and at the invitation of the Tripartite Committee for Standardization, the Comité Européen de Coordination des Normes Electrotechniques (CENEL), which comprises all E.F.T.A. and E.E.C. countries, has accepted overall responsibility for launching the system in these countries.

The responsibility for the quality assurance aspects of the Harmonized System is to be handled by an independent committee known as the Electronic Components Quality Assurance Committee (ECQAC). It is hoped that the system can be introduced on a world-wide basis later.

Radar base at Lahr

On August 18th Marconi handed over a radar base to Brigadier General R. E. Mooney who was acting on behalf of the No. 1 Canadian Air Group. The

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station, at Lahr in West Germany, is situated on a rather difficult site being ringed with mountainous country and suffering a high annual rainfall. The base will provide air traffic control terminal services for the Canadian bases at Lahr and Soellingen. In emergency the base will provide the German airfield at Bremgarten with assistance. Air traffic problems in the area are complicated by the civil airports at Strasbourg and Basle which are not very far away.

The radar installation, which cost £425,000, consists of a type S654 surveillance radar together with a temporary aerial which will be replaced by a dual beam aerial next year, a secondary radar system and a central operations control room. Under the contract the Canadian Marconi Company, acting as sub-contractors, modified the existing precision approach radar for remote operation from the central control room.

Announcements

"Compendium of Degree Courses 1970", providing information about more than 279 sandwich, full-time and part-time courses, is available from the Council for National Academic Awards, 3 Devonshire Street, London W1N 2BA.

The following courses are to be held at **Hendon College of Technology**, London N.W.4. "Thyristor applications", six lectures commencing 15th October; "Electronics for non-electrical engineers", fourteen lectures commencing 13th October; "Construction and operation of digital computers", sixteen lectures commencing 15th October. Fees are £3, £6 and £8 respectively.

Two-day and three-day **colour television training courses** are being conducted at the Thames Hotel, Bridge Street, East Molesey, Surrey, by Electronic & Colour Television Training Ltd, of 180 St. Johns Road, Woking, Surrey. Fees 15gn and 20gn.

Coincident with next year's Electronic Components Show at Olympia (May 18-21) an **Electronic Components Conference** is to be held at the Royal Garden Hotel, Kensington, under the auspices of the Electronic Components Board.

Truvox Ltd, which became part of the Racal group with the acquisition of Controls & Communications Ltd, has ceased manufacture and their range of audio products will not be available in future. The service department will continue to operate from Hythe, Southampton, providing an advisory service, repairs and a supply of spare parts.

Philips and Brown Boveri.—An agreement has been reached between Philips' Telecommunicatie Industrie, of Hilversum, Holland, and Brown, Boveri & Company Ltd, of Baden, Switzerland, by which the two companies will co-operate in the manufacture and development of sound broadcast transmitters.

REMO is the brand name of a new company, Rectifier Modules International Ltd, of Remo House, Rye Street, Bishops Stortford, Herts. The range of products will begin with three basic types: e.h.t. rectifiers and modules, encapsulated rectifiers and rectifiers to meet British Post Office specifications.

Mobile Radio Communication

Congestion of bands: Proposals for improving the service

The number of private mobile radio licences in the U.K. which at present stands at about 85,000, is likely to double every five years according to figures given at a recent conference on mobile communications arranged by the Society of Electronic & Radio Technicians. Sectors of the frequency spectrum allocated to this mode of communication are already saturated and equipment manufacturers and users are desperately seeking means whereby they can increase the number of available channels.

Speaking before 130 delegates at the conference at Brunel University, Uxbridge, J. R. Brinkley, of S.T.C. suggested that one way out of the dilemma would be to extend the present u.h.f. mobile band (450-470 MHz) to 420-512 MHz. This would entail, at the top end of the extended band, sharing television channel 21 in Band IV. Channel 21 has been assigned to 15 B.B.C. TV transmitters including three high-power stations at Rowridge (I.o.W.) Divis (N. Ireland) and Sandy Heath. D. B. Balchin, from the Ministry of Posts and Telecommunications, pointed out that the TV band was fixed by international agreement. There would be great difficulty in sharing this band with mobile radio. Undeterred Mr. Brinkley asserted that frequency sharing with TV broadcasts had been successfully introduced in America in three densely populated areas. The situation in the U.S.A. is somewhat different to the U.K. in that densely populated areas in the U.S. are more widespread than in this country. These areas are each served by a large number of relatively low-power transmitters. Therefore there would be little likelihood of interference from a mobile operator sharing the frequency of a TV station located several hundreds of miles away, but unused in his own area. In the U.K. on the other hand, an attempt is made to saturate the whole island with television signals, with most of the area covered by 8 high-power transmitters.

Users as a whole, however, cannot be unaware that v.h.f. television Bands I and III may be phased out of use for the existing programmes and the two broadcasting organizations are expected to present a convincing case for retention of the v.h.f. bands for broadcasting before their charters come up for renewal in 1976. Otherwise industry will have its sights firmly fixed on this substantial sector of the frequency spectrum. The threat by mobile radio to amateur frequencies was pointed out in "World of Amateur Radio", in the July issue.

The conference posed the question, whose need is greater, mobile radio or television? This is a choice which may not necessarily have to be made because several delegates spoke of new technologies, or the updating of old ones, which can provide a greater number of effective communicating channels within the existing frequency allocation. There is, for example, the possibility of reducing the u.h.f. channel spacing (currently 25kHz) to 12.5kHz. This would virtually double the number of available channels but an increase in noise would result from a reduction in deviation ratio.

Frequency modulation is in general used on u.h.f. but systems might be improved by changes in the modulation characteristic. E. W. Crompton of the Home Office (police and fire services) described work his department was currently carrying out on double sideband diminished carrier amplitude modulation (d.s.b.d.c.) which has the desirable feature of concentrating most of the transmitted energy in the sidebands and not in the carrier as with conventional a.m. and f.m. A more complex detector than the simple diode demodulator would be required[1] but the post detector circuits could be relatively simpler than in the conventional receiver. Delegates were played a tape recording of speech from the output of a d.s.b.d.c. receiver with a $1\mu V$ signal input. Noise was virtually absent but the intelligibility of the speech left much to be desired. Mr. Crompton postulated that more than one information channel could be accommodated in a single transmission by quadrature modulation of a subcarrier, in the same way that colour TV systems carry chrominance information, and the use of synchronous detection.

A system was described for dialling from a vehicle into a private telephone exchange and another for using the mobile radio as a data link. This latter can provide information on the location of up to 1000 vehicles by a hyperbolic system employing four fixed stations disposed round the vehicle movement area. By measuring the phase shift of a 2.7kHz audio tone and collecting 600-2700-band data in a small computer at the base terminal, the system is capable of giving the positions of up to 2000 vehicles per minute.

[1] Macario, R.C.V. "How Important is Detection" *Wireless World*, April 1968.

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Electronic Building Bricks

5. The electronic circuit

by James Franklin

In order to transmit and process information, say in a television set or a computer, we must provide means for moving the lectrons which we are using to represent the information. In Part 3 we looked at the general nature of this movement—conduction. The medium most widely used for electron conduction is one in which the process can take place most freely—a metal—and which is also reasonably cheap, which means copper. A path for electron conduction is called a *circuit,* and the simplest form of electronic circuit is a continuous loop of wire as shown at Fig. 1. Almost all "electronic building bricks" are elaborations of this.

To cause the free electrons in a conducting path such as Fig. 1 to move,

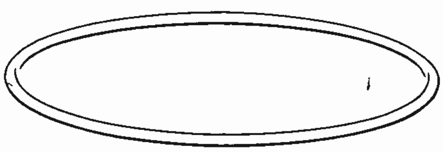

Fig. 1. Basic electronic circuit, a continuous loop of wire or other conduction path for electrons.

one must apply a force to them. Two forces which will act on electrons are the electric field (e.g. the region of "attraction" surrounding a plastic comb which has been rubbed to make it pick up pieces of paper) and the magnetic field (e.g. the region of "attraction" surrounding a magnet). For circuits we use principally the electric field. In practice this field is produced by the rotating generator (to which we make connection through the electricity mains) and the dry battery. These, of course, are both devices for converting some other form of energy—mechanical in the generator, chemical in the battery—into electrical energy. Because the electric field they produce is a force used for moving electrons it is known in the context of circuits as an electro-motive force (e.m.f.). A strong electro-motive force will move more electrons in a given time than a weak electro-motive force.

The unit by which this force is measured is the volt* Thus in a given circuit, 2 volts will move twice as many electrons in a given time (cause twice the current to flow) as 1 volt. There is a very precise scientific definition of the volt, but most people know from experience the relative strengths of different "voltages" by what can be "driven" by them—$4\frac{1}{2}$ volts from a torch battery, 9 volts from a transistor radio battery, 12 volts from a car battery, 240 volts from the electricity supply mains—and these give some practical idea of the e.m.f. presented by the volt. (Anything above about 50 volts will give you a nasty jolt!)

Fig.2(a) shows how a source of e.m.f. is inserted into the Fig.1 circuit to cause the free electrons in the metal to move. The insertion of the e.m.f. source breaks the loop of wire, but the electrons flow through the source, as indicated by the dotted lines. We may also insert some electronic component or device in the circuit, as indicated. The flow-rate of electrons (current) in the whole loop containing the e.m.f. source and the component can be measured by inserting a meter as shown at (b)—and, of course, to allow the current to continue to flow this meter must not break the circuit. Such a meter for measuring current will be calibrated in amperes (Part 3) or, for small currents, milliamperes (thousandths of an ampere) or microamperes (millionths of an ampere).

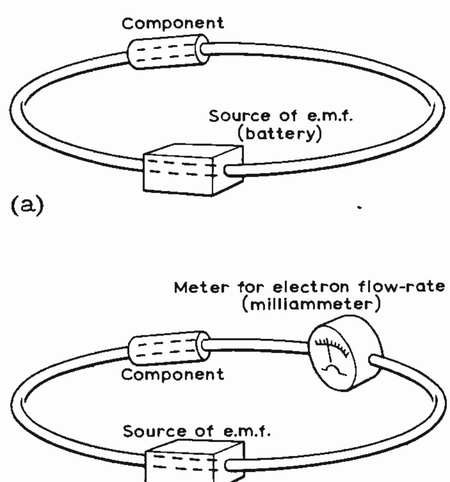

Fig. 2. The Fig. 1 loop with (a) a source of electro-motive force and an electronic component inserted; and with (b) a meter included to measure current.

As will be seen from other articles in *Wireless World,* circuits are normally drawn in a simplified form by the use of symbols. To draw the physical form of the wires and components, as was done in the early days of radio, would be extremely laborious for the more complex circuits, and in any case is unnecessary, except sometimes as an aid to the construction of equipment. All the essential information about the *functioning* of electronic circuits and systems—and this is what we are really interested in—can be given in the simplified, generalized form known as a theoretical circuit diagram (or just "circuit diagram"). Thus the simplified form of Fig.2(b) is shown at Fig.3. It will be seen

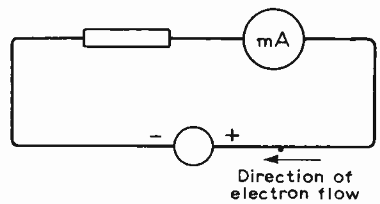

Fig. 3. Theoretical circuit diagram, using graphical symbols, representing the actual circuit shown in Fig.2(b). (The rectangle symbolizes a particular type of component, to be explained later.)

Fig. 4. Graphical symbol for a source of electro-motive force giving an alternating current (a.c.).

that this takes no account of the physical size or shape of the wire, the source of e.m.f. or the meter. We have a generalized representation that could mean any type of conductor (e.g. areas of metal film), any type of e.m.f. source (battery, rotating generator, thermo-electric device, electrostatic machine) and any shape or size of meter. Thus with the removal of non-essentials, our attention is concentrated on the functional aspect.

It will be seen that diagram Fig.3 contains some extra information, the + and − signs on the e.m.f. source. Without going into the full meaning of electrical "positive" and "negative", it is sufficient to say at this point that these signs on this particular source indicate (a) the direction in which the electrons are made to move round the loop by the e.m.f. and (b) the fact that this direction is always the same. A source of e.m.f. producing such a uni-directional current is called a direct-current (d.c.) source—a practical example is a battery— and the positive side of this source is identified as that towards which the electrons travel.

Another type of e.m.f. source used in electronics causes the electrons to flow alternately in opposite directions. This type of flow is called alternating current (a.c.), and the theoretical circuit symbol used to represent an a.c. source—which might be a power-station generator or a laboratory testing instrument (oscillator)—is shown in Fig.4.

*Named after Alessandro Volta, (1745-1827), Italian physicist and inventor of the electric battery, which was originally known as a Voltaic pile

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Personalities

Professor A. L. Cullen, O.B.E. Ph.D., D.Sc. (Eng.), Pender professor of electrical engineering, University College, London, is to receive from the Paul Instrument Fund Committee the sum of £10,717 over three years for the construction of a wideband microwave impedance bridge of high accuracy, using a new and very precise absolute impedance standard. Professor Cullen, who graduated at Imperial College, London, in 1940 and was for six years at R.A.E. Farnborough, occupied the chair of electrical engineering at Sheffield University from 1955 until his appointment to University College in 1966.

George Millington, M.A., B.Sc., F.I.E.E., who recently retired from the Marconi Company, has been appointed consultant to the Directorate of Radio Technology of the Ministry of Posts and Telecommunications. Mr. Millington is well known for his studies of radio-wave-propagation phenomena and was at one time international vice-chairman of the C.C.I.R. study group investigating the problems of ground-wave propagation. A graduate of Clare College, Cambridge, he joined Marconi's in 1931 and headed the company's propagation section for many years.

W. F. Hawes has been appointed general manager for marketing by Pye Telecommunications Ltd. Mr. Hawes, who is 49, was formerly overseas marketing manager and succeeds **J. C. Turnbull** who has become managing director of the recently formed Pye Business Communications Ltd.

Alec Kravis, O.B.E., who is 51, has been appointed sales manager of the Radio Communications Division of Marconi Communication Systems Ltd. He has been with the Marconi Company since 1950 and was at one time project co-ordinator in Research Division for advanced communication and radar projects. In 1964 he was appointed project co-ordinator for space communication studies, and when Marconi won the contract for the first British military satellite communication stations Mr. Kravis was made project manager. After the completion of this project he became manager of administration and technical services in the Research Division, and since 1967 has been in the Computer Division, as deputy manager.

Donald W. Morrison, B.A., who joined the Sprague organization earlier this year as special assistant to G. V. Tremblay the president of Sprague World Trade Corp., has been appointed managing director of Sprague Electric (U.K.) Ltd, of Yiewsley, Middx. From 1960 to 1967 Mr. Morrison was director of marketing and general manager for the Far East of A.M.F. International after which he spent two years as regional manager for Europe with the American Air Filter Co.

Alan Hall has joined Marconi Communication Systems Ltd as sales manager of the Specialized Components Division at Billericay, Essex. After studying at Sheffield University Mr. Hall, who is 45, served in the Royal Signals from 1945 to 1948. He worked as a sales engineer with Solartron Electronic Group and with Muirhead, then as assistant to the technical manager in the Electronic Component Division of Johnson Matthey at Burslem, and latterly as sales manager for Oxley Developments Co. Ltd, of Ulverston, Lancs, Mr. Hall is an amateur radio trasmitter using the call G3UWA.

Alan Hall

The appointment of **John S. Walker** as managing director is announced by Cosmocord Ltd, of Waltham Cross, Herts, manufacturers of Acos electro-acoustic products. Mr. Walker has been associated with the electronics industry for nearly 20 years. He was until recently with De La Rue Instruments, before which he spent 10 years with Texas Instruments Ltd, where he was in turn responsible for the application laboratories and for research and development.

Ernest M. Hickin recently joined Microwave Associates Ltd where he is now responsible for co-ordinating all the technical activities of the Luton-based subsidiary of Microwave Associates Inc., of Burlington, Mass. Mr. Hickin was previously for six years with GEC-AEI Telecommunications Ltd at Coventry as chief radio engineer, Transmission Division, having earlier been responsible for transmission research at the GEC Hirst Research Centre, Wembley. He is vice-chairman of the I.E.E. South Midland Electronics & Control Section.

J. D. Rhodes, B.Sc., Ph.D., an authority on microwave selective linear-phase filters, has been appointed design consultant to the Valve & Microwave Group of Ferranti Ltd at Dundee, Scotland. Dr. Rhodes, who is 26, is a lecturer in the Department of Electrical & Electronic Engineering at Leeds University, where he graduated and obtained his doctorate. He then spent one year as a post-doctorial fellow before going to work for Microwave Development Laboratories Inc., Natick, Mass., U.S.A. There he was engaged upon general research and development into microwave techniques with the emphasis on linear-phase and elliptic function filters. During 1969, Dr. Rhodes returned to Leeds to take up his present post at the University.

J. E. Diggins, M.B.E., who joined Racal in 1963 as an electronics engineer at Bracknell, is appointed deputy managing director of Racal-BCC Ltd, the largest company in the Racal group. In 1966 he headed a team of engineers whose task was to develop a new h.f. mobile radio-telephone. The project was so successful that a separate company was formed, named Racal-Mobilcal Ltd, and production transferred to a new factory in Reading. **E. T. Harrison** is managing director of Racal-BCC Ltd, in addition to being chairman and managing director of the Group. Following Mr Diggins appointment **G. J. Lomer** has become general manager in addition to being technical director of Racal-Mobilcal Ltd. **D. C. Elsbury,** formerly chief inspector and quality assurance manager of Racal-BCC, Bracknell, becomes production director of Racal-Mobilcal and **E. Phillips,** formerly production manager of Racal-Mobilcal becomes chief inspector and quality assurance manager of Racal-BCC Ltd.

The appointment of **J. O. M. Jenkins,** M.Sc., as applications engineer for digital integrated circuits is announced by Siliconix Ltd, of Swansea. After graduating in electrical engineering at the University of Swansea, Mr. Jenkins, who is 30, went to the Steel Company of Wales in 1961. In 1966 he relinquished his appointment to study for his Master's degree at Cranfield Institute of Technology following which he joined Mullard, Southampton, to work on the development of integrated circuits.

James A. Scott, appointed sales manager of K. W. Electronics Ltd., Dartford, Kent, was previously assistant to the sales manager of the radio navigational aids division of Standard Telephones & Cables Ltd. After studying at Oxford Technical College Mr. Scott, who is 41, joined the Scientific Civil Service at the Atomic Energy Research Establishment at Harwell. Following transfer to the Overseas Civil Service he held executive positions in the security services radio branch and in the civil aviation department in Kenya, until the independence of that country. He held an amateur radio call sign in Kenya, and now operates in this country with the call G3CMI.

J. R. Tillman, D.Sc., Ph.D., M.I.E.E., deputy director of research at the Post Office Telecommunications Headquarters, London, has been appointed a visiting professor in the Department of Electrical and Electronic Engineering of the City University, London. Dr. Tillman joined the Post Office Research Station in 1936 and was appointed deputy director of research there in 1965.

B. A. Paine, B.Sc., is appointed managing director of Booker Bowmar Ltd and its subsidiaries Reliance Controls Ltd and Bowmar Instruments Ltd. A graduate of California State Polytechnic College, Mr. Paine, who is 37, was, until his new appointment, operations manager with Spectral Electronics Corp., of Southern California. He succeeds **L. M. Butler** who has gone to California to become president and general manager of another Bowmar company.

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Television Wobbulator

3. Construction and operation

by W. T. Cocking, F.I.E.E.

The whole equipment has been built into a standard case (Olson Type 75B) with the power supply assembled on the back panel ($4\frac{5}{8}$in × $11\frac{5}{8}$in). This was mounted at one end on a hinge so that it could be opened out, but this proved to be quite unnecessary and would not be done in a second model. It is mentioned merely because the hinge may be evident in some photographs.

All other parts are mounted directly or indirectly on the front panel, which is of the same size as the back panel. Connection between the two is made by an 8-way cable terminated by an 8-way connector. Another identical connector is screwed beneath the box which contains the marker oscillator and the two joined together by short lengths of No. 16 wire. The cable is long enough to enable the front unit to be withdrawn from the case and turned over, and it can then be disconnected by slacking off eight screws.

General view of the wobbulator showing all the external controls

In this way good screwed connections are obtained and, by using two connectors, there is no possibility of making wrong connections on rejoining them as there would be if wires were disconnected independently.

Components are not critical and an indication is given of a supplier of certain parts which have been used in the model described. Only one component is really important—the Motorola 1N5145—and this was obtained from Celdis Ltd. (37/39 Loverock Rd., Reading, Berks, RG3 1ED). The photographs show most of the constructional details. The marker oscillator is built on a piece of plain Veroboard with a 0.1-in matrix of holes and measuring $2\frac{1}{8}$in by $2\frac{7}{8}$in. The variable capacitor used (Home Radio type VC20, Jackson U101/SS, 25pF) is a split stator type of 25pF each half; only one half is used. It is screwed to the Veroboard, on which the other parts are mounted in the usual way. A long 6BA screw for mounting to the box is fixed in each corner of the board and nuts are run on, not only to fix the screws to the board but so that the spacing of the underside of the board from the case can be adjusted.

The screening case for it is a metal box measuring 3in × 3in × 2in (Home Radio Z127). This must have four holes drilled in its bottom to take the four screws from the Veroboard. A $\frac{1}{8}$-in hole must be drilled in one side as a clearance hole for the capacitor shaft extension and in the same side four 6BA holes for the screws by which the box is attached to the front panel. It is mounted with spacers so that the box lies 1in behind the panel. It is desirable to tap the holes in the side, since this saves getting nuts on to the screws in rather awkward positions. An extension shaft of insulating material is used for the capacitor with the usual metal coupler inside the screening can.

Another hole in the bottom is fitted with a grommet to take two power-supply leads and the twisted pair for the r.f. output. One connector is screwed to that side of the can which is at the bottom when it is mounted on the front panel.

The main assembly is on a piece of plain Veroboard, again with a 0.1-in matrix of holes and measuring $3\frac{3}{4}$in × $6\frac{7}{8}$in. It is held to the panel by brackets. The layout is not critical but all connections within a tuned circuit should be kept short.

Printed circuits do not lend themselves at all well to development work and instead we used plain Veroboard with Cir-Kit for "wiring", for it permits easy alteration. It proved very successful, and we retained it in the final model. Cir-Kit comprises copper strip and is available in widths of $\frac{1}{16}$in and $\frac{1}{8}$in. It is coated on one side with an adhesive and a paper backing to prevent it from sticking together. A piece of the required length is cut off the roll, the backing paper is peeled off and it is placed on the board in the required position and pressed firmly down. The $\frac{1}{8}$in was used only for the earth, 17V and 70V lines, the $\frac{1}{16}$in being used for everything else.

It works very successfully as long as there is no push or pull on it which tends to separate it from the board. A component on the top of the board, therefore, should not be spaced from the board but should rest on it and its leads should be bent over flush with the board on the under side, so that the component and its leads tend to clamp the Cir-Kit strips to the board. It is wise to let a lead cross the copper strip and project a little on the other side. If it should be necessary to remove a component a knife blade can then be slipped under the projecting wire while the soldering iron is applied and the lead bent up without disturbing the copper strip.

The adhesive softens with heat and this can cause trouble when unsoldering a lead from a very small piece. It is then sometimes easiest to pull out the copper strip while it is hot and replace it with a new piece.

Connecting leads and any components on the under side of the boards should be

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attached only at places where the leads from parts on the top of the board are providing a firm anchorage for the strip to the board. In other places, a connecting lead should be anchored to the board by enlarging two holes, passing the lead through one from bottom to top and then through the other from top to bottom.

It is worth while to check all components before mounting them, including semiconductors. Using the Model 8 Avometer on the ohms range most diodes and base-emitter junctions show from about 400 Ω to 2k Ω in the forward direction and appear to be open-circuit in the reverse direction. Other ohmmeters may give very different resistance values, but as long as the resistance in the forward direction measures much less than in the reverse direction there is a presumption that all is well. This applies to zener diodes as well as ordinary types, and also to the varactor diode.

Setting-up

Until further notice the vision-sound switch is to be in the vision position. No i.f. amplifier need be connected. Connect a voltmeter across R_{11} Turn R_{22} for zero a.c. drive to Tr_3. Check that as R_{20} is rotated the voltage across R_{11} can be varied from almost zero to 70V. Set R_{20} for about 20V.

Disconnect the voltmeter and connect the c.r.o. between earth and the junction of R_9, R_{10} and D_3. Turn up R_{22} to apply a.c. drive to Tr_3. Use the linear timebase of the oscilloscope and synchronize it to the waveform (50Hz). The waveform should at first appear sinusoidal but, as the amplitude increases, it should more and more tend to very rounded positive half cycles with peaky negative half cycles. If the amplitude is too great or the bias on Tr_3 is wrong, or both, there will be flats on the bottoms of the negative half cycles. This occurs when D_3 conducts it and it shows that the safety circuit is functioning. Adjust R_{20} and R_{22} for the maximum possible amplitude of the waveform without overloading. This will be about 50V p-p.

It is important to realize that the oscilloscope is connected to a high impedance point (about 0.4 MΩ) with the result that unless the oscilloscope is of much higher impedance the voltage measured by it will be a good deal less than the true voltage. If R is the input resistance of the oscilloscope, the voltage indicated by it should be multiplied by $1 + 0.4/R$. We used an oscilloscope of 1MΩ input resistance and measured on it a maximum of 38V p-p, making the true voltage 53V approximately.

Connect the c.r.o. to the "Output to c.r.o. Y-amp" socket and set its Y sensitivity to about 1V, or a little less, for full vertical deflection. Connect its X-input to "Output to c.r.o. X-amp" socket. Connect the signal generator to "Input from s.g.". The signal generator should have an open-circuit output impedance of 75 Ω and an open-circuit output of at least 200 mV. Set it at full output at 36.5MHz. Turn R_{56} to maximum. Set the internal marker oscillator control so that the variable capacitor vanes are fully unmeshed. Set the core of L_4 so that its top is a little below the top of the former.

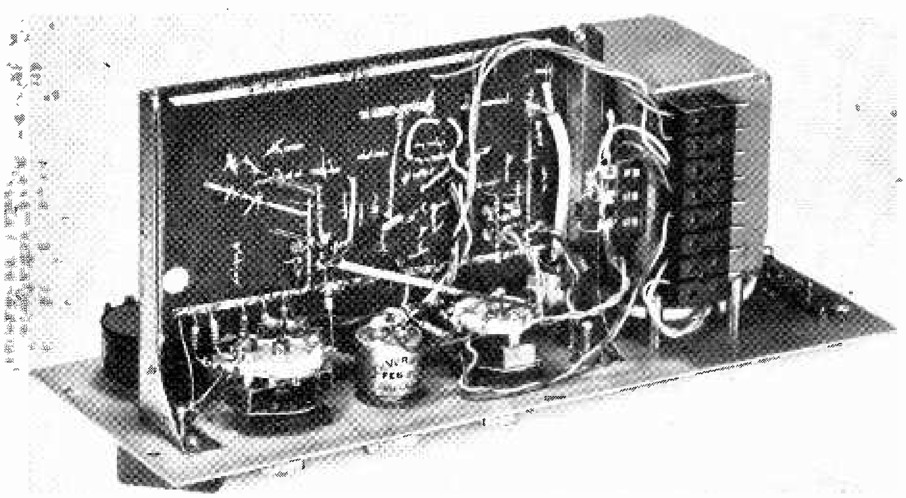
There may now be visible on the trace anything from none to three markers. If there are three the largest will usually be from the internal marker oscillator and the next from the signal generator. This may be checked by adjusting their controls independently; the markers will then move along the trace independently. The third marker is much smaller in amplitude and is always about half-way between the other two. Also it moves when either of the other markers is moved and in the same direction, but at half the rate. It arises because a frequency equal to the sum of the two marker frequencies is generated in the D_4 circuit and it beats with the second harmonic of the wobbly oscillator. Its frequency is precisely half way between the two marker frequencies and it is useful in adjusting for linearity, for this third marker should be precisely half way between the other two markers on the trace.

For the moment we are concerned only with the marker at 36.5MHz provided by the signal generator. If it is not visible adjust R_{20} to bring it on to the trace, and in any case adjust this control to bring this marker as nearly to the centre of the race as possible.

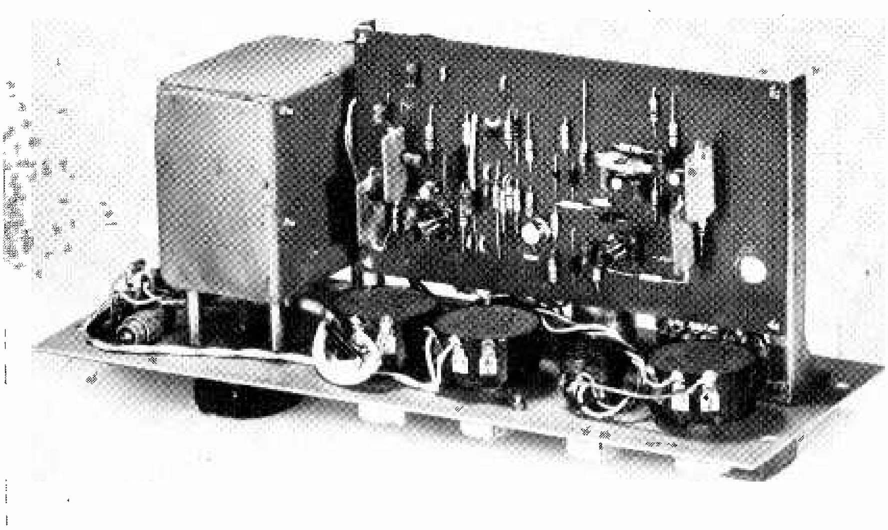
Unless the phase adjustment is by chance set correctly, each marker will be double. Adjust "Phase" to make the markers on the forward and return sweeps coincide.

Check that increasing the marker frequency makes the marker move along the trace to the right. If it moves to the left, reverse the leads to one of the 7-V transformer secondaries.

Adjust the signal generator towards 30.5MHz while watching the trace. The marker may disappear from the left-hand end of the trace before this frequency is reached or it may not have reached the end at this frequency. In either case adjust R_{20} to bring the 30.5-MHz marker precisely to the end of the trace. This is when the end of the trace comes to the middle of the marker, so that only half of the marker is visible.



Upper part of the wobbulator unit.



Lower part of wobbulator.

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Now turn the s.g. towards 42.5MHz. There are now several possibilities. The marker may disappear beyond the right-hand end of the trace at some frequency above 42.5MHz. If it does this readjust R_{20} and R_{22} so that when the s.g. is set in turn to 30.5MHz and 42.5MHz the marker is precisely on the left and right hand ends of the trace respectively.

If the marker stops before 42.5MHz is reached, it may still be possible to make it reach the end while still keeping 30.5MHz at the other end, just by adjustment of R_{20} and R_{22}. More likely, however, the third possibility will arise. This is that the marker will stop before 42.5MHz is reached but, while its left-hand half will be undistorted, its right-hand half will be drawn out into an oscillation for the rest of the trace. If this happens unscrew the core of L_1 a little and repeat the process.

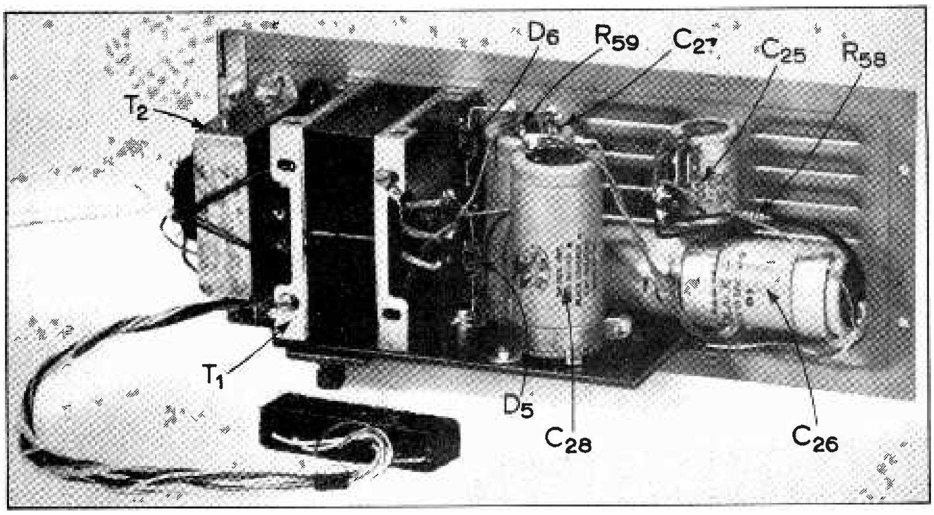
Having obtained a sweep covering 30.5MHz to 42.5MHz, calibrate the internal marker oscillator against the signal generator. It is sufficient to do this in 1-MHz steps. It is convenient to use the blank half of the dial with the slot of an adjacent screwhead as an indicator, and place a pencil mark on the dial for each calibration point.

Set the s.g. to the required frequency and turn the dial of the marker oscillator to superimpose its marker on that provided by the s.g. When the two are nearly equal in frequency a low-frequency oscillation will appear right across the trace and it will disappear at a very critical setting when the two are precisely equal. Mark this point on the dial, move the s.g. by 1MHz and repeat, and so on until the calibration has been completed.

The calibration is easily checked in this way at any time, and if the marker is needed at any frequency in between calibration points it can be set there by reference to the s.g. When the marker is calibrated, the setting up procedure is very much easier because a marker can be set at each end of the range, and it is unnecessary to be continually turning the signal generator from one end of the range to the other, possibly by a tedious slow-motion control.

Set the internal marker oscillator so that its marker is off the trace. Set the s.g. to 36.5MHz and adjust the core of L_4 for maximum amplitude of the marker. Check that the amplitude is about the same at 30.5MHz and 42.5MHz and is about 70% of that at 36.5MHz. Readjust the core if the amplitudes at the two ends are not almost the same. The adjustment is not critical.

The next step is to check the linearity. This is done with the aid of the third marker. Set the s.g. and the internal marker to 30.5MHz and 40.5MHz and see that they lie precisely on the two ends of the trace; slight readjustment of R_{20} and R_{22} may be needed because of temperature drift. Measure the distance of the small centre marker to each of the two end markers. The two distances should be the same. If they are not adjust the core of L_1 very slightly and readjust R_{20} and R_{22}, and measure the distances again. If the distances are more nearly equal continue adjusting the core in the same direction, each time readjusting R_{20} and R_{22} until they become equal. Of course, if the first adjustment to the core makes matters worse, it needs adjusting the other way.



General view of the power supply unit.

As an example, it was found in one instance that the distance of the centre marker to the left was 35 arbitrary units, whereas the distance to the right was 32 units. About one turn only of the core was needed to bring the distances to equality.

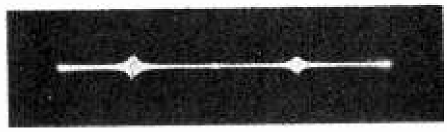
The wobbulator is now set up on vision and can be used with an i.f. amplifier to depict its response curve. Proper alignment of the vision amplifier with its trap circuits must be achieved before any attempt is made to align the inter-carrier sound channel.

There are certain traps for the unwary in doing this. The first is a short-circuit on the mains supply! Never for one moment forget that the normal television receiver is live to the mains. To avoid the risk of shock, damage to equipment and even to obtain proper operation of equipment, there is really no alternative but to use a double-wound 1:1 ratio mains transformer to feed the television set. It is certainly the only safe thing to do. Otherwise the i.f. strip must be removed from the set and operated from a separate power unit containing a transformer.

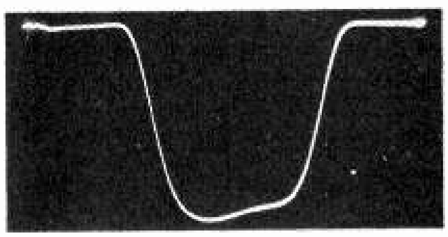
Most television sets are designed to operate with the video detector giving an output of 2V to 4V. Care should be taken to see that the peak-to-peak output with the wobbulator is of this order. However, the input from the wobbulator must never be so great that overloading occurs.

It is essential that the connection of the video output to the wobbulator should not cause feedback in the i.f. amplifier. A well-screened lead is necessary and it is sometimes better to take the output from a video stage than from the detector itself. In the case of the *Wireless World* Colour Television Receiver it was found to be best, and certainly most convenient, to use the sync separator feed. This is a long screened cable which plugs into the sync separator board, and so it can just be unplugged and plugged into the wobbulator.

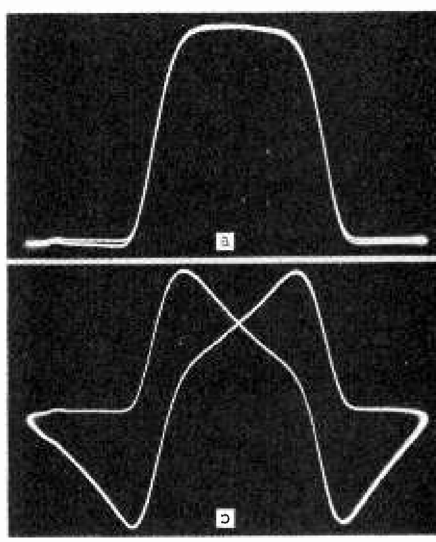
The r.f. output from the wobbulator must be by coaxial cable and its length



Here the markers are at 34MHz and 40MHz and half way between them there is a very small marker at 37MHz. This arises from the sum of the main markers (74MHz) beating with the second harmonic of the wobbly oscillator.



Taking the output from the luminance delay line of the W.W. Colour Television Receiver results in an inverted trace.



The normal input coupling time constant of the oscilloscope is 0.25 sec. These photographs show the result of reducing it, at (a) to 0.04 sec and at (b) to 0.001sec. Notice the double trace in the vertical direction in (a) and the gross distortion in (b).

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should be as short as possible. It is not practicable to match the cable perfectly at each end. At mid-band only about 4ft. of cable is needed for a quarter-wave section. It is well, therefore, to arrange the apparatus so that the cable is no more than 2ft. long if possible. It is not often practicable to feed into the tuner of the television set, so that the first i.f. circuit, which is normally in the tuner, can be in circuit. It is useful, therefore, to have a dummy tuner built to the circuit of Fig. 1 to connect the wobbulator to the i.f. strip The cable from the dummy tuner to the i.f. strip must, of course, be the same length as that from the real tuner to the i.f. strip in the receiver.

Very often additional attenuation to that provided in the wobbulator will be needed. This is conveniently obtained from Belling Lee coaxial attenuators, which are available with attenuations of 3, 6, 12, 18 dB; they plug into each other and to the normal cable plugs and sockets. It is particularly convenient to have three 6-dB (L729/6) types available. Whenever possible one should be between the wobbulator cable and the dummy tuner. This is the point at which the cable has its greatest mismatch, because of the tuned circuit in the dummy tuner, and this mismatch is greatly reduced by introducing an attenuator at this point. Even when there is sufficient range on the internal attenuator, it may pay to include a 6-dB attenuator at the cable end, and use a step or two less internal attenuation.

For the sound channel the overall picture including the discriminator is obtained by feeding the a.f. output of the sound detector to the "Input from i.f. amplifier" socket. For the alignment of the early circuits, however, it is advisable to make up an a.m. detector to the circuit shown in Fig. 2 and to take the output from this. It is connected to the collector of the last sound i.f. transistor, for instance.

To operate the wobbulator on sound, set the vision-sound switch to sound. Set the s.g. to 39.5MHz, and then set the marker oscillator to zero beat with it. This can be done even if there is no visible marker on the trace, because near the proper setting the low-frequency beat between them extends over the whole trace. It is necessary to do this because the calibration of the marker oscillator does not hold when it is switched to sound. It produces 39.5MHz when it is set roughly to the 40.5-MHz calibration point. If one wishes one can put on a special 39.5-MHz point for sound, but it is so easy to set the marker against the s.g. that it seems unnecessary. The change occurs because there is a change of loading on the marker oscillator between the two positions of S_1.

Turn the "Mid-band frequency control" R_{20} fully clockwise (if it is wired so that on vision a clockwise rotation shifts a response curve to the right) and then gradually back until the sound response curve appears. This is the correct one. If it is turned further another of less amplitude will appear, and if one goes on turning others will come into view, and there may even be several overlapping ones. The first and, usually, largest is the proper 6-MHz inter-carrier beat between 39.5MHz and 33.5MHz in the vision i.f. detector. The second occurs between 39.5MHz and 36.5MHz, when the separation is 3MHz, and it is the second harmonic of this produced in the detector which passes through the sound i.f. amplifier. The others occur for separations of 2MHz, 1.5MHz, 1.2MHz, 1MHz and so on, with the third, fourth, fifth and sixth harmonics.

By adopting the above procedure the right setting is easily found and the spurious responses cause no trouble. Notice, however, that the proper response will not always be the largest. If the sound trap in the vision i.f. amplifier gives more attentuation than usual, the second response may be larger than the proper one.

On both vision and sound an increase of marker frequency moves the marker to the right. On vision a clockwise rotation of R_{20} moves the response curve to the right on the trace, but on sound it moves it to the left. This could be avoided by extra switch contacts to reverse the leads to R_{20}, but did not seem worth while. Temperature drift is normally to the left on vision and to the right on sound.

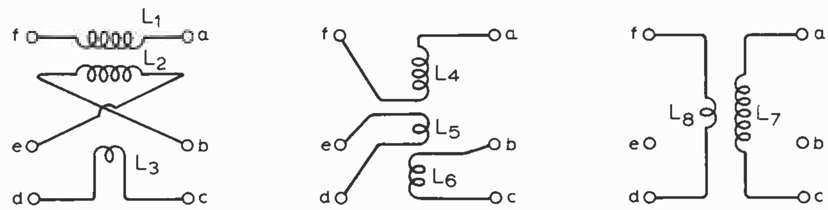
One thing must not be overlooked. It is usually necessary in some way to put out of action any a.g.c. system of the amplifier under test and the way of doing this will depend on the receiver design.

Coil data

Details of windings: L_1 10t; L_2 10t; L_3 1t; L_4 14t; L_5 1t; L_6 2t; L_7 10; and L_8 2t.

Formers are Neosid 722/1 with terminal bases 5027 and Neosid long screw cores 4 x 0.5 x 12.7. The former diameter is $\frac{3}{16}$ in. with an available winding length of $\frac{3}{8}$ in. Except L_2, all coils are wound with No. 24 enamelled wire; L_2 is wound with No. 32.

In every case the start of the main winding is next to the base of the former and is the 'hot' end of the coil. All coupling coils, except L_2, are adjacent to the main winding at the earthy end. L_1 and L_2 are bifilar. Wind L_1 first, close-wound with the start soldered to its pin. Leave the other end free. Solder the wire for L_2 to its start pin and wind it in



between the turns of L_1. The turns of the latter will spread out to accommodate the wire and it will unwind very slightly. Terminate L_2 on its earthy pin and then L_1 on its earthy pin. Note that the "hot" and earthy pins of the two coils are not adjacent. It is important that the same ends of the two coils should be earthy. Reversed connections to one coil will prevent the proper frequency coverage from being obtained.

The pin spacing of the bases does not fit the hole spacing of Veroboard. It is advisable, therefore, to make a metal template so that the Veroboard can be drilled in spite of the existing holes. The base is fixed to the board by putting small washers or loops of wire around four of the pins and soldering.

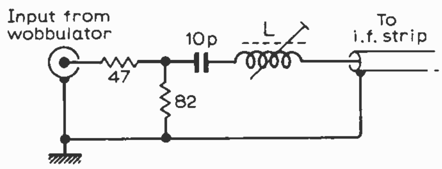


Fig. 1. Circuit of dummy tuner. The coil L can be 15 turns of No. 30 wire wound on a former like those used for the wobbulator coils.

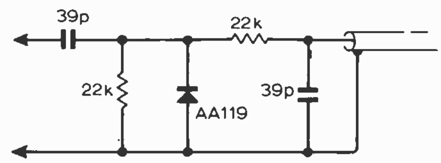
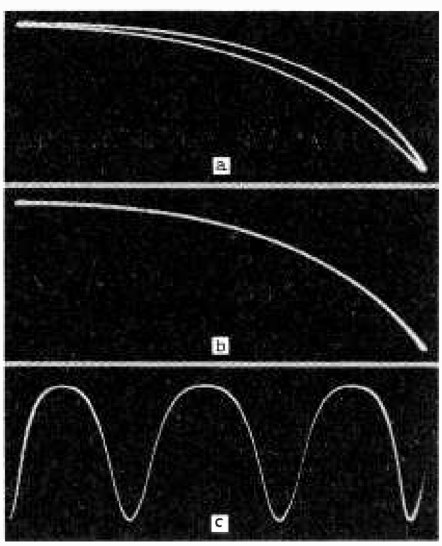


Fig. 2. Circuit for connection between earth and the last sound i.f. collector to view the overall response apart from the discriminator.



The voltage waveform at the collector of the BF177 wave shaper is shown in these photographs. At (a) the waveform is shown with a sinusoidal timebase and the double trace is due to phase shift caused by the connection of the oscilloscope. Adjusting the phase control removes this (b). The waveform with the more familiar linear timebase is shown at (c).

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A YIG Radiometer and Temperature Controller

Solid-state device for temperatures of 100-250°C

*by I. J. Kampel**

The radiation pyrometer or radiometer provides a convenient way of measuring surface temperature where more conventional sensing elements may not be employed. The system to be described also utilizes an experimental solid-state optical modulator—the YIG—and in so doing eliminates the bulky and inconvenient motorized chopper unit normally associated with such systems. Use of this new electro-optical component therefore answers the need for a sensing head requiring no maintenance, and provides a rugged, miniature solid-state construction.

All bodies emit electromagnetic radiation, the amount of radiant energy being dependent upon their temperature and emissivity. The maximum theoretical emission at any temperature is given by a black-body radiator, and the "emissivity" factor determines the ratio of emittance by a body to the emittance of a black-body at the same temperature.

The Stefan-Boltzman law states that the total spectral emittance of a black-body is directly proportional to the fourth power of its absolute temperature. The radiometer—or radiation pyrometer—utilizes this principle. Emitted radiation is measured, and this is then equated to source temperature. A control is normally provided to allow for emissivity factor corrections.

While the disadvantage of temperature measurement by radiometer lies in the need for approximate evaluation of source emissivity, the system does offer unique advantages over more conventional techniques. The infrared sensing element represents the perfect measuring transducer: it in no way affects the object being measured since it is non-contacting. This particular versatility also allows the remote measurement of moving bodies. As measurement is by means of a photo-electric as opposed to the more common thermoelectric principle, a fast response may be realized. Many control systems do not require an absolute temperature indication, and in such cases the controller may be used without prior determination of source emissivity.

For accuracy, and the convenience of a.c. amplification in radiometer systems, it is normal to chop the incoming radiation with a rotating segmented chopper-disc. The size of such a disc with its associated driving motor has necessarily led to rather bulky sensing heads in previous equipments, however the YIG solid-state optical modulator removes such size restrictions. The total absence of all moving parts has also eliminated the need for servicing.

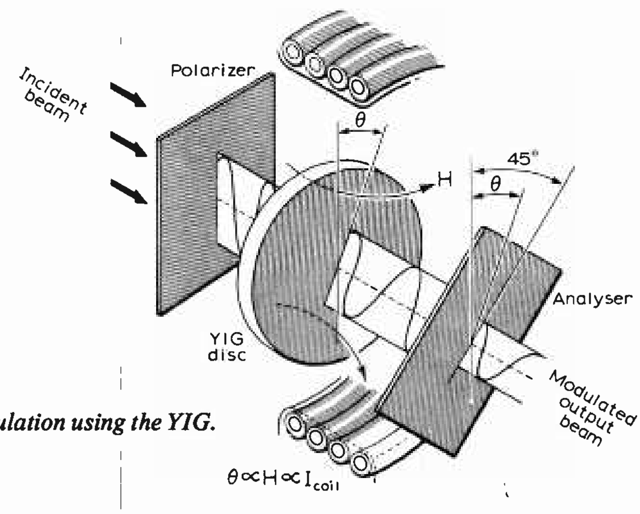

Fig. 1. Intensity modulation using the YIG.

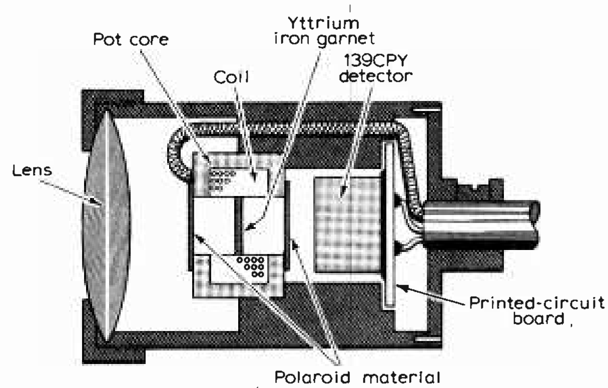

Fig. 2. The miniature radiometer sensing head.

YIG Modulator

Unbalance in electron spin of inner orbital electrons induces a magnetic moment in a crystal of yttrium iron garnet (YIG): it is therefore said to be ferrimagnetic. Magnetic domains of identically orientated magnetic dipoles form and take up a natural alignment. These domains exhibit Faraday rotation; in the presence of a magnetic field the plane of polarization of transmitted plane-polarized radiation is rotated through an angle proportional to the magnetic field strength. Faraday rotation is also proportional to path length through the material, and is inversely proportional to the wavelength of radiation.

The high transmission of YIG in the spectral range 1.1-4.5μm is enhanced by the use of an anti-reflection coating. Its usefulness is limited by the rapid deterioration in Faraday rotation with increasing wavelength. The polarizing material normally employed in the modulator also limits the spectral response to about 2.3μm. Lead sulphide is the most suitable detector material to employ in radiometric applications due to its similar spectral response characteristic.

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Current passing through a coil situated around the material provides the required magnetic field, with lines of flux lying parallel to the path of the traversing radiation. A square-wave current drive provides a suitable modulation of the plane of transmitted plane-polarized radiation, and by further passing this radiation through another polarizer—known as the analyser—an intensity modulation is derived. This principle is illustrated by Fig. 1.

A sectional diagram of the sensing head is shown in Fig. 2. Overall dimensions are 2cm diameter by 3cm length.

A glass lens is used to focus incident radiation on to the element of a 139CPY chemical lead sulphide detector. In the path of this radiation is situated an experimental YIG modulator; this was housed within a modified Ferroxcube transformer pot core which served to provide a low reluctance path for external flux, so minimizing direct electromagnetic pick-up at the detector. Polarizer and analyser are located on the outer faces of this housing.

The plane of polarization of the analyser is situated at 45° relative to the polarizer. This requires a ± 45° rotation of the plane-polarized radiation for full theoretical modulation, and this is approached by employing a true a.c. drive current in the YIG coil. A greater modulation depth for a given drive power may be achieved with this method than by any other.

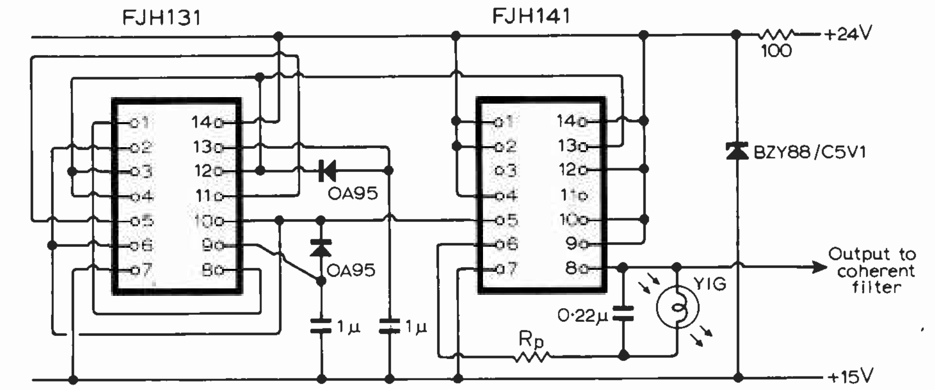

Fig. 3. Drive circuit for the YIG.

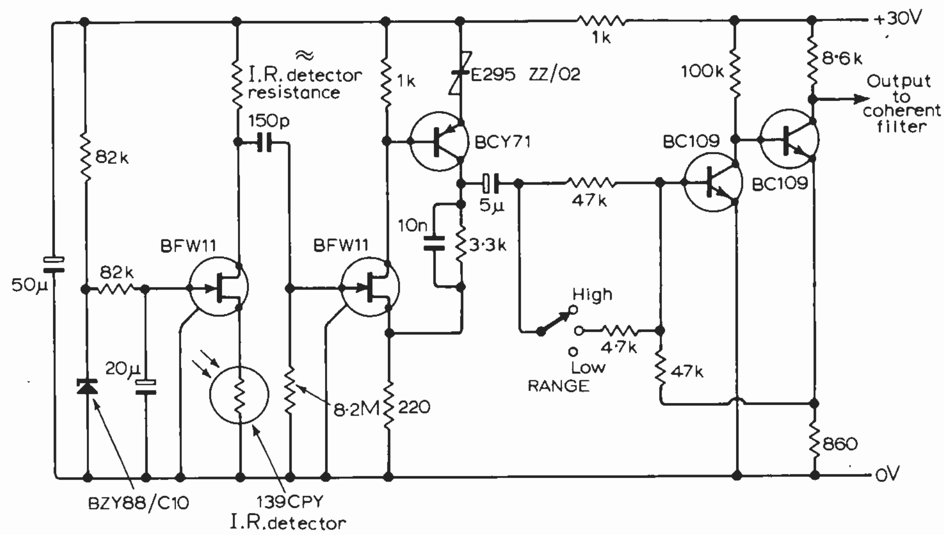

Fig. 4. Low-noise preamplifier circuit with compensatory detector bias.

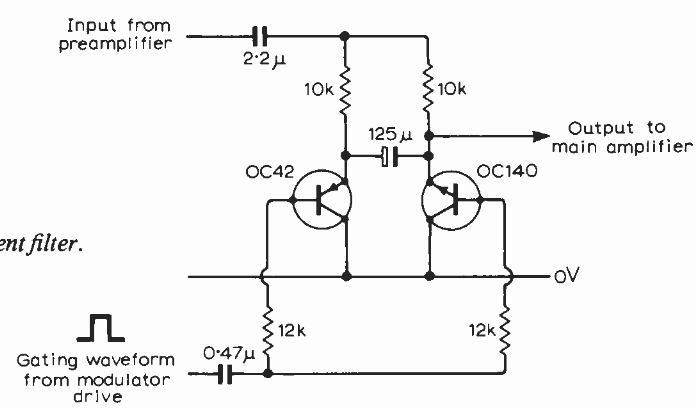

Fig. 5. A simple coherent filter.

System description

Output from the radiometer takes the form of a moving coil meter; output from the controller may either be a simple audible warning, or a change-over relay, but various operational modes may be selected to give this considerable flexibility.

A chemical lead sulphide detector was employed since this had a suitable spectral response, and was available in the small size of a TO-5 encapsulation. The system to be described is suitable for a black-body source temperature range of 100-250°C, covered by two ranges; the addition of higher temperature ranges would be straightforward. An emissivity control covering the range $e=0.1$ to $e=1.0$ was provided.

Two integrated circuits are used to modulate the YIG coil with a true a.c. square-wave current at a frequency of 400Hz as shown in Fig. 3. This frequency is sufficiently high to avoid mains pick-up problems, and low enough to allow optimum performance from a lead sulphide detector having a somewhat slow response.

A quadruple NAND gate (FJH131) is interconnected to form a square-wave generator, where external 1-μF capacitors set the periods as in a conventional multivibrator. Since the resistive component of the time-constant is contained within the integrated circuit, however, this is liable to production spreads and will introduce some spread in frequency. (See Fig. 3).

A dual buffer gate (FJH141) provides the required current drive to the YIG coil. The coil is d.c. coupled and driven in push-pull mode. The voltage amplitude from this integrated circuit is also dependent upon production tolerances, and R_p is adjusted on test to give a coil current of approximately 20mA (33-68Ω typical). A capacitor in parallel with the coil reduces the spike caused by fast current switching in the inductive load.

Modulator and detector leads should be adequately screened from one another in order to avoid pick-up problems.

Pre-amplifier Circuit

Detector resistance varies with ambient temperature. Since responsivity is approximately proportional to the square of this resistance, it is necessary to provide some correction. By biasing the detector in a constant voltage mode and utilizing the signal current, the output obtained is inversely proportional to the square of detector resistance and a much improved temperature stability is achieved.

An f.e.t. is used to provide detector bias, and was found preferable to a transistor, introducing far less current noise than the latter. The complete circuit is shown in Fig. 4. Noise from the zener diode is eliminated by decoupling the gate of the f.e.t. A second f.e.t. is used in a compound arrangement with a transistor to provide initial gain; this is also a low noise configuration. A voltage-dependent resistor in the transistor emitter allows a greater drain resistance for improved loop gain. A further gain of 10 or 100 is provided by the following virtual-earth amplifier, the gain being selected by the range switch.

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A simple coherent filter is formed from two complementary germanium transistors as shown in Fig. 5. These effectively form a dynamically synchronized narrow-bandwidth filter. A gating waveform is taken from the YIG drive circuit and applied to switch these transistors in anti-phase. The output from the pre-amplifier is fed into the circuit, and is used to charge the 125-μF capacitor through one or another of the 10-kΩ resistors.

The two transistors are switched in synchronism with the signal, and earth alternate sides of the capacitor so that a d.c. charge builds up which is proportional to the signal waveform. Any noise or spurious signals present at frequencies other than the modulating frequency average to zero. An output is taken from one side of the charge capacitor for further a.c. amplification.

It is important to ensure a true 1:1 mark-space ratio for optimum performance from the coherent filter. It will be obvious that drift of the chopping frequency will in no way affect performance. The transistors are connected in an inverted manner to minimize leakage current and breakthrough of the gating signal.

An operational amplifier (TAA811) is used to provide a further stage of variable gain and an emissivity control. The configuration of Fig. 6 will provide a linear emissivity scale calibration between 0.1 and 1.0 provided that the potentiometer is truly linear (selection normally required). Voltage gain will be 100 at e =0.1, and 10 at e=1.0.

The radiometer section of the instrument terminates in a peak detector circuit and is shown in Fig. 7. The 1-μF capacitor charges to a d.c. potential proportional to the signal amplitude. This is fed to a moving-coil meter and provides the temperature indication. The zener diode is a protective device for the meter.

The d.c. level on the charge capacitor is also monitored by the controller circuitry. Provision is made to switch the meter to a variable d.c. potential independent of the signal: this greatly facilitates initial adjustment of the controller.

Temperature controller

Totally independent adjustable maximum and minimum limits are provided, and either or both may be used. In the "basic" mode, if the meter pointer crosses either limit, the relay or audible warning switches on; hence the output does not distinguish between maximum and minimum (although the meter indicator will). In the "cycle" mode, the relay or audible warning switches on when the pointer exceeds the maximum limit, and will remain on until the lower limit has been crossed.

Thus by using both limits in this mode, a cycling controller is available between preset temperatures. By using only one limit in this mode, a one-shot function is available provided that the initial state has been correctly set.

The input to the control circuit is taken directly from the output of the radiometer circuit, and responds either to a d.c. potential set by the signal waveform, or a potential manually set on the "meter-set"

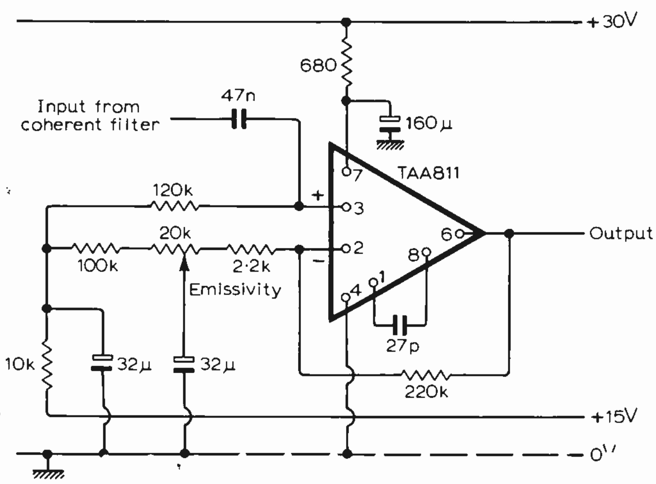

Fig. 6. Main amplifier with emissivity control.

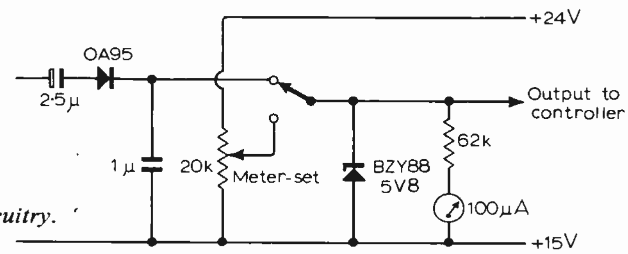

Fig. 7. Radiometer output circuitry.

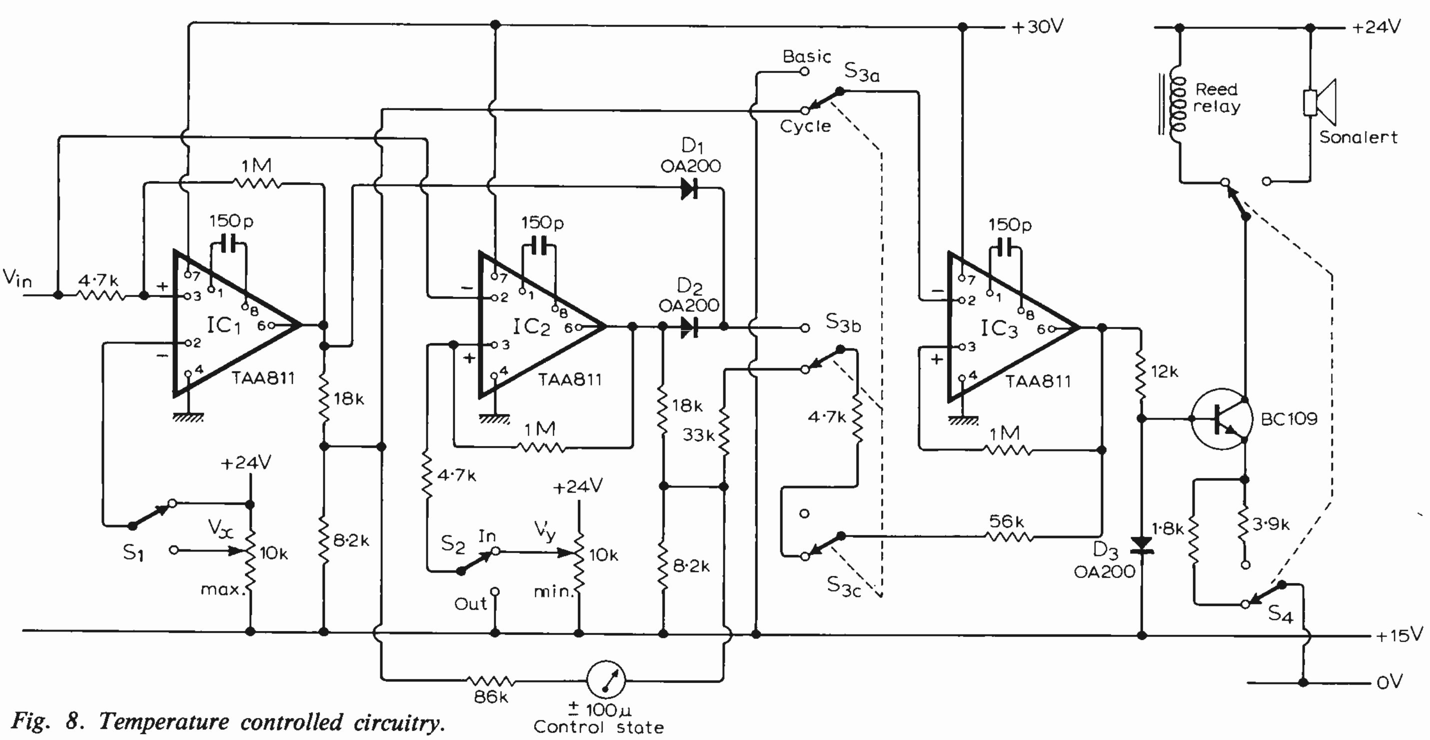

Fig. 8. Temperature controlled circuitry.

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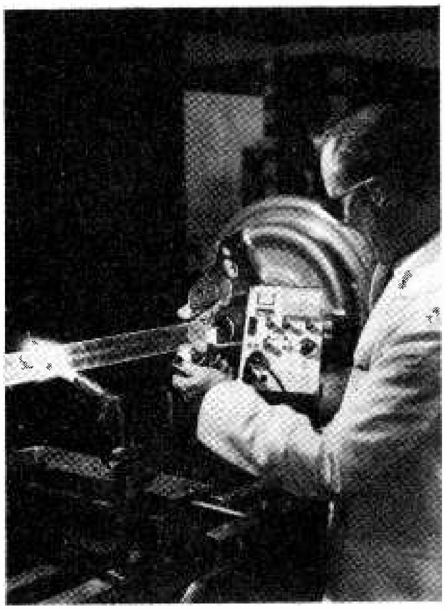

The picture shows the temperature of a silica tube being measured while undergoing a production process. The chopper and temperature sensor form the small unit held in the operator's left hand and pointed at the hot part of the silica tube.

control; the latter considerably eases the accurate setting of limits.

Two integrated circuits (IC_1 and IC_2, Fig. 8) monitor the input voltage, and both their outputs will be at 0V until a limit has been crossed. When S_1 is closed the maximum limit is set by adjusting V_y to a suitable level. If V_{in} exceeds V_x then IC_1 output rapidly switches to +30V. When S_2 is closed, the minimum limit is set by adjusting V_y to a suitable level. If V_{in} falls below V_y then IC_2 output switches rapidly to $+30V$.

With S_3 in the "basic" position, the − input of IC_3 is referred to +15V. The output of IC_3 is normally at 0V. If either limit is crossed, the anode of either D_1 or D_2 will be taken to +30V, and the + input of IC_3 will follow; thus the output of IC_3 will then switch to +30V.

Hysteresis circuit

With S_3 in the "cycle" position, IC_3 forms a hysteresis circuit, and the output voltage will switch to +30V if IC_1 output goes +ve, and will remain there until the output of IC_2 goes +ve, whence it returns to 0V.

When the output of IC_3 switches to +30V, the base of the output transistor rises from 0V to a little above +15V, and either the audible warning or the control relay will operate. The choice is selected by S_4.

A second small centre-zero meter ($\pm 100\mu A$) may be used to give a constant indication of the control state under all conditions. The pointer will be centred until a limit is crossed, and will deflect when a limit is crossed, the direction of deflection being indicative of maximum or minimum.

Olympia Audio Fair Lectures

The 1970 International Audio and Music Fair, to be held from 19th to 24th October at Olympia, London, differs from previous fairs of this kind (in this country) in that there will be a programme of "presentations" sponsored largely by the audio press. The full list is given below.

It will be seen that *Wireless World* is sponsoring five lecture-demonstrations. Each session will be given by a well-known *Wireless World* author, who will relate engineering principles and procedures to the performance actually given by items of audio equipment. The lecturers will be Arthur Bailey, Peter Baxandall, Jack Dinsdale, John Linsley Hood and Ted Jordan. Tickets for each of these sessions will be available at the *Wireless World* stand at the Fair on the day of the lecture.

Tuesday, 20th

2.0 Types of recorded sound quality
by John Crabbe *(Hi-Fi News)*
Different musical and acoustical balances to be found on modern recordings will be discussed and demonstrated.

4.0 The progression of electronic music synthesizers
by Dr. Robert A. Moog
The inventor of the Moog Synthesizer will demonstrate the instrument used for the famous "Switched on Bach" LP.

6.0 The heart of hi-fi
by J. Dinsdale *(Wireless World)*
The important features of modern amplifier circuitry and specifications will be explained and demonstrated.

8.0 Cassettes and cartridges
by W. Woyda (Precision Tapes)
Cassette and 8-track cartridge formats will be described in detail and several recently released recordings will be played in both media.

Wednesday, 21st

2.0 Power levels, distortion and the enjoyment of music
by P. J. Baxandall *(Wireless World)*
The inventor of the Baxandall tone control will discuss the technical requirements of hi-fi equipment and relate them to musical enjoyment.

4.0 How good is your gramophone?
by John Borwick *(The Gramophone)*
The lecturer will play a selection of records and suggest how these can be used to assess the performance of record reproducing equipment.

6.0 Both sides of the record
by Joan Coulson (EMI Records)
All kinds of music on record with glimpses behind the scenes.

8.0 The progression of electronic music synthesizers
by Dr. Robert A. Moog (See Tuesday)

Thursday, 22nd

2.0 From cylinders to 78s to today
by G. Child (formerly of Decca Records)
A short history of the gramophone record with a display of vintage players and examples of records old and new.

4.0 Audio facts and fallacies
by A. R. Bailey *(Wireless World)*
The well-known designer examines hi-fi ideas and terminology.

6.0 Personal appearance of Sir Arthur Bliss
The Master of the Queen's Musick will introduce excerpts from the first recording of his "Pastoral" and "Knot of Riddles" to be issued in October by Pye Records.

8.0 Stereo for beginners
by Clement Brown *(Hi-Fi Sound)*
A simple introduction to the subject of stereophonic recording and reproduction.

Friday, 23rd

2.0 Cassettes and cartridges
by W. Woyda (Precision Tapes)
(see Tuesday)

4.0 How good is your gramophone?
by John Borwick *(The Gramophone)*
(see Wednesday)

6.0 Little and good
by J. L. Linsley Hood *(Wireless World)*
An illustrated examination of the related subjects of power and quality in sound reproduction.

8.0 The funny side
by Donald Aldous *(Hi-Fi News & Record Review)*
A light-hearted look at the record repertoire—"in the groove, but out of the rut".

Saturday, 24th

2.0 Sound sense
by E. J. Jordan *(Wireless World)*
This well-known loudspeaker designer will talk about sound reproduction aims and play some recorded examples.

4.0 The funny side
by Donald Aldous *(Hi-Fi News & Record Review)* (see Friday)

6.0 Types of recorded sound quality
by John Crabbe *(Hi-Fi News)*
(see Tuesday)

8.0 Live pop recital
presented by *New Musical Express*

Throughout the Exhibition there will also be a programme of semi-technical films.

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Active Filters

14. Bandpass types

by F. E. J. Girling* and E. F. Good*

It has already been shown that low-pass filters can be made up from one or more feed-back loops containing integrators, simple lags, or a mixture of the two. The band-pass counterparts of both integrators and simple lags are tuned circuits tuned to the chosen centre frequency. Consequently band-pass filters may be made by substituting such tuned circuits for the lags and integrators of suitable low-pass models, and the method is described for synthesis by factors and for active-ladder synthesis. A brief discussion of stagger tuning is also given.

A symmetrical band-pass characteristic may be considered as a transform of a low-pass characteristic, and all the methods of design used for low-pass filters may be carried over into b-p filter design, the most important being (a) by factors, (b) as a ladder structure. The nature of the transformation has been described in Part 2; and if, for clarity, frequency for the l-p model is designated by the upper-case letter Ω, and for the b-p counterpart by the conventional ω, an algebraic statement of the transformation is

$$\begin{aligned}\Omega &= \omega - \omega_0{}^2/\omega \\ &= \omega_0(\omega/\omega_0 - \omega_0/\omega). \qquad (1)\end{aligned}$$

This produces, Fig. 1, a b-p characteristic in which the bandwidth, $\omega_1 - \omega_2$, between any two points of equal amplitude, and of equal but opposite phase shift, is equal to Ω_1, the bandwidth from zero to the corresponding point on the l-p characteristic. The b-p characteristic is centred on ω_0 in the sense that $\omega_0 = \sqrt{(\omega_1\omega_2)}$, i.e. $\omega_2 = \omega_0{}^2/\omega_1$.

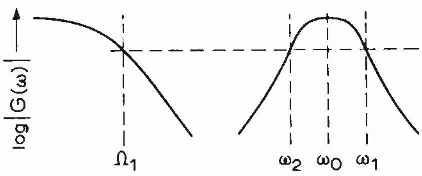

Fig. 1. Corresponding b-p and l-p responses: for any pair of points on the b-p curve and the corresponding point on the l-p curve $\omega_1 - \omega_2 = \Omega_1$ and $\omega_1\omega_2 = \omega_0{}^2$.

Realisation by factors

Since resolution of a l-p characteristic into 1st- and 2nd-order factors provides a basis for the design of a l-p filter as a cascade of non-interacting simpler (i.e. lower-order) l-p sections, transformation of the l-p factors to b-p provides a basis for the design of the corresponding b-p filter as a cascade of non-interacting simpler b-p sections.

1. 1st-order

The frequency response function for 1st-order l-p response (a simple lag)

$$G(\Omega) = 1/(1 + j\Omega/\omega_c) \qquad (2)$$

transforms into

$$G(\omega) = 1/\{1 + j\omega_0(\omega/\omega_0 - \omega_0/\omega)/\omega_c\}, \qquad (3)$$

which may be simplified to

$$G(\omega) = 1/\{1 + jQ(\omega/\omega_0 - \omega_0/\omega)\} \qquad (4)$$

where

$$Q = \omega_0/\omega_c \qquad (5)$$

the well known definition.

In transfer-function form the functions are

$$G(p) = \frac{1}{1+pT} \text{ (l-p)} \qquad (6)$$

$$G(p) = \frac{1}{1+Q\left(pT_0 + \dfrac{1}{pT_0}\right)} \text{ (b-p)} \qquad (7)$$

which rearranges into

$$G(p) = \frac{pT_0/Q}{1 + pT_0/Q + p^2T_0{}^2}. \qquad (8)$$

the familiar form for "tuned-circuit" response. ($T_0 = 1/\omega_0$). The transformation may be considered, therefore, to consist in replacing pT, where $T = 1/\omega_c$, by

$$Q\left(pT_0 + \frac{1}{pT_0}\right) \qquad (9)$$

Physically this means, for example, that a differentiator (say the capacitor of a Blumlein integrator) has an integrator placed in parallel with it.

2. 2nd-order

2nd-order l-p response may be defined

$$G(p) = 1/(1 + pT/q + p^2T^2) \qquad (10)$$

When $q > \frac{1}{2}$ the response cannot be resolved into real 1st-order factors, and when $q > 1/\sqrt{2}$ the response shows a resonant hump or peak. The corresponding b-p response is of the type associated with coupled tuned circuits.

2nd-order l-p response may be obtained from an integrator-and-lag loop, a two-lag loop, etc. The b-p transform of a simple lag is a single tuned circuit of finite Q and tuned to ω_0 (above), and it is a simple piece of reasoning† to show that the transform of an integrator is an infinite-Q tuned circuit, also tuned to ω_0. It follows that 2nd-order b-p responses may be obtained from feedback loops containing two tuned circuits tuned to the required centre frequency.

Tuned circuits of finite Q are matched to the lags of the l-p model so that the same bandwidths are obtained, and the gain at ω_0 (the peak) is made equal to the gain of the l-p simple-lag circuits at zero frequency. Tuned circuits of infinite Q are matched to corresponding integrators in the l-p model by making the unity-gain bandwidth the same, Fig. 2.

In a b-p filter there is the possibility of a type of error which has no counterpart in the l-p model. There can clearly be no error in the position of zero frequency, but the centre frequencies of the constituent tuned circuits of a b-p filter can be misaligned both with respect to each other and with respect to the desired ω_0. In a loop containing an infinite-Q tuned circuit and a damped tuned circuit misalignment causes the loop gains in the critical regions near the band limits (where the Nyquist plot comes nearest to the $-1, j0$ point, Fig. 3) to be different, with more phase shift on one side than on the other for points of equal magnitude of loop gain. This causes the response when the loop is closed to be unsymmetrical, showing more peaking on one side of the passband than on the other (i.e. show different equivalent values of q). Two tuned circuits of equal Q give a symmetrical response even when misaligned, Fig. 4, and consequently a symmetrical response with the feedback

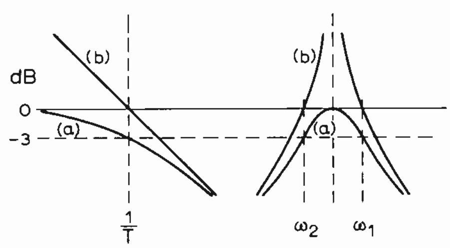

Fig. 2. Relationship between (a) a damped tuned circuit and a simple lag, (b) an infinite-Q tuned circuit and an integrator. For each pair $\omega_1 - \omega_2 = 1/T$.

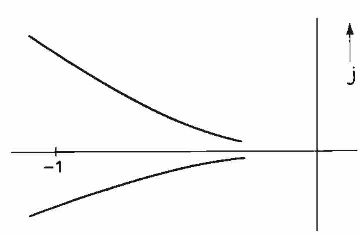

Fig. 3. Misalignment of tuned circuits of different Q gives an unsymmetrical loop response, which shows in the Nyquist plot.

†The integrator has infinite gain at zero frequency and a constant 90° phase lag: the infinite-Q tuned circuit has infinite gain at ω_0, a constant 90° phase lag for $\omega > \omega_0$, and a constant 90° phase lead for $\omega < \omega_0$.

* Royal Radar Establishment.

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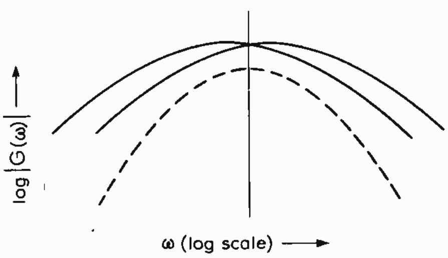
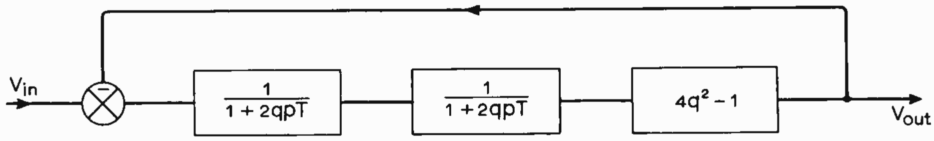
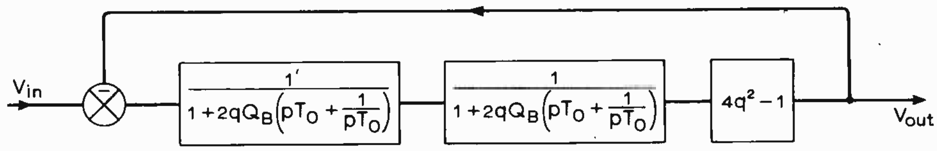


Fig. 4. Misalignment of two tuned circuits of equal Q gives a still symmetrical loop response.



(a) $$\frac{V_{out}}{V_{in}} = \frac{4q^2-1}{4q^2}\cdot\frac{1}{1+\frac{1}{q}pT+p^2T^2}$$



(b) $$\frac{V_{out}}{V_{in}} = \frac{4q^2-1}{4q^2}\cdot\frac{1}{1+\frac{1}{q}Q_B\left(pT_0+\frac{1}{pT_0}\right)+Q_B^2\left(pT_0+\frac{1}{pT_0}\right)^2}$$

Fig. 5. A two-lag loop and its band-pass counterpart.

loop closed—although bandwidth and shape factor (q) are affected. In practice there may be errors in Q as well as in centre frequency, but the effect of error is still likely to be more tolerable than in a loop containing one infinite-Q circuit. It appears, therefore, that the preferable l-p model for a b-p filter is the loop with two equal lags.

The properties of such a loop have been considered in Part 4, and the important result obtained that closing the loop increases both q and bandwidth by the factor $\sqrt{(A_0+1)}$, where A_0 is the loop gain at zero frequency. For two equal lags the open-loop q is $\frac{1}{2}$. To obtain a closed-loop value of q, therefore,

$$\sqrt{(A_0+1)} = 2q \qquad (11)$$

i.e. $$A_0 = 4q^2-1; \qquad (12)$$

and for a closed-loop bandwidth $\omega_c = 1/T$, the open-loop bandwidth, which is also the bandwidth of the individual lags, must be $\omega_c/2q$, i.e. the time constant of the lags must be $2qT$. This is shown in Fig. 5(a).

Now the bandwidth of a tuned circuit $= \omega_0/Q$. So for the tuned circuits of Fig. 5(b) to have the same bandwidth as the lags in Fig. 5(a)

$$\frac{\omega_0}{Q} = \frac{\omega_c}{2q}$$

i.e. $$Q = 2q\omega_0/\omega_c. \qquad (13)$$

It is now convenient to define a measure of the relative sharpness or selectivity of a 2nd-order band-pass circuit,

$$Q_B = \omega_0/\omega_c. \qquad (14)$$

This is comparable with the definition of Q for a single tuned circuit, although now ω_c is the nominal bandwith equal to the corner frequency of the corresponding 2nd-order l-p circuit, and the −3dB bandwidth only for the particular case of $q = 1/\sqrt{2}$.

Making this substitution in equn. (14) gives

$$Q = 2qQ_B \qquad (15)$$

and substituting for Q in equn. (7) then gives for the transfer functions of the elementary tuned circuits of Fig. 5(b)

$$G(p) = \frac{1}{1+2qQ_B\left(pT_0+\frac{1}{pT_0}\right)} \qquad (16)$$

Comparison with the transfer functions of the simple lags of Fig. 5(a)

$$G(p) = \frac{1}{1+2qpT} \qquad (17)$$

shows that the transformation may be considered to be the replacement of pT, where $T = 1/\omega_c$, by

$$Q_B\left(pT_0+\frac{1}{pT_0}\right). \qquad (18)$$

Transfer functions for corresponding l-p and b-p 2nd-order sections may therefore be written

$$\frac{1}{1+\frac{1}{q}pT+p^2T^2} \text{ (l-p)} \qquad (19)$$

$$\frac{1}{1+\frac{1}{q}Q_B\left(pT_0+\frac{1}{pT_0}\right)+Q_B^2\left(pT_0+\frac{1}{pT_0}\right)^2} \text{ (b-p)} \qquad (20)$$

though the method of synthesis just described does not depend on explicit statement of the b-p transfer function.

The rule for l-p to b-p transformation given above, expression (18), is only apparently different from the rule given for 1st-order sections, expression (9). By dividing by $T(=1/\omega_c)$ both reduce to the basic rule: i.e. p is replaced by

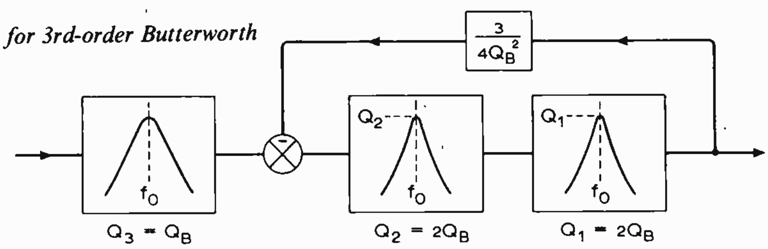
$$\frac{1}{T_0}\left(pT_0+\frac{1}{pT_0}\right) \qquad (21)$$

Equation (15) and the well known relationship expressed by equn. (5) provide the basic theory for the design of any symmetrical simple (all-pole) band-pass filter as a cascade of active 1st- and 2nd-order band-pass sections.

It is also worth noticing for future reference that because of the identity

$$\frac{pT_1}{D_1(p)}\times\frac{pT_2}{D_2(p)} = \frac{T_1}{T_2}\times\frac{1}{D_1(p)}\times\frac{p^2T_2^2}{D_2(p)} \qquad (22)$$

Fig. 6. Scheme for 3rd-order Butterworth filter.



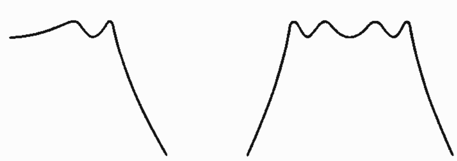


Fig. 7. The band-pass response has twice as many ripples as the low-pass response used as model.

where in the present case

$$D_1(p) = 1+pT_1/Q_1+p^2T_1^2 \qquad (23)$$

$$D_2(p) = 1+pT_2/Q_2+p^2T_2^2 \qquad (24)$$

two tuned circuits in tandem may be replaced by a low-pass circuit and a high-pass circuit also in tandem.

3. Higher-order filters

The general pattern for a higher-order filter realized as a cascade of sections designed as above is n tuned circuits all tuned to ω_0 and connected in $n/2$ feedback pairs, or when n is odd $(n-1)/2$ feedback pairs with one single tuned circuit. The feedback pairs form the 2nd-order sections, and the Q factors of the constituent tuned circuits and the degree of back coupling are chosen to give the required closed-loop bandwidth and q (shape factor). Similarly the Qs of any single tuned circuits (the 1st-order sections) are chosen to give the required bandwidth.

Butterworth filters are a special case, as the nominal bandwidth of each section is equal to the −3dB bandwidth of the complete filter. So $Q_{B1} = Q_{B2} =$ etc. $= \omega_0/\omega_c$; and for n odd the Q factor of the single tuned circuit is also equal to ω_0/ω_c. But the q of each 2nd-order section is different, and, using equns (15) and (12), the Q factors of

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the constituent tuned circuits and the loop gains must be calculated as follows:

$$Q_1 = 2q_1Q_B, \text{ loop gain } = 4q_1^2-1,$$
$$Q_2 = 2q_2Q_B, \text{ loop gain } = 4q_2^2-1, \text{ etc.}$$

Thus the block diagram for a 3rd-order Butterworth filter, which needs only one feedback pair (with $q = 1$), may be drawn as in Fig. 6. Values of q for up to $n = 10$ were given in Part 9, Table 6.

In general, however, both q and bandwidth will be different for each factor, as they are for the factors of an l-p filter, e.g. Table 8 of Part 9. The parameters q pass straight over into the band-pass design. The parameters T are inversely proportional to the nominal bandwidths of the factors (ω_c), and consequently directly proportional to the band-pass parameters Q_B (or Q for a 1st-order factor). As a b-p response repeats itself on the other side of ω_0 it will show twice as many ripples as the l-p response, e.g. Fig. 7.

1/*n*th octave filter

The bandwidth of a b-p filter is often specified as a fraction of an octave (2:1 frequency ratio), especially when it is one of a bank of filters required to divide a broad spectrum into a number of relatively narrow contiguous bands.

If the filters all have the same relative bandwidth, and the first one is centred on $\omega_0 = 1$, the nth filter will be centred on $\omega_0 = 2$, Fig. 8. Consequently

$$x^2 = 2^{\frac{1}{n}}, \qquad (25)$$

$$\frac{1}{Q_B} = x - \frac{1}{x} = \frac{2^{\frac{1}{n}}-1}{2^{\frac{1}{2n}}}, \qquad (26)$$

from which the figures shown in Table 1 may be calculated. The last row corresponds to one filter for each note of the equally tempered scale. Sometimes of course, the relative bandwidth is made different from the spacing between adjacent filters. This is so for B.S.⅓rd-octave filters, which have a band width much less than the pitch between centre frequencies.

From the first column of Table 1 the Q factors required of individual tuned circuits may be calculated via equn. (15). Thus for 3rd-order Butterworth response Q factors

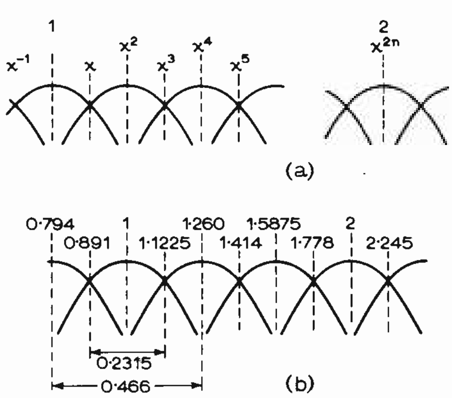

Fig. 8. (a) n filters to the octave, each with a band width of 1/nth of an octave. (b) Frequency relationships for 1/3rd-octave filters.

TABLE 1

1/*n*th-octave filters—Q_B against *n*

n	1	2	3	4	5	6	8	10	12
Q_B	1·414	2·87	4·32	5·76	7·21	8·65	11·55	14·47	17·3

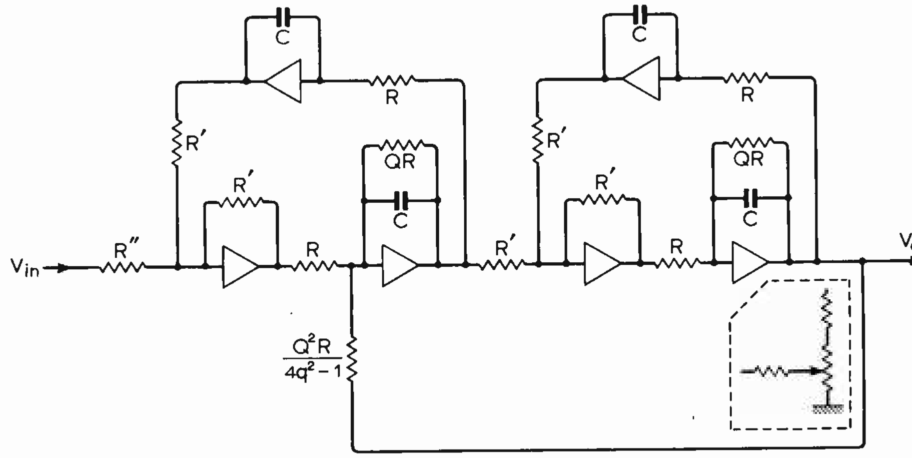

Fig. 9. 2nd-order band-pass filter using two-integrator loops.

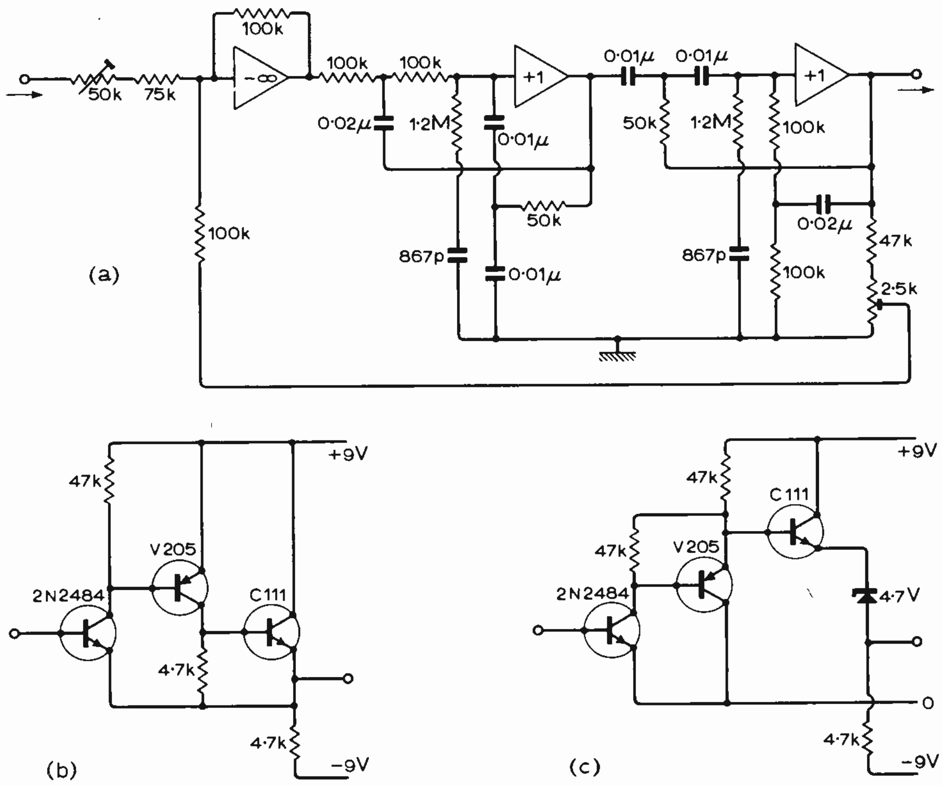

Fig. 10. (a) Schematic for ⅓rd-octave 2nd-order band-pass filter, centre frequency 159 Hz, peak gain 18 approx. (b) Enhanced emitter follower using three transistors. (c) Simple high-gain amplifier.

equal to $2Q_B$ are needed. This gives a quide to the type of circuit needed.

Circuits

The most powerful type of active tuned circuit is the two-integrator loop, especially when high Q is needed. In its ordinary form (Parts 7 and 8) the circuit uses three amplifiers, and two circuits may be put together as in Fig. 9 to form a feedback 2nd-order band-pass section. With the usual balance of component values shown each tuned circuit gives a peak gain of Q, not 1 as assumed in Fig. 5 for example. So the "gain" of the feedback path must be divided by Q^2 to compensate, and unless Q is very low the feedback resistor may have an inconveniently high value. This can be avoided by using a resistor of moderate value fed from a potential divider as shown inset. It may also be helpful to make the feedback variable over a small range. This will allow some adjustment of bandwidth, although not without varying q simultaneously.

The ease with which both tuning and Q can be adjusted in a two-integrator loop can also be made use of. Thus each tuned circuit can be individually adjusted to the centre frequency, allowing components of relatively broad tolerance to be used, or perhaps the nearest preferred value. Similarly

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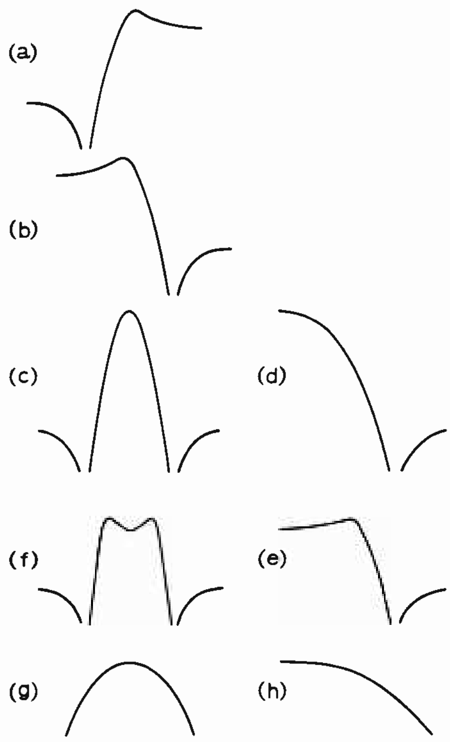

Fig. 11. Building up a 3rd-order b-p response. On the right, corresponding l-p responses.

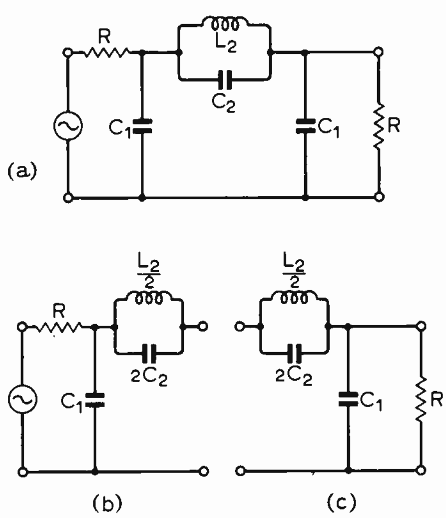

Fig. 12. Bisection of a symmetrical filter.

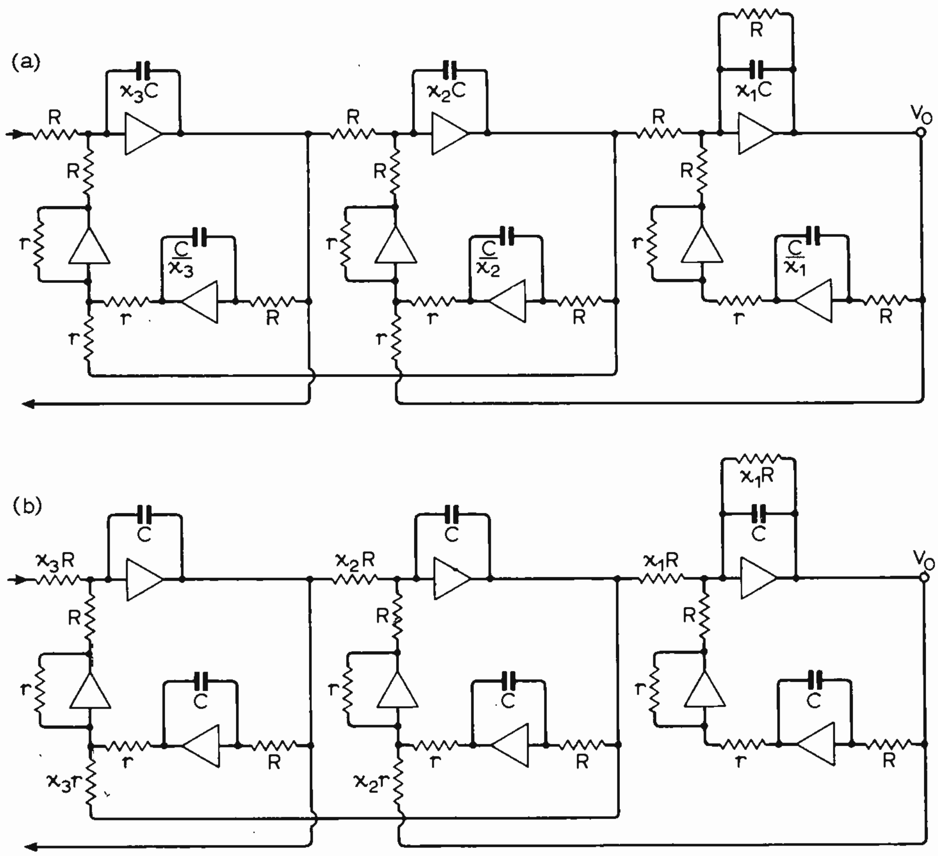

Fig. 13. As first derived the integrators of the elementary tuned circuits have different Ts (CR products). This can be corrected by making the changes shown in (b).

in a 3rd-order filter tilt in the passband can be corrected by adjusting the tuning of the single tuned circuit. If only a single resistor (or T) is varied, Q is varied with ω_0, but for a small range of adjustment this is likely to be unimportant. In any case it is quite easy to adjust Q also.

Probably most active band-pass filters at the present time use a balanced parallel-tee network (see Parts 10 and 11). Fig. 10 shows a circuit for a 2nd-order filter which was designed for $\frac{1}{3}$rd-octave bandwidth approximately. The tuned stages use the Sallen-and-Key type of connection and are arranged as low-pass × high-pass, which also usefully multiplies the forward gain by Q^2. An untuned input stage puts the necessary sign reversal into the loop, and allows adjustment of overall gain without affecting the tuned stages. Adjustment of the feedback is also provided, but if the required combination of bandwidth and q is not obtained the damping of one or both stages must be adjusted.

As the bandwidth is over 20% of the centre frequency, a reasonable tolerance on the time constant of the tees is ±1%. To ensure that each twin tee is well balanced, however, so that the expected value of Q is obtained, it is advisable to make each up from matched pairs of components as explained in Part 11 (June). As a b-p system is not required to pass zero frequency, the amplifiers do not need to have low zero drift; and as Q is only moderate, the internal gain need not be very high. The amplifiers can, therefore, be quite simple—though they can be high-gain types provided they will give the required output at the highest frequency with low distortion.

The attenuation of a 2nd-order Butterworth filter is given by $(1+x^4)^{\frac{1}{2}}$, where x = actual bandwidth/-3dB bandwidth. So 20 dB is obtained at $x = 99^{\frac{1}{4}} = 3{\cdot}155$; 40 dB at $x = 10$ approximately.

An advantage of the parallel-tee circuit is that zeros are easily added. Thus a 3rd-order b-p filter can be built up as in Fig. 11. The 2nd-order loop contains a l-p and a h-p stage, and these are given zeros (a) and (b) by feeding a fraction of the input to the other tee (Part 11, Fig. 11) so that an open-loop response as (c) is obtained. If this is symmetrical it can be represented by the transfer function of the corresponding l-p response (d) of the form

$$\frac{1+k^2p^2T^2}{1+pT/2+p^2T^2},$$

and the amount of feedback calculated which will give a l-p response such as (e) and to the b-p circuit an "over-coupled" response (f). The hole in the middle can then be "filled up" with a single tuned circuit (g) corresponding to a simple lag (h).

A suitable model is a 3rd-order Darlington filter, Fig. 12(a). Being a symmetrical network it can be divided into two halves with $Z_{out(1)} = Z_{in(2)}$, and its transfer function may be written down as $\frac{1}{2}$ the product of the transfer functions of the subnetworks so formed, Figs 12(b) and (c). These are:

$$\frac{1}{1+pC_1R}$$

$$\frac{1+p^2L_2C_4}{1+pL_2/2R+p^2L_2(C_1+C_4)/2}$$

and can act directly as l-p models for the single tuned circuit and the 2nd-order loop of the active b-p filter, the required bandwidths and q being obtained by comparing these transfer functions with transfer functions written in standard form, $1/(1+pT_1)$ and $(1+ap^2T_2^2)/(1+pT_2/q+p^2T_2^2)$.

Parallel-tee stages can be given variable tuning (Part 11); but this usually requires extra amplifiers, and the short-circuited-output time constants of the two tees must still be equal. So the practicability of a parallel-tee filter, or bank of filters, depends very much on the cost and difficulty of obtaining accurate components for the parallel-tee networks. As the price of gain comes down, therefore, the preference is likely to go to circuits which, while needing more amplifiers, have less severe requirements for passive components.

Active ladders

The first stage in preparing a design for a band-pass active ladder filter has already

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been given (Part 13, Fig. 17). But when the centre frequency is much greater than the bandwidth ($\omega_0 \gg \omega_c$), L_1/R, C_2R, L_3/R of diagram (b), where L_1, C_2, L_3, come from the l-p filter of equal bandwidth, are much greater than $C_1'R$, L_2'/R, $C_3'R$, where C_1' etc. are the reactances added to transform the l-p filter to b-p. Thus, for example, for 5th-order Butterworth response with $\omega_c = 20$ Hz, $\omega_0 = 225$ Hz,* $T_1 = T_5 = 4{\cdot}92$ and $T_1' = T_5' = 0{\cdot}1017$, $T_2 = T_4 = 12{\cdot}88$ and $T_2' = T_4' = 0{\cdot}03885$, $T_3 = 15{\cdot}92$ and $T_3' = 0{\cdot}03143$ (all in milliseconds). This is undesirable. As $T_0 = 1/\omega_0 = 0{\cdot}707$ ms, the integrators of low T would work with voltage gains of up to twenty and more. So they would develop output voltages up to that number of times greater than the voltages at the outputs of the integrators feeding them, which is bad for dynamic range; and the elementary tuned circuits which they form with those integrators would be likely to have a lower Q factor than need be, since if the zero-frequency gains of the integrators are equal the best performance is obtained with Ts equal.

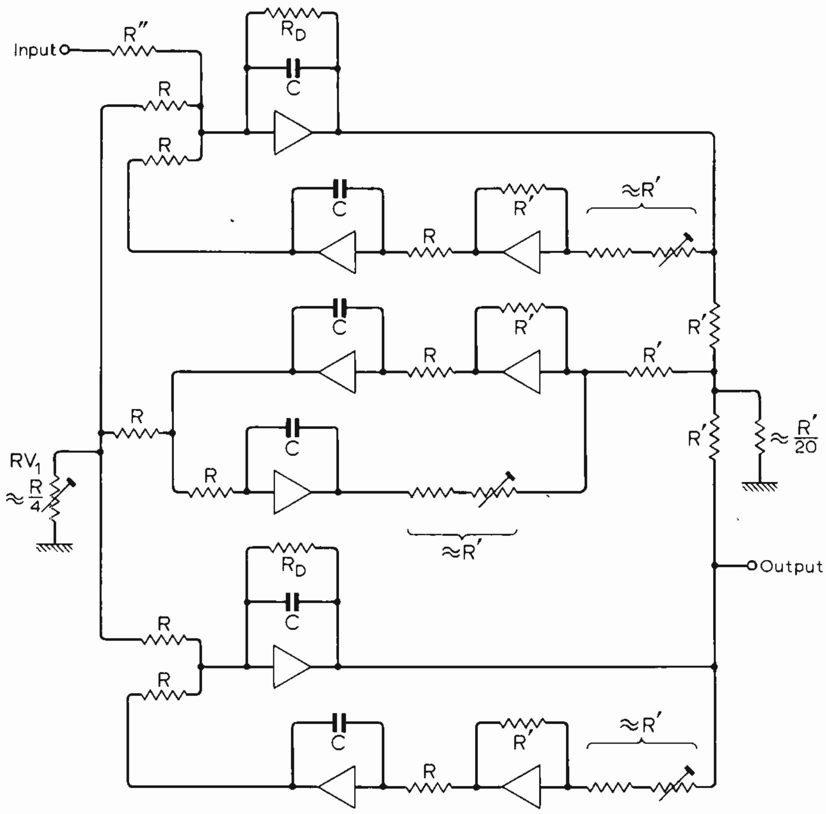
This condition is obtained by making the T of each integrator $= T_0$, and readjusting the linkages between the tuned circuits so that all loops have the same loop gains as before. If the T of one of the integrators in the forward path is changed from T_1 (say) to T_0, the signal voltage at its output is multiplied by T_1/T_0. The loop gain of the minor loop, the elementary tuned circuit, has simultaneously been restored to its former value by changing the T of the other integrator from T_1' to T_0—and $T_1T_1' = T_0^2 = 1/\omega_0^2$. Compensation in all other paths including the T_1 integrator is effected by introducing a multiplying factor T_0/T_1, which may in practice mean multiplying one or more resistances by T_1/T_0, Fig. 13.

The ratios T_0/T_1, T_0/T_2 etc., although not in general equal, are all roughly equal to bandwidth/centre frequency, say $1/Q_B$; and the ratios T_0/T_1', T_0/T_2' etc., roughly equal to Q_B. And, if more convenient, all the Ts of the first set of integrators may be multiplied by $1/Q_B$, and all of the second set by Q_B, as this will bring them near enough together for the elementary tuned circuits to give good performance. The scaling factors to be introduced into the linkages between the tuned circuits will now all be $1/Q_B$. It will still be practicable to use the same value of C for all the integrators, by obtaining the required spread of T by variation of R.

After scaling, a filter with a relatively narrow pass band ($Q_B \gg 1$) is seen to consist of n two-integrator loops, where n is the order of the l-p filter used as model, loosely coupled together. This points to a rough analogy between a coupled-tuned-circuit filter and a ladder b-p filter, in which the loose coupling needed for a relatively narrow bandwidth is obtained by making the reactances of the shunt branches much smaller than those of the series branches.

The analogy may be put to use in a 3rd-order b-p filter as shown in Fig. 14. RV_1 is placed so that the symmetry of the system is preserved and allows simultaneous adjustment of the loop gains of the two inter-

* 1 Hz = 2π radians/second.



C = 0·01μ, R = R' – R'' = 100k, R$_D$ = 16R

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Fig. 14. 3rd-order b-p filter with adjustable coupling.

meshing feedback loops. As it is increased from a low value the response changes from single-peaked, through a maximally flat response (Butterworth), to the overcoupled type of response, Fig. 15. Ideally the three peaks have equal height. This depends on the central tuned circuit having infinite Q. If it has appreciable damping the outer peaks are lower than the middle peak and may become only shoulders. Compensation can be made by applying to the central stage regeneration or positive feedback.

Other types of tuned circuit can be used in place of the two-integrator loops, provided the inside ones have high Q (approaching infinity). Thus Fig. 16 shows a 3rd-order filter using parallel-tee circuits. By making use of the identity expressed in equn. (22), and by making the middle stage sign-inverting and the outer stages not, the feedback loops have been closed in the right sense, without the use of separate inverting or adding stages. So the filter is economical in amplifiers, using only one third as many as the previous circuit. But although the equally terminated structure is some help against the effects of mistuning, the Q factors of the tuned circuits are still very dependent on accurate balance in the twin-tee networks, and at least twice as many capacitors are needed.

Electronic tuning

An all-integrator circuit lends itself to electronic (voltage-controlled) tuning (Part 8, Fig. 11). The electronic switches are equivalent to accurately ganged potentiometers, and the effective T of each integrator in series with a switch varies inversely with the mark/period ratio.

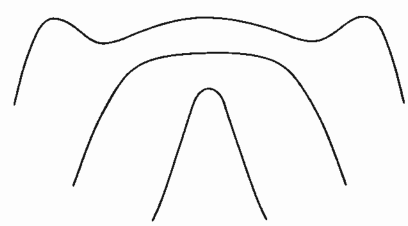

Fig. 15. Showing effect of increasing coupling in a 3rd-order b-p ladder filter.

In the passive band-pass model (Fig. 17(a) of Part 13) centre frequency is determined by the LC product of the elementary tuned circuits, $L_1C_1' = C_2L_2' = $ etc. $= 1/\omega_0^2$. The bandwidth is given by only L_1/R, C_2R, L_3/R, etc., L_1, C_2, L_3, and R, being the components derived from the l-p counterpart. If all reactances are multiplied by x, both $1/\omega_0$ and 1/bandwidth are multiplied by x; so the frequency response curve is moved down the frequency scale and keeps the same ratio of bandwidth to centre frequency. By analogy with a single tuned circuit this may be called constant-Q tuning. Because of the linear relationship between x and $1/\omega_0$, it is quite easy in the active ladder to reduce ω_0 say ten times. But the tuning reduces the effective A (zero-frequency gain) of each integrator; so as the constant-Q nature of the tuning means that ideally the same Q factors are called for, some deterioration of the sharpness of the corners of the passband may be experienced.

If only the primed reactances are varied tuning is obtained with constant bandwidth. The centre frequency, $\omega_0 \propto 1/\sqrt{x}$; and, because of the constant bandwidth, the Q factors called for are proportional to ω_0 and therefore fall as ω_0 is reduced. It is likely,

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therefore, that the tuning range found practicable will be smaller, and that in compensation deterioration of bandshape at the lowest frequency will also be less.

Once again we are indebted to Dr R. L. Ford for showing that the versatility of the active filter does not end here and that it is possible to vary bandwidth independently of ω_0. To do this it must be possible to increase the effective value of L_1, C_2, L_3, etc., for transmission round the larger loops, while leaving the tuning of the minor loops unchanged. The switches are therefore placed as shown in Fig. 17. Operation of switches k_2 varies the bandwidth, operation of switches k_1 varies the centre frequency, operation of both sets simultaneously gives the constant-Q type of tuning. If zeros are introduced by adding linkages as described in Part 13, these also move under the action of the switches, as their positions also depend on the effective Ts of the integrators.

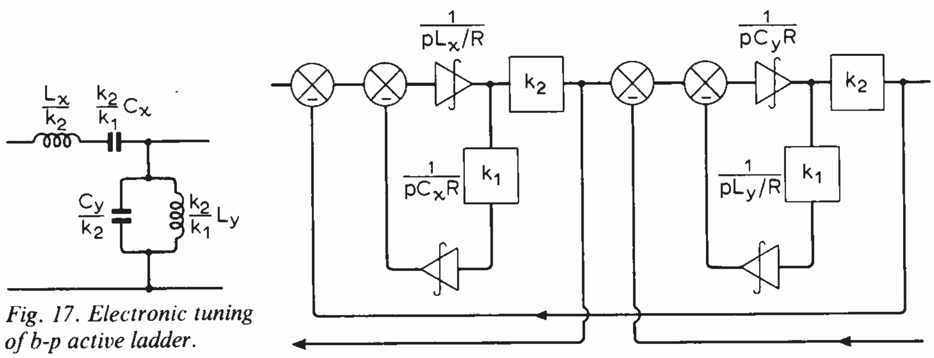


Fig. 17. Electronic tuning of b-p active ladder.

Stagger tuning

Consideration of symmetry shows that when two tuned circuits are staggered to give a 2nd-order b-p response they must be of equal Q, and that the centre frequency of the combined response is the geometric mean of the centre frequencies of the two factors. Hence the pole positions are as shown in Fig. 18, and the normalized responses may be written

$$G_1(\omega) = \frac{1}{1+jQ(\omega/x-x/\omega)} \qquad (27)$$

$$G_1(\omega) = \frac{1}{1+jQ(\omega x-1/\omega x)} \qquad (28)$$

The product gives an expression for the overall gain, and from this by substituting $\Omega = \omega - 1/\omega$, which is equn. (1) when $\omega_0 = 1$, the frequency-response function for the corresponding l-p response is obtained as

$$G(\Omega) = \frac{1}{1+Q^2\left(x-\frac{1}{x}\right)^2+jQ\left(x+\frac{1}{x}\right)\Omega-Q^2\Omega^2}. \qquad (29)$$

If S is defined by

$$x-1/x = 1/S, \qquad (30)$$

$$x+1/x = 2\sqrt{(1+1/4S^2)} \qquad (31)$$

So equn. (29) becomes

$$G(\Omega) = \frac{1}{1+Q^2/S^2+j2Q\sqrt{1+1/4S^2}.\Omega-Q^2\Omega^2} \qquad (32)$$

and comparison with

$$G(\Omega) = \frac{1}{1+j\Omega/q\omega_c-\Omega^2/\omega_c{}^2} \qquad (33)$$

gives

$$q = \frac{1}{2}\sqrt{\frac{1+Q^2/S^2}{1+1/4S^2}} \qquad (34)$$

$$\omega_c = \frac{1}{Q}\sqrt{1+\frac{Q^2}{S^2}} \qquad (35)$$

So, for example, if $S = Q$, the bandwidth obtained is $\sqrt{2}/Q$, i.e. $\sqrt{2}$ times the bandwidth that the tuned circuits would have if not staggered.

For the general case where the overall response is centred on ω_0

$$\omega_c = \frac{\omega_0}{Q}\sqrt{1+Q^2/S^2} \qquad (36)$$

Now ω_0/ω_c is the parameter Q_B. For brevity let $B = Q_B$. Equn. (36) may then be written

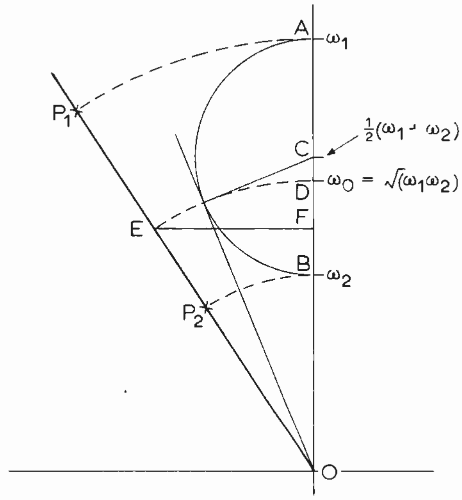


Fig. 18. Pole positions for 2nd-order band-pass response.

$$\frac{1}{B} = \frac{1}{Q}\sqrt{1+Q^2/S^2} \qquad (37)$$

or

$$\frac{1}{B^2} = \frac{1}{Q^2}+\frac{1}{S^2}. \qquad (38)$$

Thus q, the shape factor, and $1/B$, the relative bandwidth, may be calculated for any combination of Q and S.

For relatively narrow bandwidths $1/4S^2 \ll 1$, so equn. (34) becomes

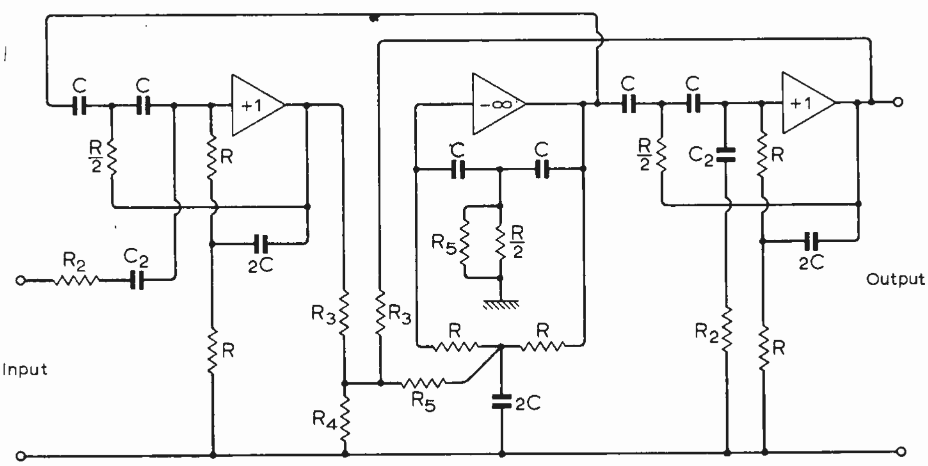
$$q = \frac{1}{2}\sqrt{1+Q^2/S^2} \qquad (39)$$

and substitution in equn. (37) gives

$$Q = 2qB. \qquad (40)$$

The above results may be used to analyse the performance of a h-p filter and a l-p filter in tandem, using the identity already given, equn. (22). Thus equn. (34) shows that for very great staggering ($4S^2 \ll 1$ and $S^2 \ll Q^2$) $q \rightarrow Q$; but that for $Q > \frac{1}{2}$, as S increases q falls, until for no stagger $q = \frac{1}{2}$.

Stagger tuning is not a good practical method of making a filter, as obtaining a flat pass band depends so much on balancing one slope against another (e.g. see Ref. 1, Fig. 10.16). But the analysis is of theoretical interest as it gives the resonant frequencies and Q factors of a b-p system.



C = 0·05μ, R = 50k, C_2 = 1,250p, R_2 = 2·0M, R_3 = 27k, R_4 = 1·2k, R_5 = 470k

Amplifiers type 741

Fig. 16. 3rd-order b-p active ladder filter using parallel-T circuits.

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Current Generators

Some circuit techniques used to design discrete component and integrated circuit current generators

*by B. L. Hart,*B.Sc., M.I.E.R.E.*

Good approximations to voltage sources are commonplace in electronic circuits and are widely used in equipment design. Familiar examples are the low impedance d.c. power supply and the "follower" circuit in all its guises (cathode, emitter, etc.).

Less well known, but often very useful, are current source circuits or current generators. These are required in the measurement of semiconductor d.c. characteristics when a specified current is caused to flow between the terminals of a device and a breakdown voltage observed. Also a current source is required for the linear charging of a capacitor to produce an accurate linear sawtooth voltage waveform in timing applications. Such a scheme has the advantage over a conventional "Miller" circuit of producing no initial jump and not requiring a floating clamp.

Review of fundamentals

The first quadrant characteristics of an ideal controllable current generator are evident from Fig. 1(a); these are:

(i) Infinite l.f. incremental resistance for $V \geqslant 0$.

(ii) The d.c. current is dependent on only one chosen variable λ. This parameter could represent the effect of variation in a circuit resistance, current, voltage, charge, applied pressure, or radiation intensity, etc. The practical situation corresponding to Fig. 1(a) is shown, exaggerated for clarity, in Fig. 1(b). Curve (1) is now only approximately straight over a region between two arbitrarily defined points A and B being limited beyond these points by physical mechanisms such as voltage saturation and breakdown. The incremental output resistance at point P, $r_0[= (\delta I/\delta V)|_\lambda]$ is not infinite and may be a function of λ and V, depending on the precise nature and spacing of the curves. Furthermore, λ may not represent only the effect of one variable. It is desirable that the variable at our disposal, e.g. resistance, is dominant, and that there is a relative insensitivity, calculable in magnitude, of set current I, and resistance r_0 with respect to the other variable parameters. In most cases of practical interest the disturbing functions are temperature and rail voltage variation.

Before passing to specific circuit realizations and assessing to what extent they succeed in satisfying the ideal criteria discussed above, it is worth noting that current generators may be simply paralleled. One easily-met requirement is that the common load resistance is much less than the r_0 of each source.

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Basic circuit schemes

One of the simplest and most frequently encountered current sources is that shown in Fig. 2(a), in which D_2 is a zener diode (preferably temperature-compensated) operating in the breakdown region.

Normally $V_Z \gg \delta_1, \delta_2$, and $\delta_1 \approx \delta_2$: hence if α = common-base current gain of Tr_1, then for $BV_{CBO} > V \geqslant (V_z + \delta_2)$,

$$I \doteqdot \alpha V_Z/(R_E + R_V) \qquad (1)$$

$$r_o \doteqdot 1/h_{ob} = f(I) \qquad (2)$$

Potentiometer R_V is the controlling variable. This may be manually operated or mechanically driven, or might represent an f.e.t. operated in the pre-pinch-off region. The main disturbing influences are variations of temperature (T), and change in V_1. The latter produces a change in V_Z because of alteration in zener current; this may be reduced by replacing R_B by a similar current source using a p-n-p transistor. Since α and V_Z are weak functions of T, I is not sensibly dependent on this variable. Sometimes the lower limit of the voltage range is unacceptably high.

If a negative rail is available then the simple arrangement of Fig. 2(b) is useful. The stability of I is directly dependent on $-V_1$ for this configuration.

For the case where a negative rail voltage is not available the circuit of Fig. 3(a) may be used. This has been used by the author

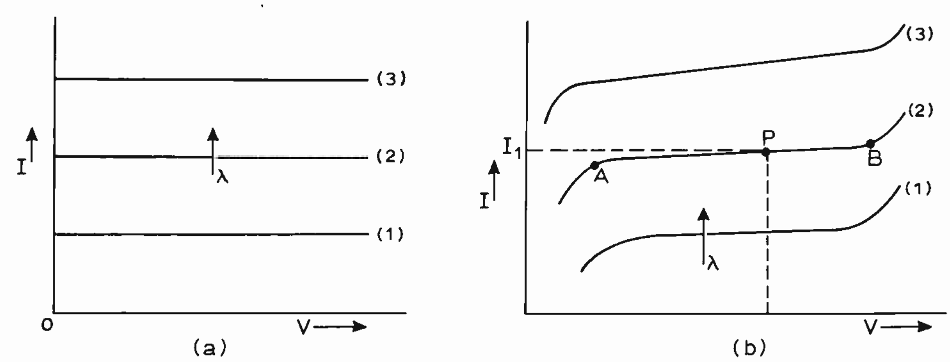

Fig. 1. (a) First quadrant characteristics of ideal current source. (b) Characteristics of practical current source.

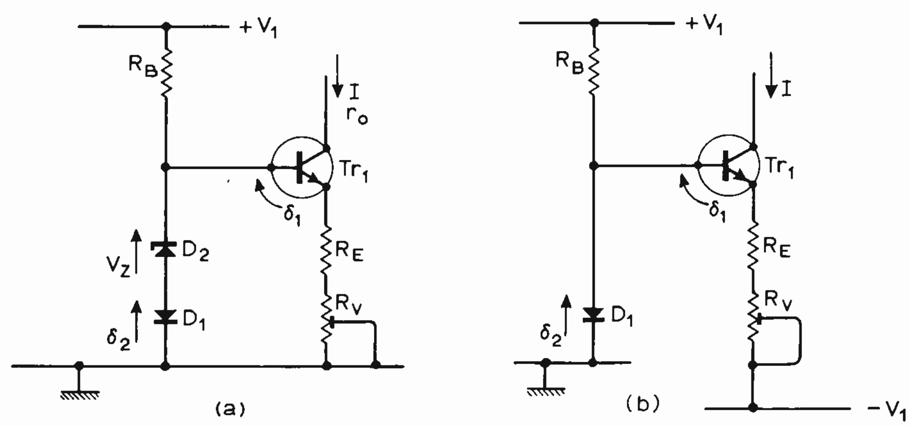

Fig. 2 (a) Popular current source circuit. (b) Current source using two rail supplies.

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to obtain the maximum input voltage range in a long-tailed pair differential comparator.

Tr_1 is a germanium transistor whilst D_1 is a silicon diode.

It follows that, for $V_u > V \geqslant \delta_2$, where V_u in the upper limit of the voltage range,

$$I = \alpha(\delta_2 - \delta_1)/(R_E + R_V) \quad (3)$$

$$1/h_{oe} < r_o < 1/h_{ob} \quad (4)$$

V_u, and the actual value of r_o both depend on $(R_E + R_V)$ and the incremental resistance of D_1.

Assuming a logarithmic volt-amp. relationship for D_1, a $\times 4$ change in V_1 causes a change in I of about 10% for $\delta_2 \approx 0{\cdot}6$ V and $\delta_1 \approx 0{\cdot}2$ V.

Because of the larger value of I_{CBO} for germanium transistors and the fact that $|d\delta_2/dT| \neq |d\delta_1/dT|$ there is an increased temperature sensitivity for this configuration compared with that of Fig. 2(a).

To obtain the output characteristics, shown in Fig. 3(b), on a curve tracer (Tektronix 575) R_B is removed and the rest of the circuit treated as a transistor—the emitter of which is the junction of D_1 and R_V—subject to base current drive in the common-emitter configuration. The logarithmic relationship between applied current and δ_2 causes two effects which are visible on the photograph.

In the first place the curves become almost parallel to the voltage axis at points successively shifted to the right by a small amount. This occurs because the saturation condition $V_{CB} = 0$ is a function of drive current.

Secondly, the characteristics are unequally spaced in the vertical direction.

In a practical circuit a selected value of R_B corresponds to a particular base current step. The vertical spacing thus gives a measure of the effect of R_B tolerance on circuit performance. For predictable behaviour it is arranged that $(V_1/R_B) \geqslant (I/\beta)$ where β = common emitter d.c. current gain of Tr_1. The results were obtained using an OC139 for Tr_1 and a low cost 1N4148 for D_1. R_V was set at a convenient value: in cases where R_V is dispensed with, a closer tolerance on a fixed I may be obtained by using a well-specified diode, such as the 1N3595, for D_1.

If the magnitude of the minimum voltage across the terminals of the current source is not a problem the output resistance of the circuits in Figs. 2(a), 2(b) and 3(a) can be increased[1] by employing a field-effect operated in the common gate mode (see Fig. 4(a)). The output resistance is now r_o' where,

$$r_o' = r_{ds} + (\mu + 1) r_o \quad (5)$$

r_{ds} = incremental drain-source resistance of f.e.t.

μ = f.e.t. amplification factor.

For this scheme to be successful the f.e.t. must be a "normally-on" type working in the current saturation region.

A mathematical treatment of the d.c. circuit conditions is not difficult but a pictorial representation conveys the information most appropriate to a practical design. Curve (a) represents the base-emitter characteristic of Tr_1. The load line for R_E intersects this at I_D, δ_1. Curve (b) shows drain-source current, I_{DS}, as a function of source voltage. Since r_o is large relative to the impedance seen looking in at the source the actual source voltage, V_s, is found at the point on curve (b) corresponding to $I_{DS} = I_o$. Clearly it is required that $V_S > \delta_2$ if Tr_1 is not to saturate.

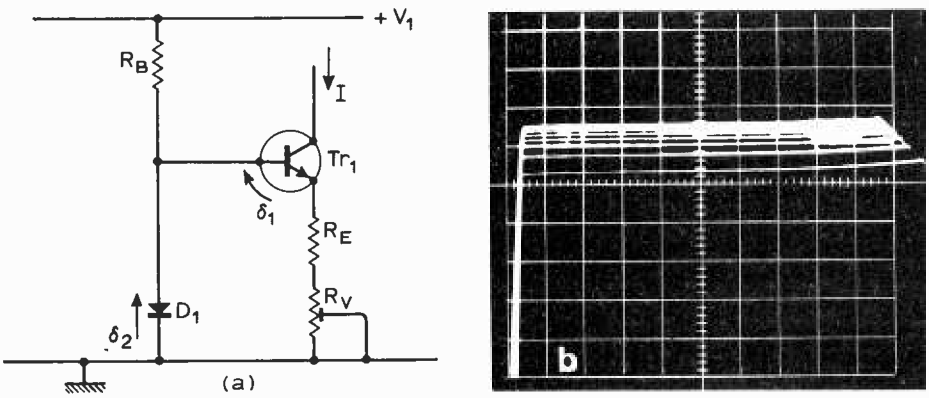

Fig. 3. (a) Useful alternative circuit to Fig. 2(a). (b) Output characteristics for circuit of Fig. 2(b). Scale: vertical 1 mA/div; horizontal 2 V/div. Steps 0.5 mA.

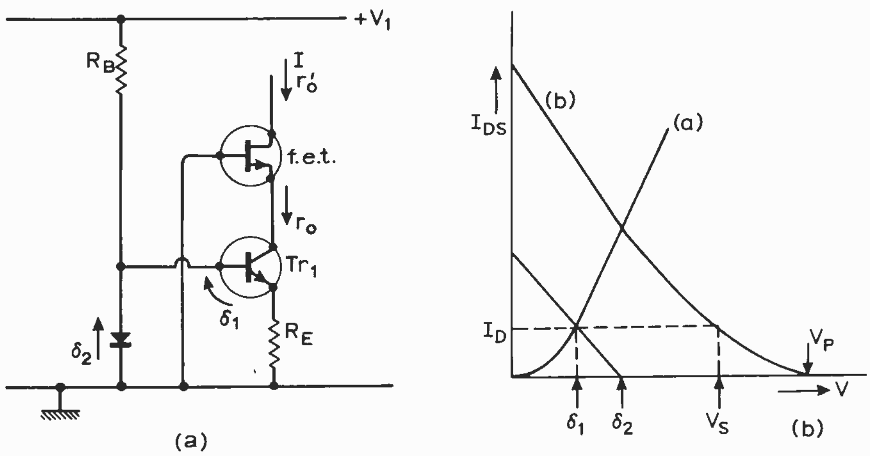

Fig. 4. (a) Improved form of circuit in Fig. 3(a). (b) Load line construction for 4(a).

Exploitation of the matching principle

One result of modern monolithic i.c. technology is the close parameter matching and temperature tracking of components of the same type fabricated on the same semiconductor slice. This feature is widely exploited in circuit design, giving rise to some arrangements which would not normally be encountered when using discrete components.

In Fig. 5(a), two transistors Tr_1, Tr_2, having an emitter area ratio $1{:}m$ respectively are made in close proximity on the same wafer. Tr_1 is given a collector-base strap in order to function as a diode. The circuit has been variously described as a "compound diode-transistor structure"[2] and "current mirror".

Ignoring base width modulation effects in Tr_2 and the difference in power dissipation in the two devices—both legitimate assumptions for small values of collector-

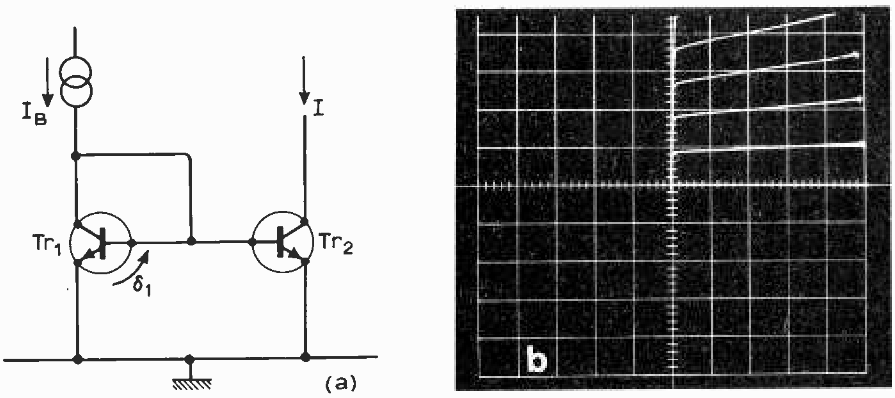

Fig. 5. (a) Integrated transistor pair in "current mirror" configuration. Emitter area ratio 1:m. (b) Current mirror characteristics using SL303A. Scales: vertical 1 mA/div; horizontal 2 V/div. Steps: 1 mA.

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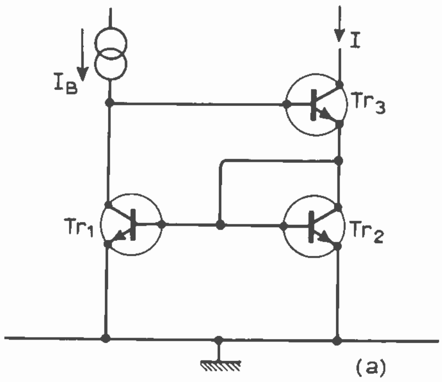


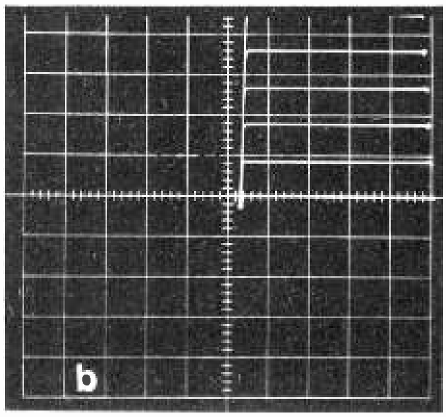


Fig. 6. (a) "Enhanced" current mirror circuit. (b) Characteristics of Fig. 5(a) using SL303A. Scales as in Fig. 5(b).

emitter voltage— it follows (see Appendix 1) that,

$$I = m\beta I_B/(m+\beta+1) \qquad 6(a)$$

or

$$I \approx mI_B \qquad 6(b)$$

for the usual case $\beta \gg 1$ and $m \leqslant 5$. The effective current gain of the combination is $\beta^* = I/I_B \approx m$. Gain precision has thus been obtained at the expense of gain magnitude—a familiar feature in feedback systems.

Equation 6(a) is true provided Tr_1 functions as a "well-behaved diode": in the present context this means a logarithmic current-voltage characteristic whatever the current level. Now the collector-base region of Tr_1, is, in fact, <0 by $\approx IR_{sat}$ where R_{sat} is the collector saturation resistance: it is thus necessary to use low R_{sat} devices, obtained for example by employing an n^+ buried layer technique, in order that equation 6(a) be valid.

Fig. 5(b) shows curve tracer results obtained using two matched devices of an SL303A (Plessey): in this case $m \approx 1$. Since m is a function of device geometry I is independent of T to a weak, calculable, extent. For the same approximations used in obtaining equation 6(b) from 6(a) it is simply shown that, at constant I_B,

$$(1/I)(dI/dT) \fallingdotseq (1/\beta^2)(d\beta/dT) \qquad (7)$$

In fact the application of local heating to the T0-5 header (via a Thermoprobe) caused no noticable shift in the characteristics on

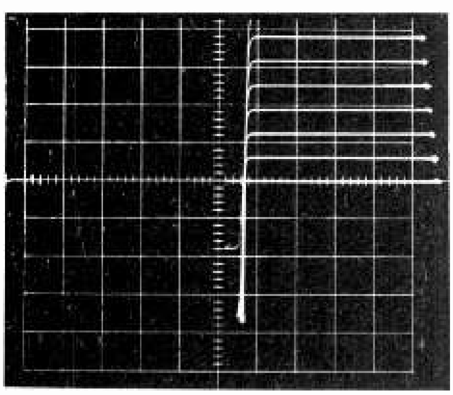


Fig. 7. Characteristics of Fig. 6(a) using sample X (Plessey). Scales: vertical 0·05 mA/div; horizontal 1 V/div. Steps: 0·1 A.

the scale used over a 50°C temp. range. However equation (7) must be used with caution since in a circuit I_B is not always fixed. If it is obtained by taking a resistor, R_B, from the collector of Tr_1 to a constant positive rail voltage, V_1, then R_B and δ_1 both vary with T. This means that, in principle, by suitable choice of a discrete component R_B variation of I_B with T can be made to almost cancel variation of β with T. One shortcoming of the current mirror so far described is the noticeable output resistance at the collector of Tr_2: this is not unexpected since as far as collector circuit of Tr_2 is concerned the transistor is connected in the common-emitter configuration with only a small incremental resistance between emitter and base.

At low current levels a resistance in the emitter lead of Tr_2 gives negative feedback and a consequent increased output impedance: for this case,[3]

$$R_E = (V_T/I)\log_e(I_B/I) \qquad (8)$$

where V_T = "thermal voltage" = (KT/q).

This approach permits the design of current sources in the μA range: this avoids the necessity of large values of R and the resulting expensive use of chip area in monolithic circuits.

The technique is not really suitable at current levels of a few milliamps because R_E becomes very small; this means no significant output impedance improvement due to feedback, and a significant lack of precision in I.

High output impedances for currents in the milliamp range can be achieved by using the circuit of Fig. 6(a). This scheme —which might be termed the "enhanced current mirror"—is a modification of Fig. 5(a), and appears to have been first used by Wilson.[4] Analysis, in Appendix 2, shows that if all the devices have equal emitter areas,

$$I = I_B/[1+\{2/\beta(\beta+2)\}] \qquad (9)$$

Fig. 6(b), which should be compared with Fig. 5(b), shows the increased output resistance occurring in this circuit.

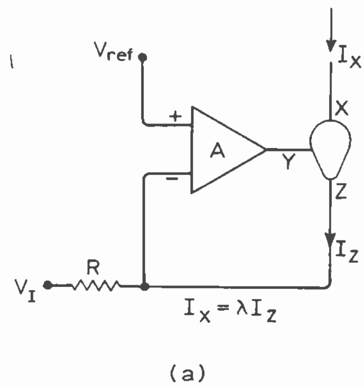
To obtain the trace an SL303A was used: this contains two matched transistors (Tr_1, Tr_2) suitable for use in a long-tailed pair and a third transistor (Tr_3) suitable for use in its current tail. Fig. (7) shows results for a specially supplied device in which the area ratio of the matched transistors used in the circuit of Fig. 6(a) is 1 : 3.

Feedback amplifier schemes

Some of the limitations of the simple current generators so far considered may be overcome by using circuits having a large amount of negative feedback. This condition is easily met if plenty of gain is available. Such gain is readily and cheaply obtained with i.c. operational amplifiers. The input stages of these amplifiers will, incidentally, almost certainly contain a current-mirror type of current source functioning as an emitter or common-emitter load.

Fig. 8(a) shows the arrangement of a class of precision current generators in which a discrete active device is combined with an i.c. amplifier. The device—represented by the pear shaped symbol[5]—may, in principle, be any amplifying element with three electrodes. Terminals X, Y, Z are defined in Fig. 8(b) for the three types of unit most likely to be used. Control is effected between Y and Z and the output current, I_x, appears at terminal X.

It follows from established operational



	Thermionic valve	Bipolar transistor	Unipolar transistor
X	Anode	Collector	Drain
Y	Grid	Base	Gate
Z	Cathode	Emitter	Source

(b)

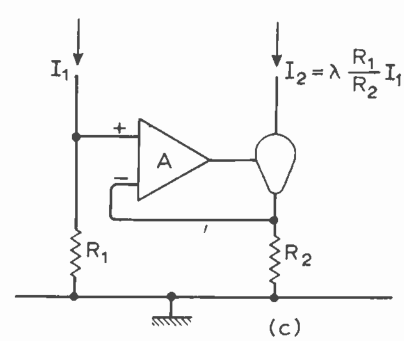


Fig. 8 (a) Current Source using 0 p. amp. and active device. (b) Terminal identification for general active device of 8(a). (c) Circuit for obtaining precise current ratio.

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amplifier theory that if $A \gg 1$ and common-mode effects are neglected,

$$I_x \approx \lambda(V_{ref} - V_I)/R \qquad (10)$$

The variables at our disposal for fixing I_x are thus V_1, V_{ref}, R. Suppose V_{ref} is considered as the "input". Then I_x increases with V_{ref}. Also, the current supplied by V_{ref} to the positive input is very low, being only the amplifier input bias current. The latter can be reduced to negligible proportions (<1pA) if an f.e.t. input stage is employed. Fig. 8(c) indicates a method whereby a current I_2 may be made strictly proportional to I_1. This follows from Fig. 8(a) by making $V_I = 0$ and $V_{ref} = I_1R_1$. When V_I is regarded as the "input", I_x decreases with V_I and the current supplied by V_I is the output current (I_x).

To reduce the current demanded of V_I while at the same time maintaining the same I_x as before the circuit of Fig. 9 may be used. V_I now exerts control over the current furnished through additional resistor R_z connected to constant supply V_{zz}. R' ($>R$) gives a reduced input current.

It must be noted that I_x is only as constant as the d.c. source(s) which define it. The effect of feedback is to increase the output resistance of the active device and to reduce the sensitivity of I_x to parameter variations in the active device due to ageing, replacement, etc. This is shown, for $V_I = 0$, by a more accurate expression for I_x, viz,

$$I_x \doteqdot \lambda(AV_{ref} - V_{YZ})/(A+1)R \qquad (11)$$

V_{YZ} is thus effectively reduced by a factor $1/(A+1)$.

The question arises as to whether it is better to use a bipolar transistor or an f.e.t. for the active device. This cannot be answered directly unless the requirements of the intended application are more closely specified. In some cases e.g. high current levels, a bipolar device is obligatory for current handling capability.

However, in making a choice the following points should be borne in mind. They apply to Fig. 8(a) for specified V_I, V_{ref}, R, and A.

1. It is the current in R that is "sensed" and kept constant by the feedback connection. Now with an f.e.t. $\lambda = 1$, so I_x is kept constant also. With a bipolar device $\lambda = \alpha$ (common-base current gain). Since α varies with T keeping the current in R constant will not ensure I_x's constancy.

2. Errors due to V_{YZ} uncertainty will be slightly greater for an f.e.t. simply because V_{YZ} is in general much larger.

3. Because of the large impedance mismatch between the output of the amplifier and the input of the f.e.t. (operated at zero gate current) using an f.e.t. is likely to give marginally more voltage loop gain.

The extent to which the last two factors are troublesome is obviously dependent on the magnitude of A.

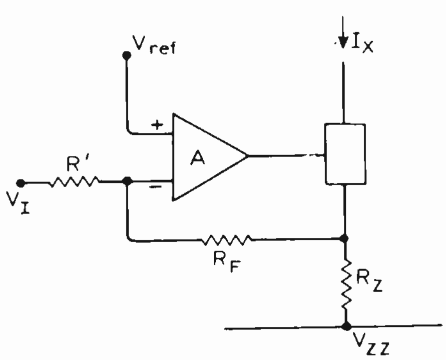

Fig. 9. Variation of Fig. 8(a) for reduced current from V_I.

Appendix

1. "Current mirror" analysis

If, in Fig. 5(a), I_{E1} and I_{E2} are the emitter currents of Tr_1, Tr_2 respectively, current addition at the commoned base terminal yields,

$$I_B = I_{E1} + \{I_{E2}/(1+\beta)\} \qquad \text{A.1.}$$

but, $I_{E1} = I_{E2}/m$ A.2.

$\therefore$ $I_{E2}[(1/m) + 1/(1+\beta)] = I_B$ A.3.

now, $I = \alpha I_{E2}$ A.4.

hence $I = \alpha I_B/[(1/m) + 1/(1+\beta)]$ A.5.

or, rearranging

$$I = m\beta I_B/(m+\beta+1) \qquad \text{A.6.}$$

and $(I/I_B) = \beta^* \approx m$, for small "$m$" and $\beta \gg 1$ A.7.

2. "Enhanced current-mirror" analysis

If, in Fig. 6(a),

I_{C1} = collector current of Tr_1,

I_{E3} = emitter current of Tr_3

then from A.6 above, with $m = 1$,

$$I_{C1} = \{\beta/(\beta+2)\}I_{E3} \qquad \text{A.8.}$$

summing currents at the base of Tr_3,

$$I_B = I_{C1} + I_{E3}/(1+\beta) \qquad \text{A.9.}$$

or, $I_B = I_{E3}[\{\beta/(\beta+2)\} + \{1/(1+\beta)\}]$ A.10.

now $I = \alpha I_{E3}$ A.11.

$$I = \alpha I_B/[\{\beta/(\beta+2\} + \{1/(1+\beta)\}] \qquad \text{A.12.}$$

rearranging

$$I = I_B/[1 + \{2/\beta(\beta+2)\}] \qquad \text{A.13.}$$

This gives (I/I_B) closer to unity than A.6 with $m = 1$.

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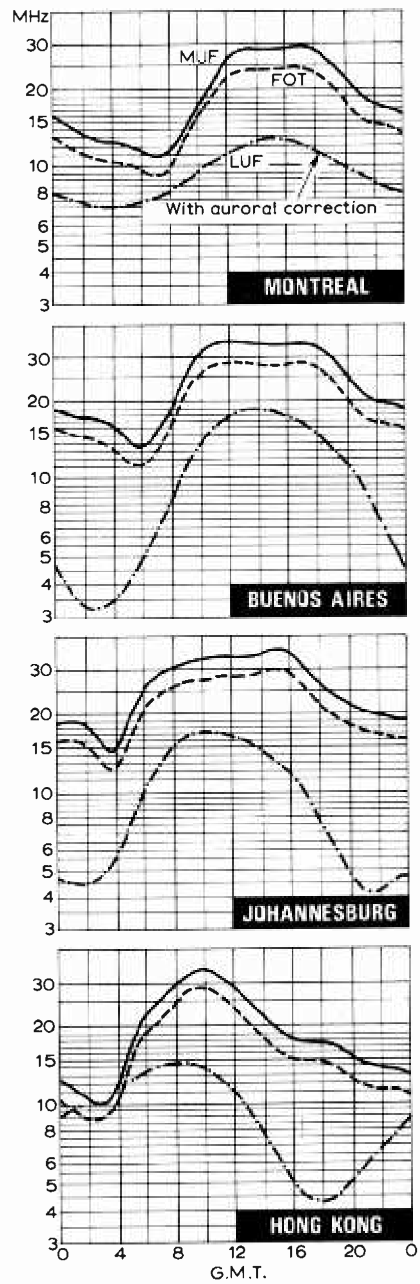
H.F. Predictions—October

The curves show median standard MUF (frequency at which high-and low-angle ray paths coincide), optimum traffic frequency (MUF exceeded on 90% of days) and lowest usable frequency (LUF) for reception in the UK. LUF is the frequency above which the signal/noise ratio at receiver input exceeds a required value on 90% of days. The curves were drawn by Cable and Wireless Ltd for commercial telegraph operation using high power and directive aerials. Curves for a 50kW broadcast service would be very similar and rough estimates for other services can be made by increasing LUF 1.5MHz for each 10dB drop in e.r.p.

Day-to-day variations in height and density of ionospheric layers and hence deviations from chart values are greatest during equinox months around sunspot maximum as at present.



World of Amateur Radio

Amateur radio in Poland

The Polish amateur radio society Polski Zwianek Krothofalowcow (PZK) is this year celebrating its 40th anniversary. A special jubilee meeting is being held on October 25th. Today there are 6000 members of PZK, about half of them licensed amateurs. Although the Polish authorities closed PZK in 1950, and then advised amateurs to join the official "League of friendship with the Army", the Polish amateurs re-established their own society in 1957 following internal changes in Poland in 1956. Current PZK activities include a journal, contests, QSL bureau, maintaining regional amateur measuring laboratories, running amateur radio courses and supplying emergency communications during floods or forest fires. Over 400 local radio clubs are in affiliation with PZK which, in 1963, became the first Iron Curtain country to re-join the International Amateur Radio Union. There are now over 2500 Class I licenses (h.f./v.h.f.), 450 Class II (v.h.f. only) and 400 club stations. Class I permits range from 20 watts (age 15 to 18), 50 watts (over 18), 250 watts (after 6 years) to 750 watts (on request after 10 years).

Long distances on MHz?

Two recent studies underline dramatically the potential of frequencies as high as 70 MHz to support long-distance ionospheric propagation at low-power. The Canadian amateur, Geoff Kennedy (VE2AI0), of Valois, Province of Quebec, has successfully received 70.275 MHz signals from the low-power TF3VHF beacon transmitter on Iceland, and is now striving to make a 50/70 MHz cross-band contact with the U.K. or continental Europe (he transmits on 50.055 MHz c.w.). The Canadian, already well-known for his intensive work on 50 and 70 MHz, believes there may be a connection between auroral, sporadic E and transequatorial (t.e.) modes of propagation.

This theory receives support from the results of an extended Japanese investigation into t.e. propagation on 32, 48 and 72 MHz over a 4850-km path from Darwin, Australia, to Yamagawa, Japan (K. Tao *et al, Journal of the Radio Research Laboratories* [Japan] January 1970). This demonstrates that, even in years of low sunspot activity, t.e. propagation frequently extends, particularly in the evening-to-midnight equinoctial periods, to at least 72 MHz. It is noted that t.e. conditions appear to follow "spread F" occurrences and correlate with local sporadic E conditions. The Japanese team suggests that around 32 MHz there are two separate t.e. modes; one during daylight subject to a violent interference-type fading range of about 25 dB; and the night-time mode with a fading range of about 10 dB. The interest of propagation research workers in t.e. was aroused initially by long-distance amateur contacts across the equator in the 50-MHz band during the period 1947-51; this interesting "chordal hop" mode has since been confirmed by many professional and amateur studies—but there clearly remains much opportunity for further study of t.e. in the 28-, 50- (not available in the U.K.) and 70-MHz amateur bands.

Rise in Morse Test charges

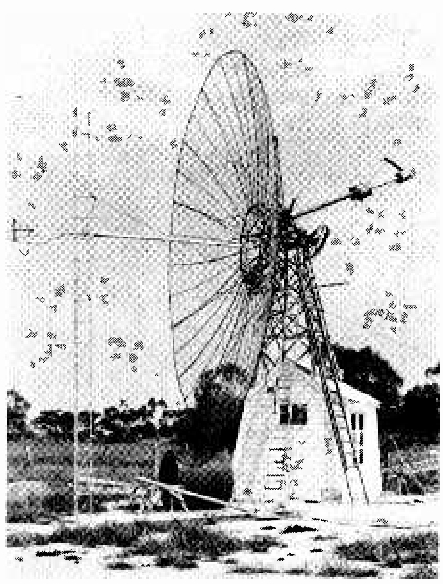
From October 1st, Minpostel is increasing the fee for the Morse Test needed to obtain a U.K. Amateur (Sound) Licence A from ten shillings to £1. The Federal Communications Commission has recently raised the fees charged to U.S. amateurs (who for many years paid no licence fees) though these remain modest by European standards. The charge is now $9 for new, renewed or upgraded operator licenses, but this covers five years. No fees are charged for Novice licences.

Trans-arctic expedition lecture

Radio communications played a vital role in the four-man polar crossing, led by Wally Herbert, in 1968-69. At an R.S.G.B. lecture on Monday, September 28th at 18.30 at the I.E.E., Savoy Place, London W.C.1, Squadron Leader Freddy Church (the expedition's base station operator), Dennis Collins (G2FLG) and Roly Shears (G8KW)—who organized and operated the G7AE "weekend" stations—are to describe the expedition's communications system.

In Brief: The Mullard Jubilee Exhibition at Mullard House, Torrington Street, London W.C.1 (October 5th to 24th, excluding Sundays) is to feature an amateur station built and operated by Mullard radio amateurs and using some of the firm's earliest transmitting valves.... The Port Talbot R.S.G.B. group won the National Field Day shield with a score of 2386 points; Oxford and District Amateur Radio Society was the runner-up with 2174 points. The Bristol Trophy for the leading one-station entry goes to Cannock Chase society with 1633 points. . . . An unusal callsign—HG 100UA/D—has been frequently heard recently on h.f. bands; this is an exhibition station operating from various venues in Hungary to mark the centenary of the birth of Lenin . . . . Membership of the International Amateur Radio Club, with headquarters in Geneva, is now 394.... A Scottish v.h.f. convention is being held at the Queen's Hotel, Dundee, on Sunday, October 11th. An afternoon programme of technical lectures will be followed by a dinner. Details from G. C. Somerville, GM3KYI 73 Balerno Street, Dundee . . . . A French biological study balloon, carrying beacon transmitters on 145.22 MHz (tone modulated) and 129.6 MHz (tone one-second pips) is to be launched in the Nancy area on Sunday, October 11th at 1400 G.M.T. and should reach maximum height about one hour later. Cardiac data on a rat will be telemetered on 27.4 MHz (reception reports to Georges Guinard, 15 Route de Villers, 54. Laxou).

PAT HAWKER, G3VA



This 28ft diameter dish was made by the Australian amateur VK3ATN for his moonbounce experiments. The reflector uses half-inch mesh. Gain is 38dB at 1296MHz and 28dB at 432MHz. The photo was taken by L. A. Moxon (G6XN) during a recent visit.

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October Meetings

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

LONDON

1st. I.E.R.E.—"Some aspects of the design of a universal radar viewing unit" by R. W. Elbourn at 18.00 at 9 Bedford Sq., W.C.1.

1st. S.E.R.T.—"Design and performance of modern loudspeakers" by R. L. West at 19.00 at the Royal Commonwealth Society, Craven St., W.C.2.

7th. I.E.R.E.—"Solid state h.f. communications receiver" by B. M. Sosin at 18.00 at 9 Bedford Sq., W.C.1.

8th. I.E.E.—"Communication by glass fibre" by R. B. Dyott at 17.30 at Savoy Pl.. W.C.2.

14th. I.E.R.E.—"The logistics of computer aided circuit design" by C. S. den Brinker at 18.00 at 9 Bedford Sq., W.C.1.

20th. I.E.E.—Colloquium on "Integrated circuits for consumer products" at Savoy Pl., W.C.2.

20th. I.E.E.—"Thyristor-bridge control systems analysis" by Dr. J. O. Flower and P. A. Hazell at 17.30 at Savoy Pl., W.C.2.

21st. R.Aero.Soc./I.E.E.—Discussion on the "The impact of modern avionics on safety" 18.00 at 4 Hamilton Pl., W.1.

21st. I.E.E.—"Technology and the Universities" by Professor John Brown the Electronics Division chairman at 17.30 at Savoy Pl., W.C.2.

21st. I.E.R.E.—"Continuing education for electronics engineers" by Dr. K. G. Stephens at 18.00 at 9 Bedford- Sq., W.C.1.

22nd. I.E.E.—"A state-space approach to modular circuit synthesis" by Dr. A. W. Keen at 17.30 at Savoy Pl., W.C.2.

26th. I.E.E.—"The training of systems engineers—a systems problem" by Dr. I. Cochrane at 17.30 at Savoy Pl., W.C.2.

28th. I.E.E.—"An introduction to image analysis" by G. Gardner and Dr Gibbard at 18.00 at 9 Bedford Sq., W.C.1.

29th. I.E.E.—"Developments in soft magnetic materials and their uses" by Prof. J. E. Thompson at 17.30 at Savoy Pl., W.C.2.

29th. I.E.R.E.—"Space communications—the present and the future" by J. M. Brown at 18.00 at 9 Bedford Sq., W.C.1.

ABERDEEN

8th. I.E.R.E.—"High fidelity recording and reproduction at 19.30 at Robert Gordon's Inst. of Tech. Schoolhill.

29th. I.E.E./I.E.R.E.—"Medical electronics" by J. M. Neilson at 19.30 at Robert Gordon's Inst. of Technology.

BANGOR

20th. I.E.E.—"Electronic measurement in the automobile industry" by M. H. Westbrook at 19.30 at the School of Eng'g Science, University College.

BATH

14th. I.E.E./I.E.R.E.—"Review of marine radio development" by G. J. Macdonald and C. S. Burnham at 18.00 at the University.

BOLTON

15th. I.E.R.E.—"Coaxial-line cavity Gunn effect oscillators" by P. W. Crane at 19.15 at the Institute of Technology.

BRIGHTON

27th. I.E.E.—Colloquium on "Advances in automatic pattern recognition" at 14.30 at the Polytechnic.

BRISTOL

19th. I.E.E.—"Telecommunications—past, present and future" by J. S. Williams at 18.00 at Queen's Bldg. the University.

27th. I.E.E.T.E.—"Zinc-air batteries" by G. W. Walkiden at 19.30 at Royal Hotel, College Green.

CAMBRIDGE

15th. I.E.E.—"Design techniques for mobile and personal radio telephones" by P. A. Webster at 18.30 at the Eng'g Labs, Trumpington Street.

20th. I.E.E.—"Current electronic developments in the deep-sea fishing industry" by R. Bennett at 19.30 at the College of Arts & Technology.

CARDIFF

1st. S.E.R.T.—"The all-transistor colour chassis" by T. Ayscough at 19.30 at Llandaff Technical College, Western Avenue.

21st. I.E.E.T.E.—"Education and training of technicians" by Dr. H. L. Hazlegrave at 19.30 at the University of Wales Institute of Science & Technology, Cathays Park.

28th. S.E.R.T.—"Test equipment for colour receiver servicing" at 19.30 at Llandaff Technical College, Western Avenue.

CATTERICK

13th. I.E.E.—"The teaching of digital techniques" by D. Brown at 18.30 at the Camp.

CHATHAM

22nd. I.E.R.E.—"Thick film microelectronics" by A. F. Dyson at 19.00 at the Medway College of Technology.

CHELMSFORD

7th. I.E.E./I.E.R.E.—"The new Post Office research station at Martlesham" by C. F. Floyd at 18.30 at the King Edward VI Grammar School, Broomfield Rd.

DUNDEE

20th. I.E.E. Grads.—"Computer memory systems" by J. Crabb at 19.30 at the University.

ENFIELD

29th. I.E.E.—"Microelectronics" by E. T. Emms at 18.30 at the College of Technology.

FARNBOROUGH

13th. I.E.E.—"Electronics character recognition" by R. H. Britt at 18.30 at the Technical College.

GUILDFORD

6th. I.E.R.E.—"A computer for teaching" by V. F. Thomas at 19.00 at the Technical College.

HULL

29th. I.E.E.—"Review of electronics in cars" by W. F. Hill, at 18.30 at the Y.E.B.

LEICESTER

6th. I.E.R.E./C.E.I.—"How an engineer can take part in the total management of a company" by Dr. F. E. Jones at 19.00 at the University.

14th. I.E.E.T.E.—"Control systems, computer simulation" by P. J. Lawton at 18.45 at the Hawthorn Bldg, the Polytechnic, The Newarke.

LETCHWORTH

21st. S.E.R.T.—"Art of computation" by C. Fleckney at 19.00 at ICL Engineering Training Centre, Icknield Way West.

LIVERPOOL

14th. I.E.R.E.—"Fast switching techniques as applied to automatic telephone exchange design" by J. Lewsley at 19.00 at the University, Dep. of Elec. Eng.

26th. I.E.E.—"The role of the polytechnics by G. Bulmer at 18.30 at the Polytechnic.

LOUGHBOROUGH

20th. I.E.E.—"Measuring the body's signals" by J. M. Ivison at 18.30 at Ed. Herbert Bldg, the University.

MALVERN

7th. I.E.R.E.—"Quasars—the most powerful transmitters in the Universe" by Dr. P. J. S. Williams at 19.00 at the Abbey Hotel.

MANCHESTER

12th. I.E.E.—"The polytechnics in perspective" by Sir Eric Richardson at 18.15 at Renold Bldg, U.M.I.S.T.

NEWCASTLE ON TYNE

7th. S.E.R.T.—" B.R.C. Colour TV 3000 power supplies" by K. R. Harris at 19.15 at Charles Trevelyan Technical College. Maple Terrace.

14th. I.E.R.E./I.P.P.S./R.T.S.—"Colour television" D. G. Packham at 18.00 at Tyne Tees Television Studios, City Road.

NEWPORT, I.O.W.

16th. I.E.R.E.—"The future of aircraft landing systems" by W. F. Winter at 19.00 at the Technical College.

23rd. I.E.E.—"Electronic performance testing of motor vehicles" by D. C. Freeman at 18.30 at the Technical College.

POOLE

20th. I.E.E. Grads.—"Integrated circuits in hi-fi systems" by B. A. Reed at 18.30 at the Technical College.

READING

6th. I.E.E.—"Review of electronic telephone switching" by D. J. Harding at 19.30 at the J. J. Thomson Lab., the University.

15th. I.E.R.E.—"Computer aided instruction" by Dr. D. G. Bate at 19.30 at the J. J. Thomson Laboratory, the University, Whiteknights Park.

ST. AUSTELL

6th. I.E.E.T.E.—"Communications satellite" by S. Pitham at 19.30 at the English Clays Lovering Pochin Ltd, Staff Restaurant, John Keay House.

SHEFFIELD

28th. I.E.E./I.E.R.E.—"Large scale integrated circuits" by C. S. den Brinker at 18.30 at the University.

SOUTHAMPTON

21st. I.E.E./I.E.R.E.—"Digital techniques in moving target indication" by I. C. Walker at 18.30 at the University.

SWINDON

6th. I.E.R.E.—"Thyristor engineering " by D. B. Corbyn at 18.15 at the College.

TAUNTON

13th. I.E.E.—"Television communications" by A. James at 19.45 at the Castle Hotel.

WEYMOUTH

22nd. I.E.E.—"Holography" by A. E. Ennos at 18.30 at the South Dorset Technical College.

WHITBY

13th. I.E.E.—"Radio astronomy" by I. W. Sheffield at 19.00 at Botham's Cafe, Skinner St.

WOLVERHAMPTON

6th. I.E.R.E.—"What the C.E.I. means to the professional engineer" by H. F. Schwarz and Sir Arnold Lindley at 19.15 at the Polytechnic.

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New Products

M.O.S. Multiplex Switches

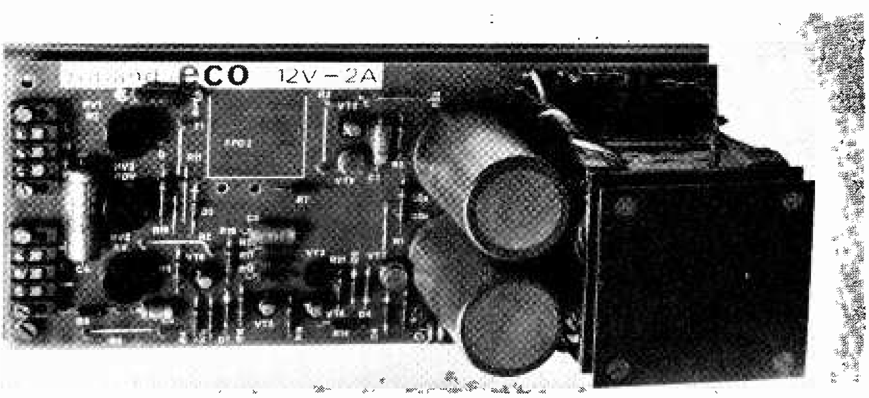
High-performance m.o.s. multiplex switches for communications are available in two ranges from Marconi-Elliott Microelectronics, one with leakage currents of 10nA and the other with leakages of the order of 0.1nA. Low-leakage types have been specially selected and are already used in a military message-switching application. Range includes single switches, with either high input or low 'on' resistance (M101, 103 and 106), dual switches, with either high input or low 'on' resistance (M203 or 206), dual pair used as multiplex switch and digital-analogue converter (M406), common-source sextet (M605), and triple pair digital-analogue converter (M606). Devices, all p-enhancement types, are made by the thick-oxide technique, used in high-voltage m.o.s. devices. Process allows increased oxide thickness by vapour deposition technique underneath non-active areas of aluminium conductors, thus avoiding spurious leakage. Marconi-Elliott Microelectronics Ltd, Witham, Essex.

WW319 for further details

Stabilized Power Supplies

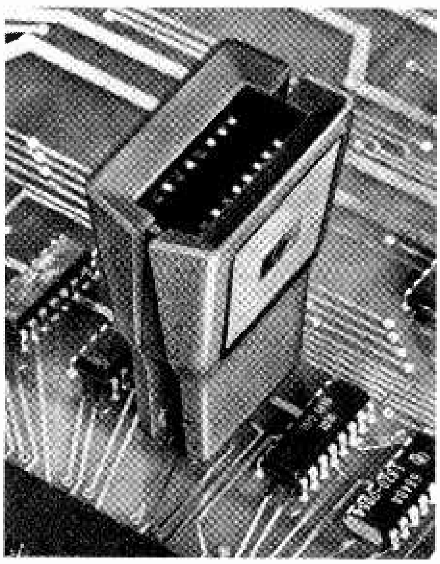
The Roband ECO range of stabilized power supplies provides outputs of up to 50V and 10A. All components are mounted on a single printed board. Stabilization is about 0.005% with ripple and noise typically 150μV pk-pk. A modified re-entrant current characteristic maintains protection of loads and supplies, while preventing lock-out on linear and non-linear loads. They can also be externally programmed and used as constant current supplies. Protection against overvoltage, without spurious shut-down, is made possible by an optional addition. The units range in price from £20 to £49. Roband Electronics Ltd, Charlwood Works, Charlwood, Horley, Surrey.

WW312 for further details



Logic-state Indicator

Hewlett-Packard model 10528A Logic Clip fits on to t.t.l. or d.t.l. integrated-circuit packages and instantly displays the

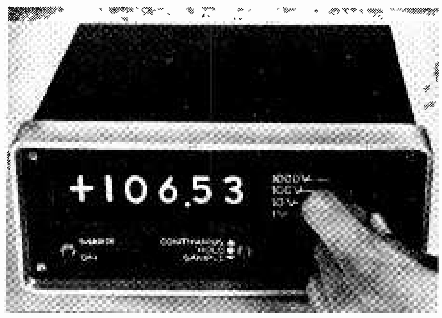


logic states of all 14 or 16 pins. The clip has 16 light-emitting diodes, each of which follows voltage-level changes on one pin. A lighted diode indicates a high logic state (+5V). The Logic Clip is self-contained. It requires no power connections or adjustments, drawing its power from the circuit being tested, and contains logic circuitry for locating the ground and +5V pins even if clipped on unsymmetrically. The buffered inputs put no more than one t.t.l. load on the circuit being tested. Hewlett-Packard Ltd, 224 Bath Road, Slough, Bucks.

WW313 for further details

Highly Accurate Digital Voltmeter

Digital voltmeter DSV4 from International Electronics has an accuracy of ±0.01% ±1 digit. Input resistance is 10MΩ on all ranges. There are 12½ readings per second. Calibration is automatic between each displayed heading and is effected by reference to a temperature compensated



Weston standard cell. Zero offset current and voltage are automatically corrected between each reading. Series-mode rejection at mains supply frequency and its harmonics is greater than 85dB, and common-mode rejection 100dB at mains frequency and d.c. with 1kΩ imbalance in the input leads. The price is £25. International Electronics Ltd, Ewood Bridge, Haslingden, Lancs.

WW303 for further details

Quarter Squares Multiplier

Quarter squares multiplier type M3-5C from Computing Techniques obtains the product of two functions by implementing the relationship $XY=\frac{1}{4}(X+Y)^2 - (X-Y)^2$. Three built-in operational amplifiers perform the summing and inverting operations, while two function generators approximate the required parabolic transfer functions. Four-quadrant operation is provided. Using D1-2 amplifiers, the phase shift is less than 1° at 1kHz; this accuracy can be maintained up to 10kHz, and full power can be obtained up to 30kHz (with reduced accuracy) by using type F1-7 amplifiers. The required inputs are variable A, variable B and reference voltages; the input impedance is approximately 3kΩ. The available outputs, selected by miniature key switch, are $-AB$, A^2-B^2 as a voltage and A^2+B^2 as a current; multiplication accuracy is 2% f.s.d. The product output voltage is AB/10 and the output current is ±5mA at ±10V. The multiplier is fitted with a standard

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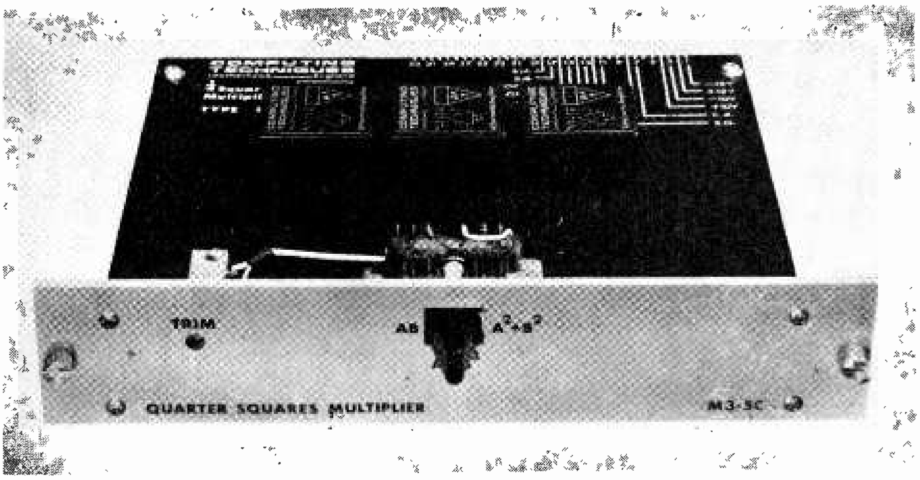


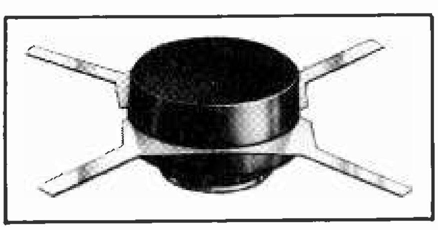
I.S.E.P. 33-way plug and is intended primarily for use with the Comtec analogue/hybrid computers Vidac 336 and Vidac 169. It may, however, be built into any analogue system operating on a machine unit of 10V. Computing Techniques Ltd, Westminster Bank Chambers, Bridge Street, Leatherhead, Surrey.
WW 330 for further details

Opto-electronic Switch for T.T.L.

Light-activated integrated switches have been produced with outputs compatible with both transistor logic and relays. Devices can be used in many applications of light beams including punched card reading and are said to be cheaper than photocells. Included in the TO-18 can is an m.o.s. Schmitt trigger circuit integrated into the same silicon chip. Response can be up to 40kHz and switching threshold is variable over three orders of magnitude by an external *RC* time constant. Devices are available with either 12-15V or 24-30V operating voltage and logic value of 0 or 1 in the dark state. Price is £2 8s for 1-50 and £1 10s for 5000 up. Arrays of 40 and 60 diodes for character recognition, edge sensing, etc., are also announced by Integrated Photomatrix Ltd, Grove Trading Estate, Dorchester, Dorset.
WW 321 for further details

U.H.F. Transistor

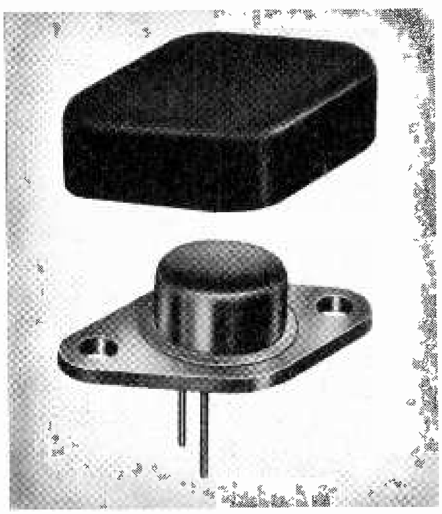
The latest addition to the Mullard range of u.h.f. transistors is intended for use as a driver in mobile telecommunications equipment or as an output stage in pocket transmitters. Type BFW98G, it will deliver an output of 500mW at 470MHz when operating with a supply voltage of 13.8V



and a drive of 80mW. The transistor is an n-p-n silicon planar device in a new stripline package. Mullard Ltd, Mullard House, Torrington Place, London W.C.1.
WW304 for further details

Cover for Power Transistors

A cover designed for use with the TO-66 packaged devices is available from Jermyn. Designated the A22/2005, it is moulded in black Rilsan BMNP40 which is capable of



withstanding temperatures up to 113°C and provides electrical insulation up to 1000V. The cover has a lip to hold it in place, Jermyn Industries, Vestry Estate, Sevenoaks, Kent.
WW308 for further details

Dynamic Shift Register

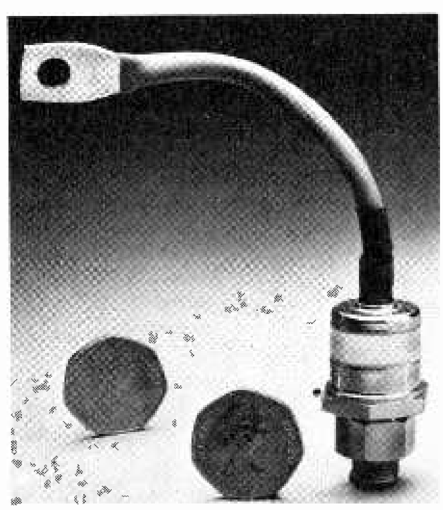
A 512-bit dynamic shift register, the MM5016, from National Semiconductor operates on a standard +5V and −12V power supply and is compatible with bipolar transistors. The device has a 600Hz guaranteed minimum operating frequency at 25°C and an input tap gives 500 or 512 bits. It is available in either TO-5 or dual-in-line packages. No pull-up or pull-down resistors are required on the input and output. Price is 50s for the TO-5 version in lots of 100 or more. U.K. distributors: Athena Semiconductor Marketing Co., Egham, Surrey; Electronic Component Supplies (Windsor) Ltd, Thames Avenue, Windsor, Berks; Farnell Electronic Components Ltd, Canal Road, Leeds LS12 2TU; and ITT Electronic Services, Edinburgh Way, Harlow, Essex.
WW 332 for further details

M.O.S. Adaptive Logic for Pattern Recognition

Adaptive logic gate, originally intended for use in pattern recognition experiments at Kent University, is now commercially available from Integrated Photomatrix. The m.o.s. gate (MC901) decodes all possible logic functions of the four binary inputs and sets one of the 16 internal bistable stores. Outputs of these stores are OR-gated to form the output. The adaptive gates can be connected into an array that will 'learn'. If, say, the array is controlling a system, feedback loops can be provided from the system to the array. The output of the array will be a function of the current input and, because of the memory stores, also a function of previous inputs and results as defined by the feedback loops. A particular input pattern does two things. First, it gates the output of one of the stores providing an output if that store contains a '1', second, it gates the input of the same store allowing it to set or reset depending on which of two input lines a $1\mu s$ pulse is applied. Two additional inputs allow all the stores to be reset. Available in dual in-line packages, price is £7 for 1-99 and £4 10s for 1000 up. Supply voltage is 27V. Integrated Photomatrix Ltd, Grove Trading Estate, Dorchester, Dorset.
WW 323 for further details

300-W Zener Diode

Type BZX86 zener diode from Mullard has a continuous power rating of 300W at a mounting base temperature of 65°C; and at an ambient temperature of 30°C it will withstand surges of 2kW. It is intended for use in preventing large electrical transients causing damage to equipment. It has a maximum diameter of 27mm and is 84mm



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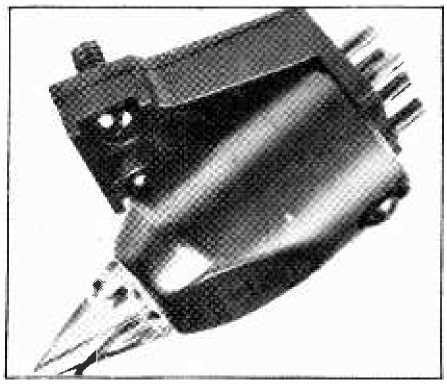
high. Zeners with voltages between 10 and 75V can be supplied. Mullard Ltd, Mullard House, Torrington Place, London W.C.1.
WW 305 for further details

Pickup Cartridge

Aluminium stylus mounting is used on a new Bang & Olufsen medium-compliance magnetic pickup cartridge (model SP14).



The cartridge is included in a turntable unit to be released later this year, but is available separately.

response	15Hz-25kHz ± 4dB 20Hz-16kHz ± 2.5dB 40Hz-10kHz ± 1.5dB
separation	> 20dB at 1kHz
channel difference	< 2dB
compliance	15×10^{-6}cm/dyne
effective stylus mass	1mg
stylus pressure	1.5-2.5g
output	1mV/cm/sec (typically 5mV)
impedance	47kΩ
stylus	15μm radius, spherical

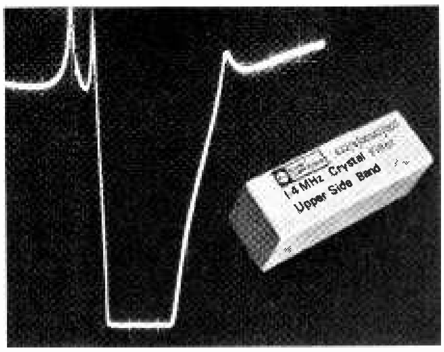
Price £8 10s. Bang & Olufsen U.K. Ltd, Eastbrook Rd, Gloucester GL4 7DE.
WW 322 for further details

Monolithic Mica Capacitors

Using monolithic construction the MS611 series of capacitors available from WEL provides high stability over a wide frequency range. The units will operate over a temperature range −55 to +100°C. Insulation resistance is about 10^5 megohms and tan d is better than 0.001 at 1 kHz and better than 0.002 at 1 MHz. The upper frequency limit is about 1 GHz. Voltage ratings up to 750V are provided and the standard E12 range of values at 1% tolerance is available from stock. Capacitance range is 10-10,000pF. WEL Components Ltd, 5, Loverock Road, Reading, Berks.
WW 327 for further details

Crystal Filters for Marine Communications

Plessey have introduced a new standard range of crystal filters for use in marine communications equipment employing an i.f. of 1.4MHz. Included in the standard



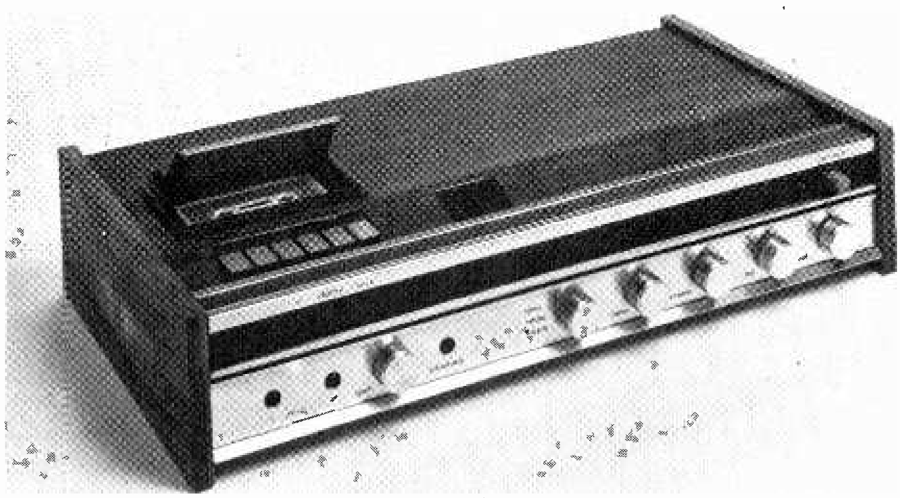
range are upper and lower sideband filters, and symmetrical filters for a.m. and c.w. applications. In each case, the design meets the selectivity requirements of the Post Office specifications, TSC105 or MFT1201. Operation is specified over the temperature range −10 to +40°C, and each filter is sealed in a metal case measuring 76 × 27.5 × 26mm. The new range is complemented by a 2.182MHz sideband filter which has been specially designed for use on emergency marine equipment, and a broad-band LC filter giving selectivity from 1.6MHz to 4.2MHz while rejecting 1.4MHz. Filter Unit, Plessey Components Group, Titchfield, Hants.
WW 334 for further details

1.5–MHz Recording Head

Gresham Recording Heads have produced a high-frequency analogue recording head for operation at 1.5MHz. The 14-channel wideband head-sets are made with a gap of only 25μ-inches. They are designed for direct analogue recording at 1.5MHz on 1-in tape. The normal tape speed is 120 i.p.s. Gresham Recording Heads Ltd, Feltham Trading Estate, Feltham, Middx.
WW 328 for further details

Cassette Recorder with Dolby System

A new cassette tape-recorder unit, the Kellar DTA 50 available from Kellar Electronics, incorporates the Dolby 'B' noise reduction system, and is claimed to give a performance as good as that obtained from gramophone records. The unit comprises a high-quality transport mechanism and a stereo amplifier capable of a continuous power output of 25W per channel. Ordinary cassette recordings can be played on the unit, but the Dolby circuit can be switched in to allow low-noise recordings to be made. The frequency response of the recorder is 40Hz to 12kHz ± 3dB. The wow and flutter is given as 0.15% w.f.d. peak according to C.C.I.R. The use of the noise reduction system results in an increase of 5dB in the signal-to-noise ratio. Price £150. Kellar Electronics Ltd, Maryland Works, 9 Brydges Road, Stratford, London E.15.
WW 324 for further details



10-turn Precision Pot.

Bourns have introduced a new ⅞in diameter, 10-turn, wirewound potentiometer with a glass-filled nylon housing and bushing, and a polycarbonate shaft with a screwdriver slot. Rotational life is guaranteed to 1 million shaft revolutions. The resistance range is 100Ω to 100kΩ, and tolerance ±5%. Deviation from linearity is ±0.25% max. Maximum power dissipation is 2.0W at 25°C. The operating temperature range is −55 to +105°C. Bourns (Trimpot) Ltd, Hodford House, 17/27 High Street, Hounslow, Middx.
WW311 for further details

Silicon Infra-red Light Detector

The MSP70 infra-red detector from MCP Electronics Semiconductor Division has, like its predecessors the MSP3 and MSP6, a black tubular housing 1in × 0.25in. It will detect modulated infra-red up to 60kHz. MCP Electronics Ltd, Alperton, Wembley, Middlesex, HA0 4PE.
WW 335 for further details

D.C. Multimeter

Multimeter measuring direct current down to picoamperes can now be supplied by Wayne Kerr. Model M300, first seen on the

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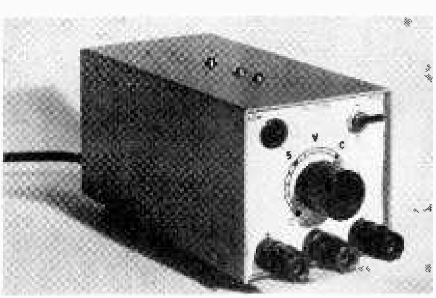
Wayne Kerr stand during the I.E.A. exhibition earlier this year, also measures direct voltage and resistance. Amplifier uses photochopper stabilization with 40-Hz switching to avoid beats with the 50-Hz supply. Instrument also features a small reverse scale for null detection, overload protection, one set-zero control for all ranges, a recorder output of 0-1mA at 1kΩ, and a hum rejection filter which rejects hum equivalent to 50 × f.s.d.

current ranges	30pA—100mA in 20 ranges
accuracy	±1.5%
voltage	30μV—300V in 15 ranges
accuracy	±1%
resistance	10Ω—1GΩ
accuracy	±2% (±5% 1GΩ range)
input Z	100MΩ
common mode rej.	120dB a.c. or d.c.
noise	< 1μV or 1pA r.m.s.
stability	±3μV, ±3pA
response time	< 1s (2s with filter)

Available for 110V operation. Price £150. Wayne Kerr Co. Ltd., New Malden, Surrey.
WW317 for further details

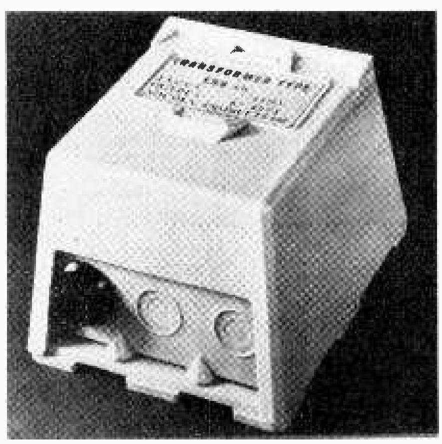
Miniature Power Supply

The PS5 supply from A.D.M. Electronics measures 70×70×180mm and provides up to 0.17A over the voltage range 0–15V. The PS5 design has been developed to eliminate damaging transient overshoot at turn-on and turn-off. The supply is fully protected against short circuits and overloads. The transient response is less than 5μs for a full load-current change and the regulation is claimed to be better than 5000 to 1. On full load the ripple and noise do not exceed 500μV, reducing to less than 30μV for load currents below 50mA. A.D.M. Electronics, P.O. Box 3, Merthyr Tydfil, Glamorgan.
WW310 for further details



Range of Transformers

A series of low-voltage, double-insulated transformers from Adcola has a primary-winding range of 110 to 380V and a secondary range of 6 to 115V. Ratings are 25, 40 or 60W. The casing is secured by clips so dismantling for access to internal parts is quick and does not require the use of a screwdriver. Adcola Products Ltd, Adcola House, Gauden Road, London S.W.4.
WW307 for futher details



Monolithic Tuning Indicator

An integrated circuit tuning indicator is available from Motorola. The addition of a miniature lamp bulb turns the i.c. into an aid to the fine-tuning of f.m. radio and colour television receivers. In use type MC1335 tuning indicator is connected across the f.m. ratio detector. When the receiver is correctly tuned equal voltages appear across each half of the ratio detector centre-tapped coil causing the lamp to light. Unequal voltages turn the lamp off. The device is encased in an eight-lead dual-in-line plastic package and requires a power supply of 20V. The price is 27s each in quantities of 100 or more. Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middx.
WW309 for further details

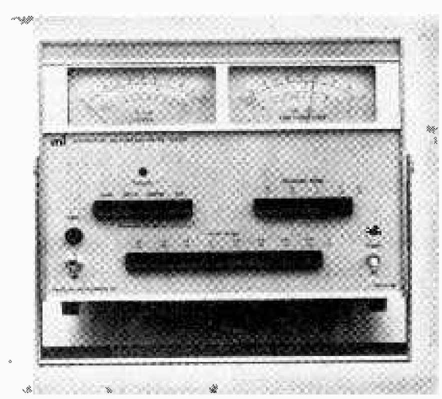
V.H.F. Personal Radiotelephone

A v.h.f. personal radiotelephone consisting of a single lightweight unit suitable for emergency, public and private usage is available from Pye Telecommunications. It is known as the PF2FMB, and is equipped for three channel f.m. operation in the 68-88 and 148-174 MHz bands. It meets the Post Office mobile specification and, with an adaptor, can be used in a car or other vehicle. Sockets on the top of the unit make connection with a small external loudspeaker/microphone unit, and an aerial. The unit can be clipped to the waist belt or carried in a specially designed case or harness slung from the shoulder with a flexible wire antenna fitted in the shoulder strap, its loudspeaker-microphone being clipped to the lapel. The whole unit, including battery, weighs only 28 oz. Power is from a single 15V rechargeable nickel-cadmium battery with a capacity of 200mAh, the unit incorporates a circuit to reduce battery drain when no signal is received. Pye Telecommunications Ltd, Newmarket Road, Cambridge.

WW 331 for further details

Distortion Factor Meter

Push-button controls make this new distortion factor meter easy to use especially for measurements on production radio and TV receivers and audio equipment.



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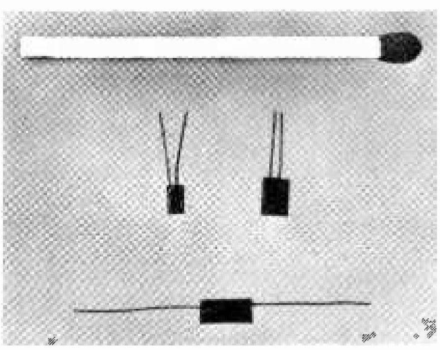
Meter, type TF2337, is complementary to the more elaborate Marconi Instruments TF2331 announced in 1964 and intended for development work. Level is set to give adequate meter deflection at either 400Hz or 1kHz and distortion factor read directly from the other meter. Provision is made for injection of external fundamental frequency in the range 30Hz-7kHz.

level ranges	10mV-30V in 8 ranges
accuracy	±5% f.s.d.
(10mV range	±10%)
distortion ranges	1-30% in 4 ranges
accuracy	±5% f.s.d.
input Z	100kΩ plus 50pF
fundamental rejection	56dB
frequencies	400Hz and 1kHz plus external facility

Available for 110V use. Price £230. Marconi Instruments Ltd, St. Albans, Herts.
WW316 for further details

Small Tantalum Capacitors

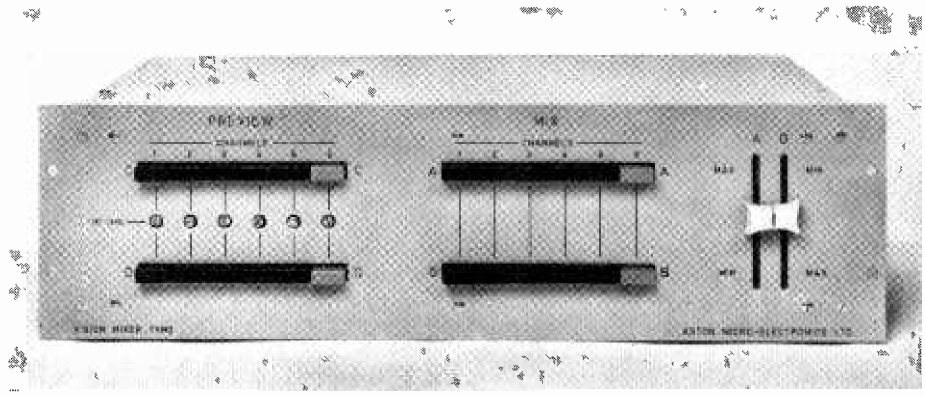
The Kemet Micron range of small tantalum capacitors, from Union Carbide U.K., are available in either rectangular or cylindrical cases with radial or axial leads. The capacitance range is 0.001 to 220 μF, and voltage range 2 to 100V. Standard capacitance



tolerances are ±20%, ±10% and ±5% and the operating temperature range is −55 to +125°C. The capacitors consist of very small tantalum anode assemblies with attached leads enclosed in insulated cases filled with epoxy resin. Union Carbide U.K. Ltd, 8 Grafton Street, London W1A 2LR.
WW302 for futher details

Video Signal Mixer

From Aston Micro-Electronics Ltd we have received details of a vision mixer TVM2 for cutting, fading and mixing the pictures from five cameras and one video tape recorder. Two rows of button switches allow any two vision sources to be selected for viewing before transmission. Automatic interlocks prevent accidental mixing of the video tape recorder picture with pictures from the cameras. Although designed for monochrome systems, the 8MHz response of the TVM2 is adequate for colour television. It is claimed that PAL-encoded colour pictures can be mixed without



noticeable degradation of picture quality. Aston Micro-Electronics Ltd, Vapery Lane, Pirbright, Woking, Surrey.
WW314 for further details

Wideband Test Oscillator

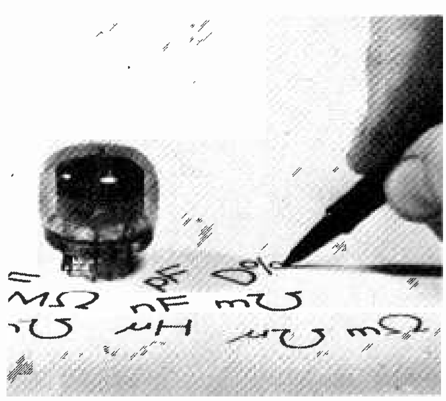
Model 4200 oscillator, manufactured by Krohn-hite and available from Omega Laboratories, provides a sine wave in the range 10Hz to 10MHz with 0.5W of power and 0.1% distortion. Frequency tuning is by means of a dial and six push buttons. Amplitude calibration of 0.2dB accuracy is obtained by an eight-position push-button attenuator calibrated in 10dB steps. A continuous output control combined with the attenuator provide 90dB of attenuation. Frequency accuracy is 2%. A fixed 1V output independent of the main output is also available. Price £235. Omega Laboratories Ltd, 59 Union Street, London S.E.1.
WW301 for further details

Units Symbol Numicator Tubes

End-viewing indicator tubes are now made by Hivac which display special symbols—units in particular. Two tubes have the following symbols.

GR8M	GR81M
pF, nF, μF, mF	nΩ, μΩ, mΩ
μH, mH, H	Ω, kΩ, MΩ
%	D%, %

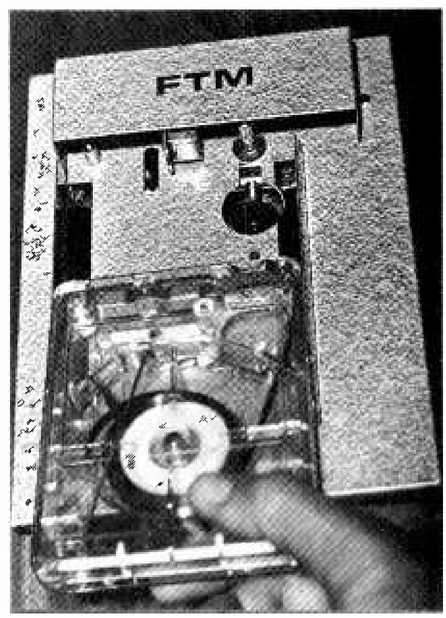
Both operate with minimum strike voltage of 170V and maintaining voltage of 140V. Minimum cathode current needed is 2mA with mean of 5mA and maximum peak current of 20mA. Because of the smaller area of the %, Ω and H symbols, current



requirement is half of these figures. Base is B13B and tube height is 26.5mm. Price is about £2 for 100 up. Hivac Ltd, Stonefield Way, South Ruislip, Middx.
WW318 for further details

Tape-transport Mechanism

Fitch Tape Mechanisms have developed and are now producing an endless-loop tape-transport mechanism using a new system of pinch wheel operation making tape stiction impossible even if the unit is



not actuated for long periods with the tape cartridge in the ready position. This mechanism, adaptable to all sizes of cartridge, can be battery or mains powered. It is fully automatic and can be remotely controlled. Fitch Tape Mechanisms, 7a Balham Grove, London S.W.12.
WW 333 for further details

Corrections

"Miniature Tape Recorder" (August p. 413): The complete address for Hayden Laboratories Ltd, who market the Nagra SN recorder, is East House, Chiltern Avenue, Amersham, Bucks.

"Switching Diode" (September p. 463): The diodes illustrated are not the Mullard BAV 44 type described.

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Literature Received

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

ACTIVE DEVICES

A range of zener diodes and varistors manufactured by Schauer Semiconductors is described in literature obtainable from LST Electronic Components Ltd, 7 Coptfold Rd, Brentwood, Essex. The varistors consist of two matched semiconductor junctions and are suitable for meter protection, fractional voltage regulators, etc. WW401

From REL Equipment and Components Ltd, Components Division, Croft House, Bancroft, Hitchin, Herts, a catalogue devoted to semiconductors, pot-cores, transformers, alkaline cells, capacitors and resistors WW402

We have received the following data sheets from Marconi-Elliott Microelectronics, Witham, Essex.
- M101, MO3. Single metal oxide silicon p-enhancement transistor, 3.5 to 6V threshold, 30V breakdown and 2nA max off leakage. WW403
- M106. Single m.o.s.t. with low 'on' resistance (30Ω), p-enhancement, V_{DS} = −30V and P_{max} = 350mW WW404
- M206. Dual m.o.s.t. on a common substrate, similar to M106 .... WW405
- M203, 406, 605. 30V m.o.s.ts—203 is a pair without gate protection; 406 consists of two common-source pairs; and M605 is six common-source m.o.s.ts in a d.i.l. package WW406

Integrated Photomatrix Ltd, Grove Trading Estate, Dorchester, have sent us a number of product data sheets.
- IPL11. Light activated switch in TO-18 case consists of silicon planar diode and an m.o.s.t. i.c. for 28V operation WW407
- IPL1100. Similar to above for −12V operation WW408
- IPL13. Light-to-frequency convertor; planar diode and i.c. in TO-18 case, output 10Hz to 100kHz WW409
- IPL14. Analogue light level sensor in TO-5 case WW410
- HA14. Paper tape reader head. Planar diodes and i.cs, no interface circuit required for driving t.t.l. or m.o.s. WW411
- IPL20. 50 × 1 light sensitive array. 50 silicon planar photodiodes and a 51-bit shift register on a chip WW412
- IPL30/D. Position-sensitive photocell system. Gives X and Y output voltages proportional to the distance of a light spot from a central position. Accurate to better than 1μm .. WW413
- WCM22. Material width monitor WW414
- IPL20R. 50 × 1 Photodiode array driver. WW415
- IPL15. Light activated switch WW416
- MR103/4/5/6 A 16-bit shift register; t.t.l. compatible m.o.s. WW417
- MC901. Adaptive logic gate WW418
- Price list WW419

We have received a 111-page brochure from British Brown-Boveri Ltd, Glen House, Stag Place, London S.W.1, called "Thyristors-Data and Diagrams" WW420

American Diodes Incorporated have just started a company in the U.K. called Diodes Ltd at Fairacres Estate, Dedworth Rd, Windsor, Berks. The following literature is available
- Short-form catalogue WW421
- Price list WW422

The 1970 catalogue of Electrovalue, 28 St. Judes Rd, Englefield Green, Egham, Surrey, lists a variety of semiconductor devices together with many other components, Price 2s.

PASSIVE COMPONENTS

We have received the following brochures from Tygadure, Littlebrough, Lancs.
- Tygadure radio-frequency coaxial cables WW423
- Tygaflor equipment wires WW424
- Tygadure p.t.f.e. equipment wire, cable, sleeving, lacing cords and tapes, and glass braided yarn WW425

Engineering Bulletin No.7451 from Sprague is called "Beryllia-core silicone-coated Acrasil precision wirewound resistors". Sprague Electric (U.K.) Ltd, Sprague House, 159 High St, Yiewsley, West Drayton, Middlesex WW426

Swift Hardmans Wholesale Supply (S-O-T) Ltd, P.O. Box 23, Hardale House, Baillie St, Rochdale, Lancs, have produced a catalogue of Belling-Lee components (plugs, sockets, fuse holders and the like) they stock WW427

Precision miniature wirewound resistors with either radial or axial leads are the subject of a booklet available from Electrothermal Engineering Ltd, 270 Neville Rd, London E.7 .... WW428

B & R Relays Ltd, Temple Fields, Harlow, Essex, have produced a leaflet describing their D-range of relays which have contacts rated at 6A at 240V a.c. or 30V d.c. WW429

The following data sheets on components manufactured by the American Wilbrecht Company are available from J. H. Associates Ltd, 1 Church St, Bishop's Stortford, Herts.
- Model 170-S potentiometer with switch .. WW431
- Model 170-R potentiometer WW432
- Model 100 variable resistor (2.54mm diameter) WW433
- Model 2000 miniature slide switch WW434

APPLICATION NOTES

From Fairchild Semiconductor Ltd, Kingmaker House, Station Rd, New Barnet, Herts, a leaflet (No.6) describing the use of the Mμ L4102 16-bit associative memory cell and the Mμ L9035 64-bit read/write memory cell in a high-speed buffer memory system WW435

Application report B61 from WEL Components Ltd, 5 Loverock Rd, Reading, Berks, examines the construction of the triac and looks at its application in several circuits. The report includes notes on protection and working. Post and packing 2s.

Integrated Photomatrix Ltd, Grove Trading Estate, Dorchester, have produced an application note for the 50 × 1 light sensitive array mentioned in the Active Devices section.

EQUIPMENT

REL Equipment and Components Ltd, Microwave and Electronics Division, Croft House, Bancroft, Hitchin, Herts, have issued a short-form catalogue which lists a comprehensive range of test equipment WW436

A logic trainer from Limrose Electronics which consists of 16-NAND/NOR gates, or five-switch input register and four indicator lamps is described in a leaflet CK1/02 from Limrose Electronics, Lymm, Cheshire WW437

EMI Electronics Ltd, Television Equipment Division, Hayes, Middlesex, have produced a 44-page booklet on monochrome closed-circuit television systems for educational purposes WW438

A Mullard booklet, 'Do-it-yourself stereo', price 5s, gives constructional (wood-working) details of several possible stereophonic record reproducing assemblies using amplifier and power-supply modules (Mullard 'Unilex') that can be wired up using only a screwdriver.

Conferences and Exhibitions

Further details are obtainable from the addresses in parentheses

LONDON

Oct. 14-16 Savoy Place
Earth Station Technology
(I.E.E., Savoy Place, London WC2R 0BL)

Oct. 19-24 Olympia
Audio & Music Fair
(C. Rex-Hassan, 42 Manchester St., London W.1.)

BRIGHTON

Oct. 13-15 Hotel Metropole
INTER/NEPCON
(P. G. Saville, 21 Victoria Rd., Surbiton, Surrey)

MANCHESTER

Sept. 28-Oct. 2 Belle Vue
Electronics, Instruments and Components Show
(Inst. Electronics, 659 Oldham Road, Balderstone, Rochdale, Lancs)

OVERSEAS

Oct. 5-7 Argonne
Pattern Recognition
(Prof. S. S. Yau, Dept. of Electrical Eng., Northwestern University, Evanston, Illinois 60201)

Oct. 5-7 Rolla
Communications Conference
(J. R. Bretten, University of Missouri, 123 EE Bldg., Rolla, Missouri 65401)

Oct. 6-11 Ljubljana
Modern Electronics Exhibition
(Gospodarsko razstavisce, Ljubljana, Titova No. 50, Yugoslavia)

Oct. 7-9 Monticello
Circuit & Systems Theory
(G. Metze, University of Illinois, Urbana, Illinois 61801)

Oct. 12-16 Helsinki
British Engineering Week
(London Chamber of Commerce, 69 Cannon Street, London E.C.4)

Oct. 13-15 Los Angeles
Telemetering Conference
(International Foundation for Telemetering, 19730 Ventura Blvd., Woodland Hills, California 91364)

Oct. 14-16 Pittsburgh
Systems Science & Cybernetics
(I.E.E.E., 345 E. 47th St., New York, N.Y. 10017)

Oct. 26-28 Washington
Electronics and Aerospace Systems
(Dr. R. Marsten, NASA Headquarters, Code SC, Washington D.C. 20546)

Oct. 28-30 Washington
Electron Devices
(I.E.E.E., 345 E. 47th St., New York, N.Y. 10017)

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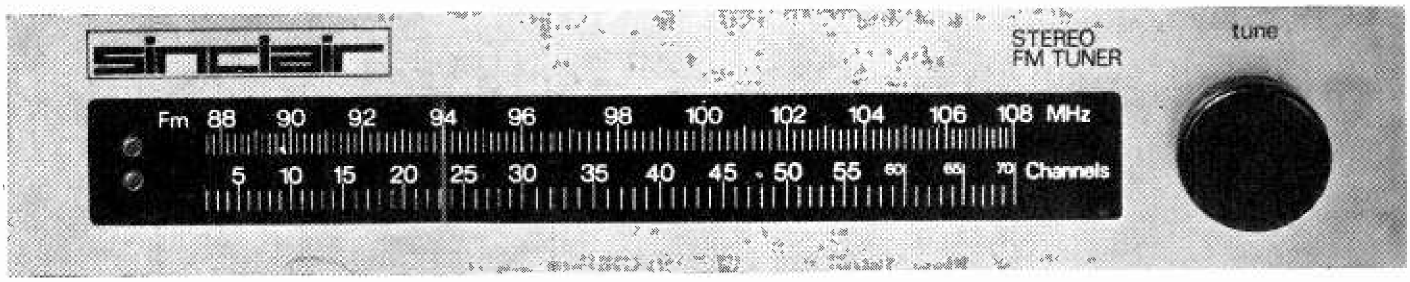
New for Project 60



the world's first high fidelity *phase lock loop* FM tuner

It has always been our policy at Sinclair Radionics to employ new and highly advanced circuitry in our products so that we can offer better performance at competitive prices. Our new F.M. tuner is the first in the World to use the phase lock loop principle. We have also incorporated such advanced features as varicap diodes for the tuning, printed circuit coils for the tuner and I.F. strip, A.G.C., A.F.C., an excellent squelch circuit to silence the tuner between stations, an Integrated Circuit stereo decoder and the option of remote control and push button switching.

The phase lock loop principle was first applied to receivers for reception from satellites because of the important improvements in signal to noise ratio that could be obtained by this technique. In addition there were the benefits of greatly improved selectivity and sensitivity. The Project 60 tuner, as the specifications show, is unsurpassed by any tuner now available yet we are able, because of the new circuitry, to sell the product at a fraction of the price.

From the high fidelity point of view this new circuit has the very important advantage of very much lower distortion than any other tuner known to us.

A voltage controlled oscillator (V.C.O.) in a phase lock loop tuner is kept in phase with the incoming signal by a phase comparator or detector which compares the two and feeds a control voltage to the oscillator. This control voltage is the audio output in the case of an F.M. signal. Since it is possible to design a V.C.O. which has an extremely linear voltage to frequency transfer characteristic excellent audio fidelity can be readily achieved. Furthermore, the oscillator can track a signal whilst completely rejecting a nearby stronger signal which would cause interference in a conventional receiver.

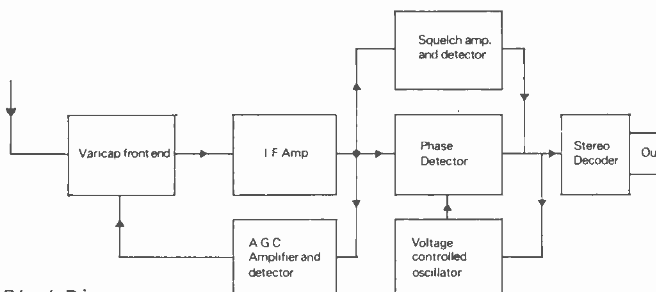
In use the tuner is especially attractive because the squelch circuit gives complete silence between stations and because fine tuning is accomplished automatically by the tuner. Accurate tuning is therefore ensured.

The use of an integrated circuit for the stereo decoder part of the circuit helps to give improved performance as it enables us to use a far more sophisticated circuit than would otherwise be possible. In particular stereo separation is excellent. Switching from mono to stereo is automatic and is indicated by a bulb.

The Project 60 tuner is supplied completely built and tested and ready to be mounted into any cabinet you choose. It may be used with any high fidelity amplifier including of course the Project 60 amplifier systems. The remarkable selectivity and sensitivity will make it possible to receive stereo transmissions in many more areas and foreign broadcasts will also be received far more readily. It is worth remembering that the Project 60 tuner will operate well on only a few inches of wire in most areas should this be necessary.

Project 60 F.M. tuner specifications

Number of transistors	16 plus 20 in I.C.
Tuning range	87.5 to 108 MHz.
Capture ratio	1.5dB
Sensitivity	2μV for 30dB quieting 7μV for full limiting
Squelch level	20μV
A.F.C. range	± 200 KHz
Signal to noise ratio	>65dB
Total harmonic distortion	0.15% for 30% modulatio
Stereo decoder operating level	2μV
Pilot tone suppression	30dB
Cross talk	40dB
I.F. frequency	10.7 MHz
Output voltage	2 x 150mV R.M.S.
Aerial Impedance	75 Ohms
Indicators	Mains on; Stereo on; tuning indicator



Block Diagram

Price: £25 built and tested. Post free.

Sinclair Radionics Ltd.

at the International Audio and Music Fair, Olympia, Stand **44**

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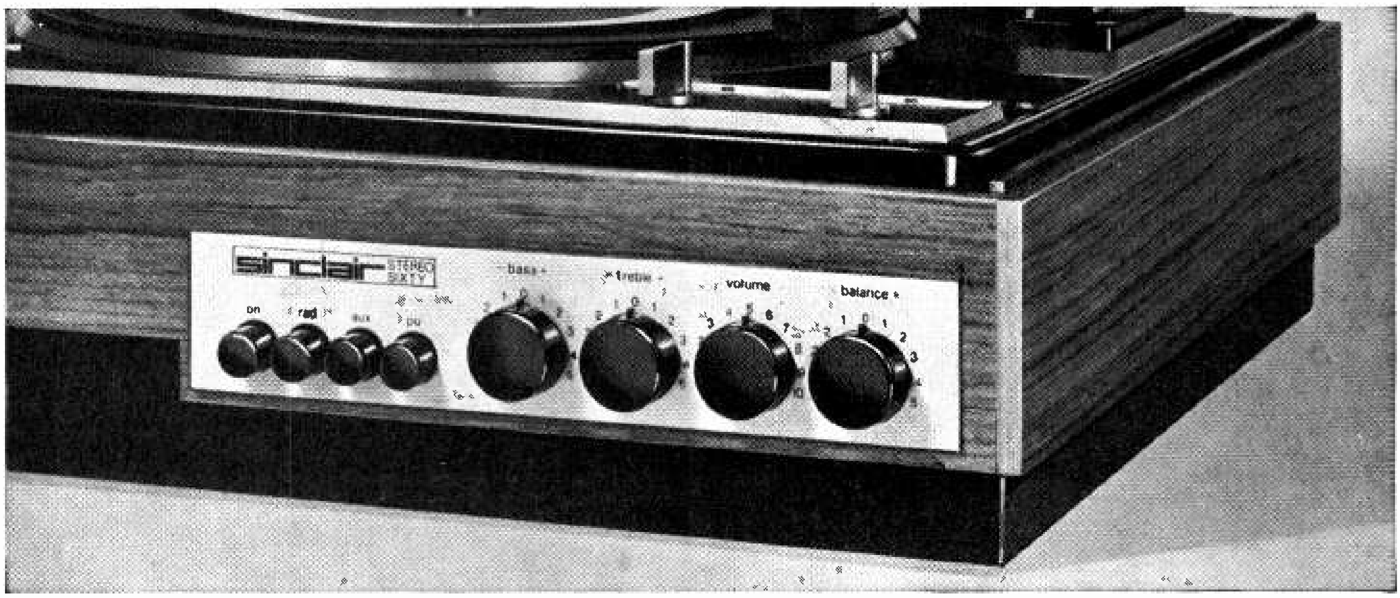
Project 60



Laboratory standard modular high fidelity

Sinclair Project 60 comprises a range of modules which connect together simply to form a compact stereo amplifier with really excellent performance. So good, in fact, that only 2 or 3 amplifiers in the world can compare in overall performance and now the constructor has choice of assemblies with either 20 or 40 watts output per channel, with or without filter facilities.

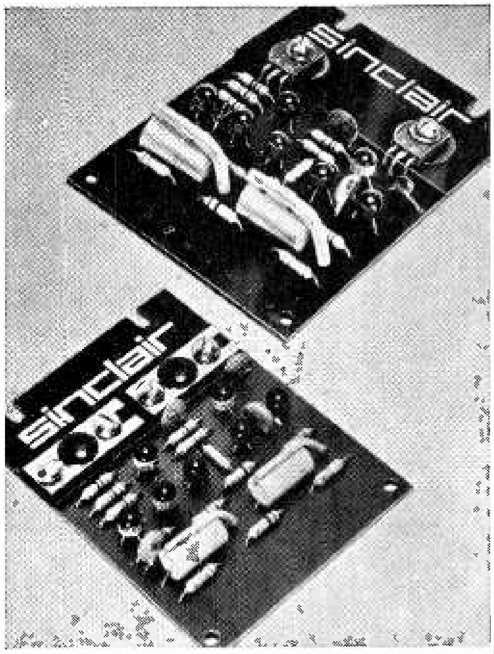
The modules are: **1.** The Z.30 and Z.50 high gain power amplifiers. **2.** The Stereo 60 preamplifier and control unit. **3.** The Active Filter Unit. **4.** 4 supply units–PZ.5; PZ.6; PZ.7 and PZ.8. In a normal domestic application, there will be no significant difference between PZ.5 or PZ.6 unless loudspeakers of very low efficiency are being used, in which case the PZ.6 will be required. For assemblies using two Z.50's there is the PZ.8 supply unit to ensure maximum performance from these amplifiers. No skill or experience are needed to build your system and the Project 60 manual gives all the instructions you can possibly want, clearly and concisely. Perhaps the greatest beauty of the system is that it is not only flexible now but will remain so in the future as new additions are made to the range. A stereo F.M. tuner is next to come. These and all other modules introduced will be compatible with those already available and may be added to your system at any time. And because Sinclair are the largest producers of constructor modules in Europe, Project 60 prices are remarkably low.

	System	The Units to use	In conjunction with	Your Project 60 Units will cost
A	Car Radio	Z.30	Existing car radio, Sinclair Micromatic	89/6
B	Simple battery powered record player	Z.30	Crystal pick-up, 12 V or more battery supply and volume control	89/6
C	Mains powered record player	Z.30 and PZ.5	Crystal or ceramic P.U. Vol. control etc.	£9.9.0
D	20+20 watts RMS stereo amplifier for most needs	Two Z.30s, Stereo 60 and PZ.5	Crystal, ceramic or magnetic P.U., most dynamic speakers, FM tuner, etc.	£23.18.0
E	20+20 watts RMS stereo amplifier for use with low efficiency (high performance) speakers	Two Z.30s, Stereo 60 and PZ.6	High quality ceramic or mag. P.U., F.M. Tuner, Tape Deck, etc. All dynamic spkrs.	£26.18.0
F	40+40 watts RMS de-luxe stereo amplifier	Two Z.50s, Stereo 60 PZ.8 and mains transformer	As for E	£32.17.6
G	Outdoor public address system	Z.50	Microphone, up to 4 P.A. speakers, 12V car battery with converter, or 45V d.c. controls	£5.9.6
H	Indoor P.A	One Z.50, PZ.8 and mains transformer	Mic., guitar, heavy duty speakers etc., controls	£17.8.6
J	High pass and low pass filters	AFU	D, E or F as above	£5.19.6
K	Stereo F.M. tuner	To be released shortly		

How to assemble and use Project 60 modules to best advantage in the above and other applications will be found in the fully descriptive Project 60 manual included with Project 60 systems. This 48 page manual is available separately, price 2/6d including postage.

sinclair SINCLAIR RADIONICS LTD., 22 NEWMARKET ROAD, CAMBRIDGE
Telephone 0223 52731

www.americanradiohistory.com



Z.30 & Z.50 POWER AMPLIFIERS

The Z.30 together with the Z.50 are both of advanced design using silicon epitaxial planar transistors to achieve unsurpassed standards of performance. Total harmonic distortion is an incredibly low 0.02% at full output and all lower outputs. Whether you use the Z.30 or Z.50 power amplifiers in your Project 60 system will depend on personal preference, but they are the same physical size and may be used with other units in the Project 60 range equally well. For operating from mains, for the Z.30 use PZ.5 for most domestic requirements, or PZ.6 if you have very low efficiency loudspeakers. For Z.50, use the PZ.8 described below.

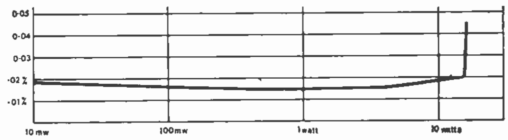
SPECIFICATIONS ***(Z.50 units are interchangeable with Z.30s in all applications.***

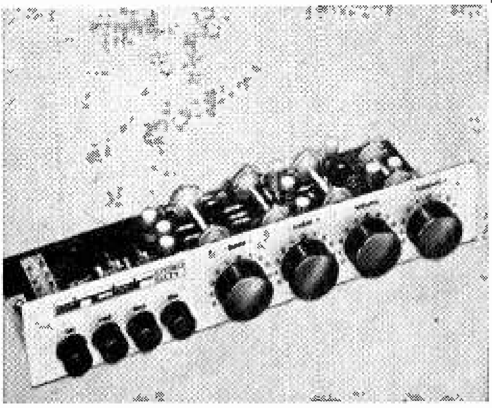
Power Outputs
Z.30 15 watts R.M.S. into 8 ohms, using 35V: 20 watts R.M.S. into 3 ohms using 30 volts.
Z.50 40 watts R.M.S. into 3 ohms from 40 volts: 30 watts R.M.S. into 8 ohms, using 50 volts.
Frequency response 30 to 300,000 Hz ± 1dB
Distortion 0.02% into 8 ohms
Signal to noise ratio better than 70 dB unweighted
Input sensitivity 250mV into 100 Kohms.
For speakers from 3 to 15 ohms impedance
Size $3\frac{1}{2}$ x $2\frac{1}{4}$ x $\frac{1}{2}$ ins.

Z.30
Built, tested and guaranteed with circuits and instructions manual **89/6**

Z.50
Built, tested and guaranteed with circuits and instructions manual **109/6**

Curve shows power versus distortion for Z.30 and Z.50.





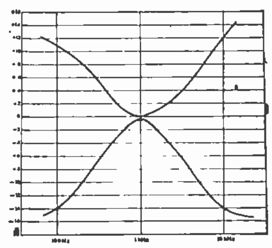
STEREO 60 Pre amp/Control Unit

Designed for the Project 60 range but suitable for use with any high quality power amplifier. Again silicon epitaxial planar transistors are used throughout, achieving a really high signal-to-noise ratio and excellent tracking between channels. Input selection is by means of push buttons and accurate equalisation is provided for all the usual inputs.

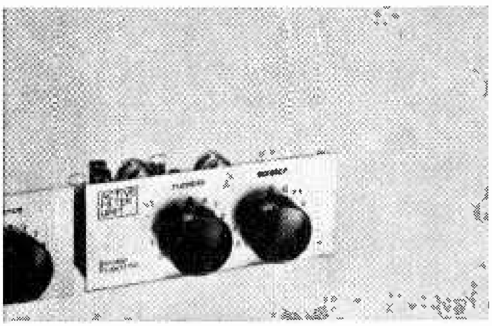
SPECIFICATIONS

- Input sensitivities – Radio – up to 3mV. Mag. p.u. – 3mV: correct to R.I.A.A. curve ± 1dB: 20 to 25,000Hz. Ceramic p.u. – up to 3mV: Aux. – up to 3mV.
- Output – 250mV.
- Signal-to-noise ratio – better than 70 dB.
- Channel matching – within 1 dB.
- Tone controls – TREBLE +15 to –15dB at 10kHz: BASS +15 to —15dB at 100Hz.
- Front panel – brushed aluminium with black knobs and controls.
- Size $8\frac{1}{4}$ x $1\frac{1}{2}$ x 4 ins.

Built, tested and guaranteed **£9.19.6**

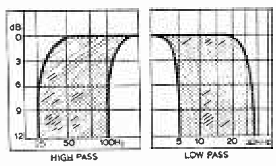


Curve to show bass and treble cut and boost.

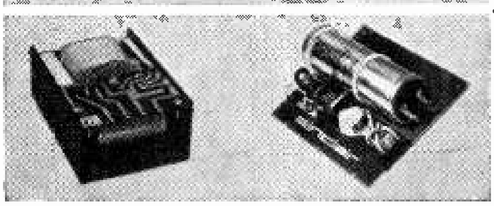


ACTIVE FILTER UNIT

For use between Stereo 60 unit and two Z.30s or Z.50s, the Active Filter Unit matches the Stereo 60 in styling and is as easily mounted. It is unique in that the cut-off frequencies are continuously variable, and as attenuation in the rejected band is rapid (12dB/octave), there is less loss of the wanted signal than has previously been possible. Amplitude and phase distortion are negligible. The Sinclair A.F.U. is suitable also for use with any other amplifier system.
Two stages of filtering are incorporated—rumble (high pass) and scratch (low pass). Supply voltage—15 to 35V. Current—3mA. H.F cut-off (—3dB) variable from 28kHz to 5kHz. L.F cut-off (—3dB) variable from 25Hz to 100Hz. Filter slope, both sections 12dB per octave. Distortion at 1kHz (35V supply) 0.02% at rated output.



Built, tested and guaranteed **£5.19.6**



POWER SUPPLY UNITS

The units below are designed specially for use with the Project 60 system of your choice. Illustration shows PZ.5 power supply unit to left and PZ.8 (for use with Z.50s) to the right. Use PZ.5 for normal Z.30 assemblies and PZ.6 where a stabilised supply is essential.

PZ-5 *30 volts unstabilised* **£4.19.6** **PZ-8** *45 volts stabilised (less mains transformers)* **£5.19.6**
PZ-6 *35 volts stabilised* **£7.19.6** **PZ-8** *mains transformer* **£5.19.6**

GUARANTEE *If within 3 months of purchasing Project 60 modules directly from us, you are dissatisfied with them, we will refund your money at once. Each module is guaranteed to work perfectly and should any defect arise in normal use we will service it at once and without any cost to you whatsoever provided that it is returned to us within 2 years of the purchase date. There will be a small charge for service thereafter. No charge for postage by surface mail Air-mail charged at cost*

sinclair

at the International Audio and Music Fair, Olympia Stand 44

To: SINCLAIR RADIONICS LTD., 22 NEWMARKET RD., CAMBRIDGE

Please send ....
....
....
....
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NAME ....
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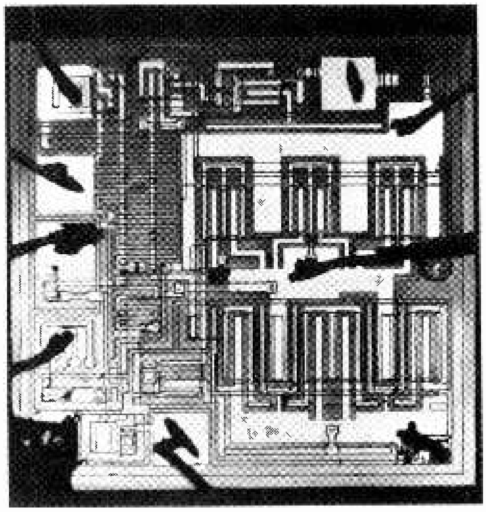
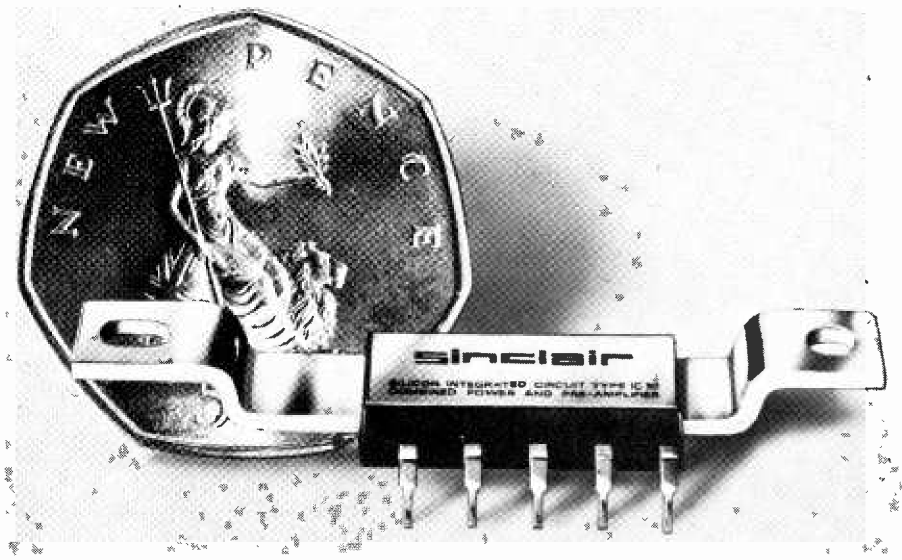
Sinclair IC-10





the world's most advanced high fidelity amplifier

Specifications

Output: 10 Watts peak, 5 Watts R.M.S. continuous
Frequency response: 5 Hz to 100 KHz±1dB
Total harmonic distortion: Less than 1% at full output.
Load impedance: 3 to 15 ohms.
Power gain: 110dB (100,000,000,000 times) total.
Supply voltage: 8 to 18 volts.
Size: 1 x 0.4 x 0.2 inches.
Sensitivity: 5mV.
Input impedance: Adjustable externally up to 2.5 M ohms.

Circuit Description

The first three transistors are used in the pre-amp and the remaining 10 in the power amplifier. Class AB output is used with closely controlled quiescent current which is independent of temperature. Generous negative feedback is used round both sections and the amplifier is completely free from crossover distortion at all supply voltages, making battery operation eminently satisfactory.

Applications

Each IC-10 is sold with a very comprehensive manual giving circuit and wiring diagrams for a large number of applications in addition to high fidelity. These include stabilised power supplies, oscillators, etc. The pre-amp section can be used as an R.F. or I.F. amplifier without any additional transistors.

The Sinclair IC-10 is the world's first monolithic integrated circuit high fidelity power amplifier and pre-amplifier. The circuit itself, a chip of silicon only a twentieth of an inch square by one hundredth of an inch thick, has 5 watts R.M.S. output (10w. peak). It contains 13 transistors (including two power types), 2 diodes, 1 zener diode and 18 resistors, formed simultaneously in the silicon by a series of diffusions. The chip is encapsulated in a solid plastic package which holds the metal heat sink and connecting pins. This exciting device is not only more rugged and reliable than any previous amplifier, it also has considerable performance advantages. The most important are complete freedom from thermal runaway due to the close thermal coupling between the output transistors and the bias diodes and very low level of distortion.

The IC-10 is primarily intended as a full performance high fidelity power and pre-amplifier, for which application it only requires the addition of such components as tone and volume controls and a battery or mains power supply. However, it is so designed that it may be used simply in many other applications including car radios, electronic organs, servo amplifiers (it is d.c. coupled throughout), etc. Once proven, the circuits can be produced with complete uniformity which enables us to give a full guarantee on every IC-10, knowing that every unit will work as perfectly as the original and do so for a lifetime.

SINCLAIR
IC-10 *with IC-10 manual Post free.* **59/6**

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At the International Audio and Music Fair, Olympia, Stand 44

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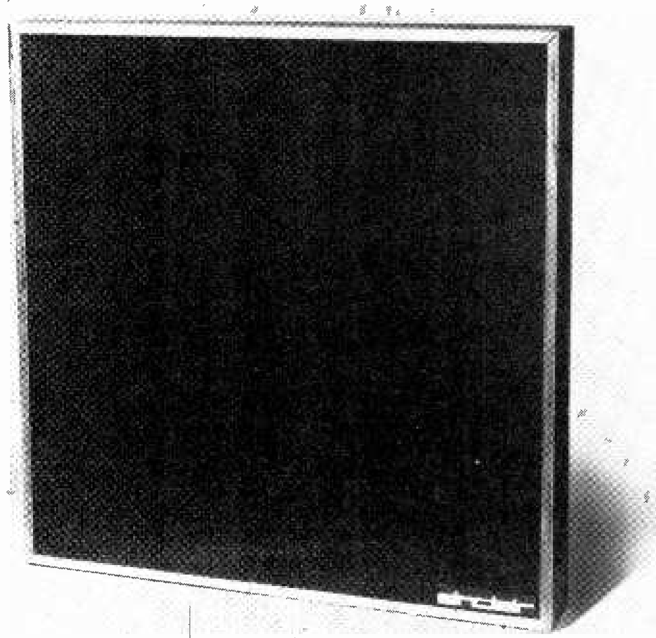
Q.16 High fidelity loudspeaker

Developed out of the revolutionary and much praised design of the original Sinclair Q.14 comes this more advanced version to meet the requirements of even greater numbers of high fidelity enthusiasts. The Q.16 employs the same well proven acoustic principles in which a special driver assembly is meticulously matched to the physical characteristics of the uniquely designed housing. In reviewing this exclusive Sinclair design, technical journals have been loud in their praise for it and it comfortably stands comparison with very much more expensive loudspeakers. The shape of the Q.16 enables it to be positioned and matched to its environment to much better effect than is the case with conventionally styled enclosures, and with its improved styling, the Q.16 presents an entirely new and attractive appearance. A solid teak surround is used with a special all-over cellular black foam front chosen as much for its appearance as for its ability to pass all audio frequencies unimpaired.

The Q.16 is compact and slim and is the ideal shelf-mounted speaker, and brings genuine high fidelity within reach of every music lover.

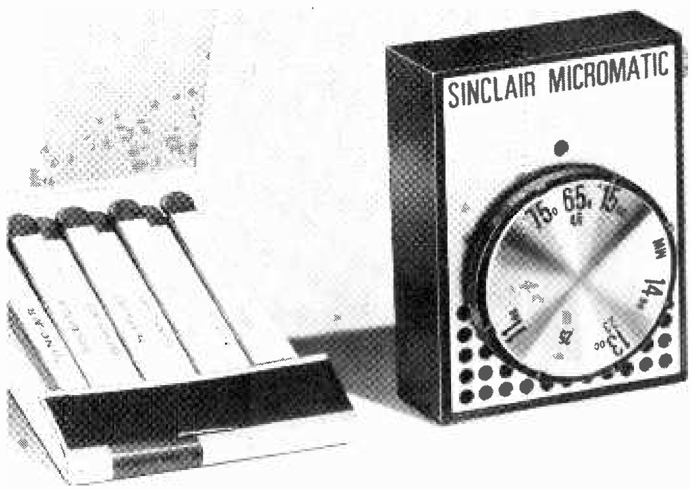


Specifications

Construction:	A sealed seamless sound or pressure chamber is used with internal baffle, all of materials carefully chosen to ensure freedom from spurious tone coloration.
Loading:	Up to 14 watts R.M.S.
Input impedance:	8 ohms.
Frequency response:	From 60 to 16,000Hz, as confirmed by independently plotted B & K curve.
Driver unit:	Specially designed high compliance unit having massive ceramic magnet of 11,000 gauss, aluminium speech coil and special cone suspension. Excellent transient response is achieved.
Size and styling:	$9\frac{3}{4}$" square on face × $4\frac{3}{4}$" deep with neat pedestal base. Black all-over cellular foam front with natural solid teak surround.
Price:	£8 19 6.

Micromatic Britain's smallest radio

Considerably smaller than an ordinary box of matches, this is a multi-stage A.M. receiver meticulously designed to provide remarkable standards of selectivity, power and quality. Powerful A.G.C. is incorporated to counteract fading from distant stations; bandspread at higher frequencies makes reception of Radio 1 easy at all times. Vernier type tuning plus the directional properties of the self-contained special ferrite rod aerial makes station separation very much easier than with many larger sets. The plug-in high fidelity type magnetic earpiece which matches exactly with the output of the Micromatic provides wonderful standards of reproduction both for speech and for music. Everything including the batteries is contained within the attractively designed case. Whether you build your Micromatic or buy it ready built and tested, you will find it as easy to take with you as your wristwatch, and dependable under the severest listening conditions.



Specifications

Size:	$1\frac{13}{16}$" × $1\frac{7}{16}$" × $\frac{1}{2}$" (46 × 33 × 13mm).
Weight including batteries:	1 oz. (28.35gm) approx.
Tuning:	Medium wave band with bandspread at higher frequency end.
Earpiece:	High-fidelity magnetic type.
Battery requirements:	Two Mallory Mercury Cells, type R.M. 675, for long working life.
Case:	Black plastic with anodised aluminium front panel, spun aluminium dial.
Controls:	Tuning dial, and on/off switching by means of earpiece plug.
Price:	Available in kit form complete with earpiece, case, instructions and supply of solder in fitted pack. 49/6. Ready built, tested and guaranteed. 59/6.

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BFX88 5/-
BFY50 4/6
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BSY28 5/-
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OC28 12/6
OC29 12/6
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OC44 4/-
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OC74 6/-
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OC76 5/-
OC77 8/-
OC78 5/-
OC81 5/-
OC81D 4/-
OC82 5/-
OC83 5/-
OC84 5/-
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OC123 10/-
OC139 5/-
OC140 7/6
OC141 15/-
OC169 4/-
OC170 5/-
OC171 6/-
OC200 7/6
OC201 9/6
OC202 12/6
OC203 7/6
OC204 8/-
OC205 12/6
OC206 15/-
OC207 15/-
OCP71 19/6
ORP12 10/-
ORP60 8/-

INTEGRATED CIRCUITS

Type	1-11	12+	Type	1-11	12+
UL914	8/6	7/9	CA3014	29/6	27/-
CA3048	45/-	42/-	CA3018	22/6	20/-
SL403A	42/6	41/-	CA3020	27/6	25/-
MC1303	52/6	48/-	CA3028	24/6	22/6
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Mullard 115 watt Silicon Power
25+13/- 100+11/-

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Texas F.E.T.
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500+4/9

2N2926 2/6
NPN Planar
All colours
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Motorola Unijunction
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BY126 3/-
Mullard 800v. 1 amp. plastic
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BY127 4/-
Mullard 1000v. 1 amp. plastic
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BYZ13 4/-
Mullard 6A. 200v.
25+3/6 100+3/2
500+2/10

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Mullard Power
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500+1/1

OCP71 19/6
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OC45 3/6
Mullard
25+3/- 100+2/6
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OC71 3/-
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500+1/9

OC75 5/-
Mullard
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500+3/-

BCY34 5/-
Mullard
25+4/3 100+3/9
500+3/6

OC20 19/6
Mullard 100v.
25+15/9 100+14/6
500+13/3

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1 amp. 100-300v.
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All voltages 3.3v.-33v. 3/6
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Any one type

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800-1000v. 1 amp.
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OC139 5/-
Mullard
25+4/- 100+3/3
500+3/-

OC140 7/6
Mullard
25+6/- 100+5/-
500+4/-

PLEASE NOTE MINIMUM ORDER 10/-

MULLARD INTEGRATED CIRCUITS

	1-9	10-24		1-9	10-24
FJH101	14/-	12/-	FJH151	14/-	12/-
FJH111	14/-	12/-	FJH161	14/-	12/-
FJH121	14/-	12/-	FJH171	14/-	12/-
FJH131	14/-	12/-	FJH181	14/-	12/-
FJH141	14/-	12/-	FJH221	14/-	12/-
FJH231	14/-	12/-	FJJ111	19/-	16/-
FJJ121	32/-	27/6	FJJ191	32/-	27/6
FJJ131	28/9	25/-	FJJ251	56/6	48/6
FJJ141	56/6	48/6	FJL101	56/6	48/6
FJJ211	56/6	48/6	FJY101	77/6	67/6

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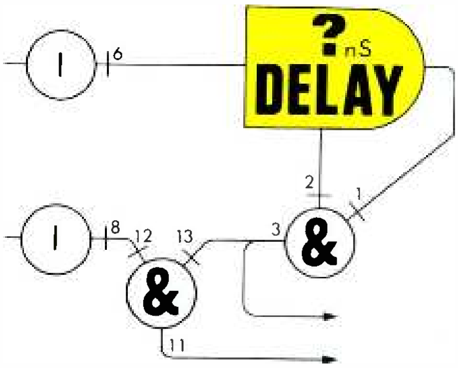
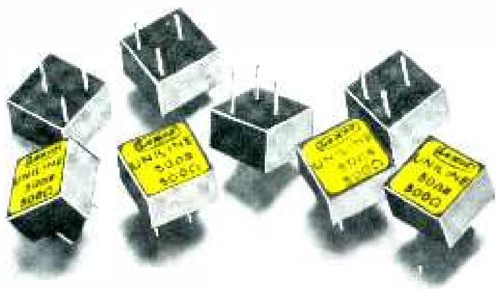


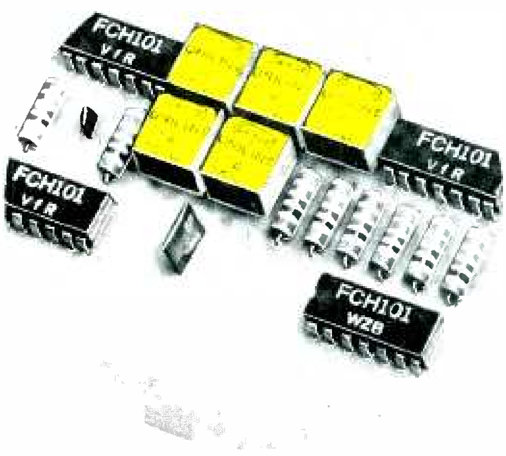
PROBLEM

what delay value?



SOLUTION

use Lexor cascadable modules



Lexor

uniline

build-a-delay system

Engineers with a delay problem will swiftly recognise the value and flexibility of Lexor **uniline** cascadable modules.
The wide range of incremental values makes delay building simple, quick and economical. An exact delay required for a circuit can be established instantly in-situ.

Lexor THE DELAY LINE SPECIALISTS

LEXOR ELECTRONICS LTD.
25/31 Allesley Old Road, Coventry.
Telephone: Coventry 72614 & 72207

With the **uniline** system the required delay value can be achieved by adding or subtracting modules in steps as small as 5nS. These miniature encapsulated units are designed to mount directly on to printed circuit boards on a 0·1" matrix and their small uniform size allows them to be positioned in any configuration which can be accommodated by an existing circuit layout.

They are available from Stock.

Having resolved the problem of delay value, take further advantage of the Lexor service by ordering—
Sets of **uniline** modules made up in composite blocks or a custom-built single unit to an equivalent value in any form required.

Leaflet L2 gives full technical specification and application notes.
KWP L2

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ERSIN Multicore 5 CORE SOLDER

OVER 400 SPECIFICATIONS USED IN MORE THAN 63 COUNTRIES FOR FAST RELIABLE SOLDERED JOINTS

- Contains 5 cores of non-corrosive high speed Ersin flux. Removes surface oxides and prevents their formation during soldering. Complies with B.S. 219, 441, DTD 599A, Din 1707, U.S. Spec. QQ-S-571d.
- Savbit an exclusive Multicore Alloy which is saturated with copper to prevent absorption of copper from copper wires, circuit boards and soldering iron bits. Ministry approved under Ref: DTD 900/4535.
- Solder Tape, Rings, Preforms and Washers, Cored or Solid, are available in a wide range of specifications.

STANDARD ALLOYS INCLUDE

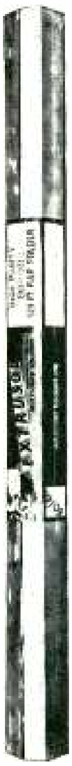
TIN/LEAD	B.S. GRADE	LIQUIDUS MELTING TEMP °C.	LIQUIDUS MELTING TEMP °F.
60/40	K	188	370
Savbit No 1	—	215	419
50/50	F	212	414
45/55	R	224	435
40/60	G	234	453
30/70	J	255	491
20/80	V	275	527

HIGH AND LOW MELTING POINT ALLOYS

ALLOY	DESCRIPTION	MELTING TEMP. °C.	MELTING TEMP. °F.
T.L.C.	Tin/Lead/Cadmium with very low melting point	145	293
L.M.P.	Contains 2% Silver for soldering silver coated surfaces	179	354
P.T.	Made from Pure Tin for use when a lead free solder is essential	232	450
H.M.P	High melting point solder to B.S. Grade 5S	296-301	565-574

COMPATIBLE PRINTED CIRCUIT SOLDERING MATERIALS

EXTRUSOL



High Purity Extruded Solder

provides the most economical soldering. Its high purity and freedom from oxides, sulphides and other undesirable elements result in the following advantages:—

- Less dross on initial melting.
- More soldered joints per pound of solder purchased.
- Less reject joints.
- Improved wetting of electronic components & printed circuit boards.
- More uniform results.

All Extrusol is completely protected by plastic film packaging from the moment of manufacture until it is used. Available in bars and pellets. Can be released under AID authority and supplied to USA QQ-S-571d.

PC.2 MULTICORE TARNISH REMOVER

removes tarnishes and inorganic residues as the second half of a pre-cleaning process before soldering. It leaves the copper unaffected.

PC.90 MULTICORE PEELOFF SOLDER RESIST

is a temporary solder resist which can be peeled off with tweezers after soldering, leaving the original clean surface. It can be used for masking gold plated edge connections and holes to which heat sensitive or other components must be added later.

PC.41 MULTICORE ANTI-OXIDANT SOLDER COVER

which forms a liquid cover on the solder bath either side of the solder wave, largely preventing the formation of dross.

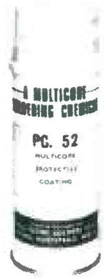
PC.80 MULTICORE SOLVENT CLEANER

removes organic contaminants such as grease, perspiration and residues of organic solutions from prior processes, as a pre-cleaning process before soldering. It is also very efficient in removing rosin-based flux residues after soldering.

PC.10A MULTICO ACTIVATED SURFACE PRESERVATIVE

is a pre-soldering coating for preserving the clean surfaces established by the PC.80 Multicore Solvent Cleaner and PC.2 Multicore Tarnish Remover. PC.10A does not need to be removed before soldering and in fact contributes to the efficiency of the soldering process. PC.10A should be used whenever there is a delay between cleaning and soldering.

Gallon Containers
All liquid chemicals and fluxes supplied in 1 gallon polythene 'easy pouring' containers, with carrying handle.



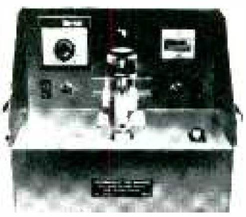
Aerosols
PC.21A, PC.10A and PC.52 available in 16 oz aerosol sprays.

SEVEN STANDARD MULTICORE LIQUID FLUXES

are now available, five of which are new:— PC.21A Multicore Non-Corrosive Liquid Flux is normally recommended for wave, dip, brush, spray and roller flux application methods. PC.25 Multicore Rosin Foam Flux is designed for foam fluxing and exhibits an unusually stable foam with a fine bubble size.

PC.52 MULTICORE PROTECTIVE COATING

is a lacquer which should be applied after soldering for protecting printed circuits from deterioration or failure in service. It can easily be soldered through if modifications or repairs are necessary at a later date.



SOLDERABILITY TEST MACHINE MARK 3

Use for testing to B.S 4393: 1969, Section 10. A simple precision instrument for assessing the solder-ability of component termination wires. Complies with B.S 2011 Part 2 Test T and comparable international standards. Essential for quality control.

SOLDERING HANDBOOK

The most comprehensive book on soldering for industrial use, containing 120 pages with 100 illustrations and invaluable reference charts. Features practical and theoretical methods of soldering in electronics and allied industries, and is divided into three headings.

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