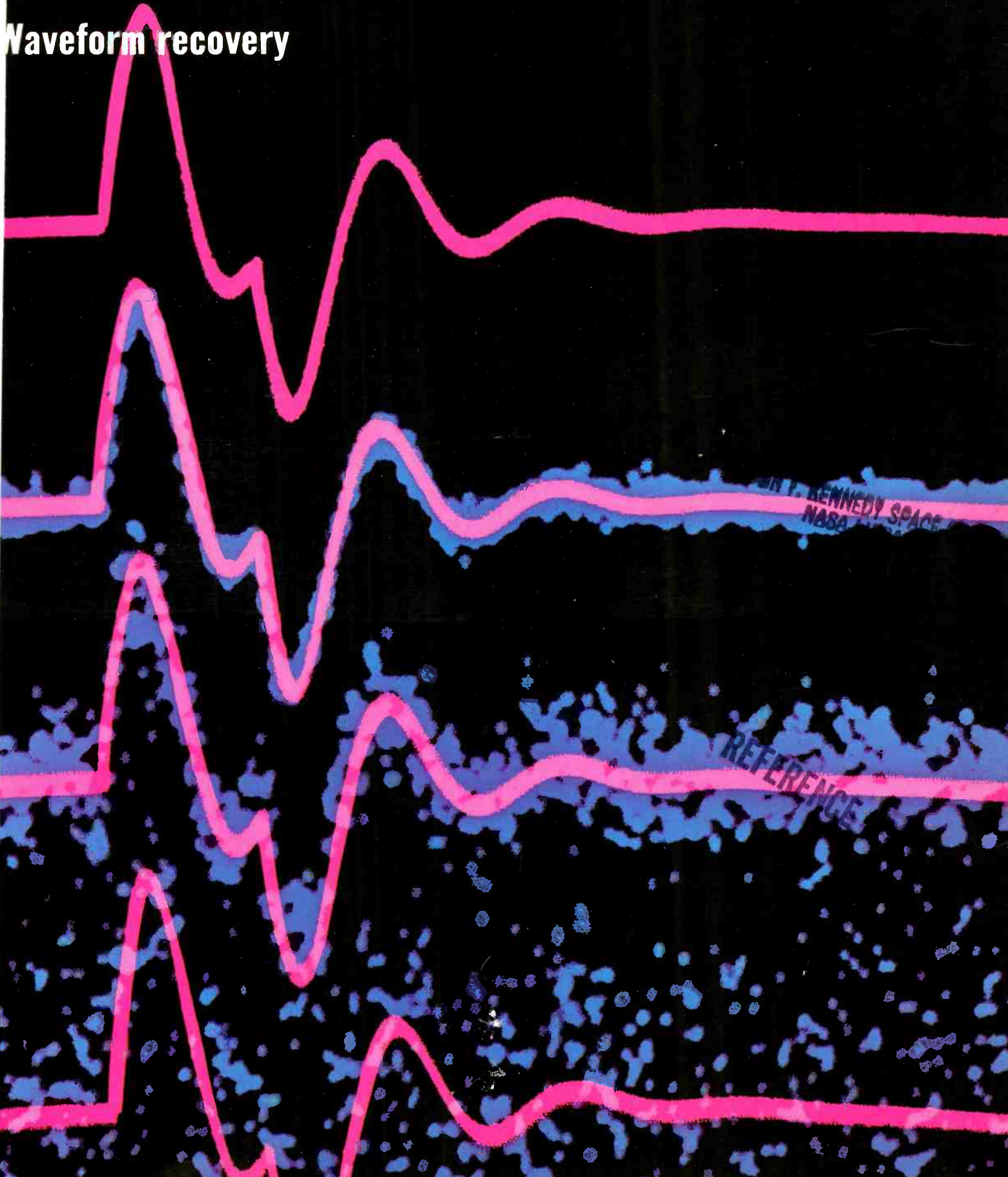
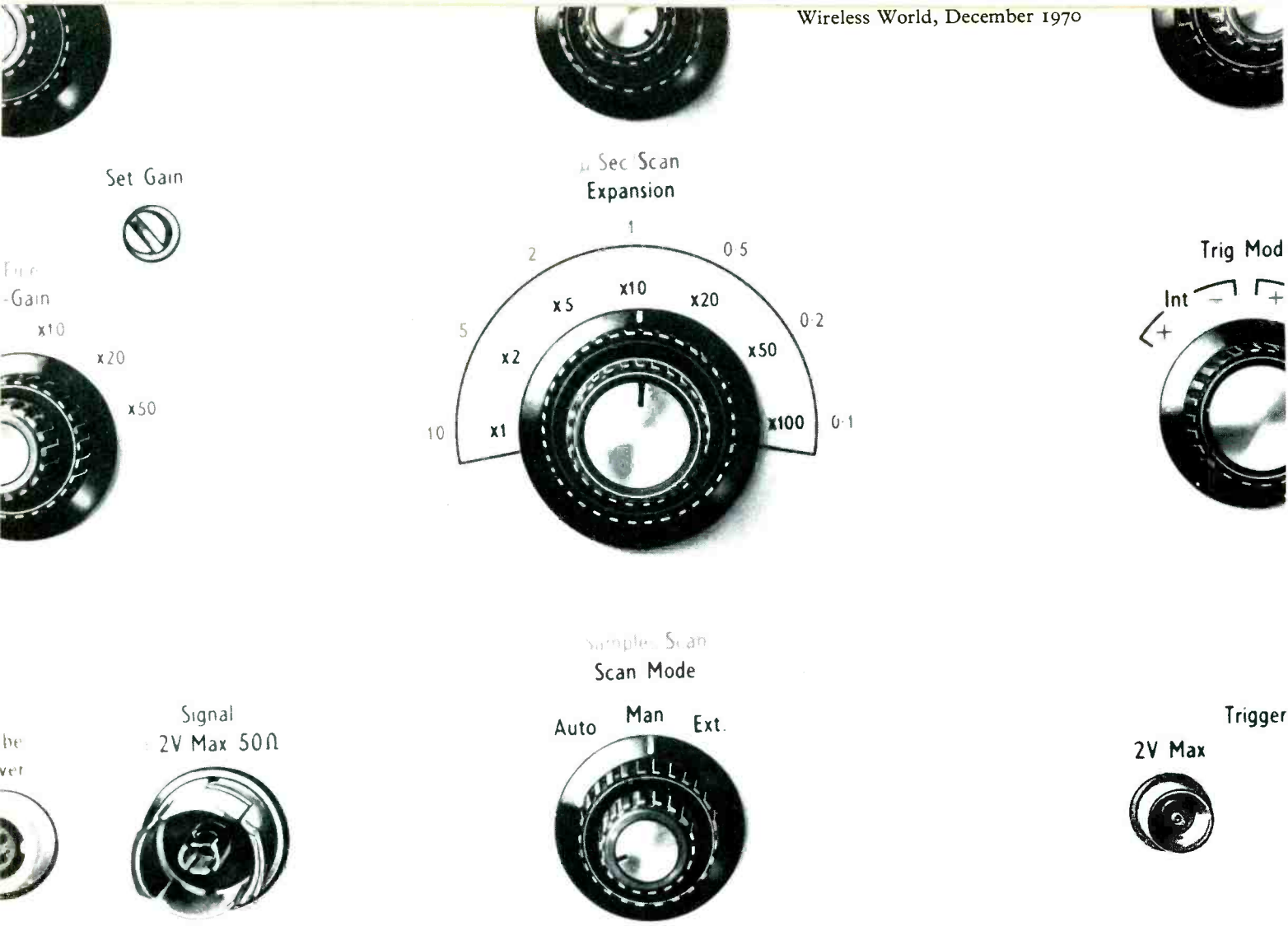


Wireless World

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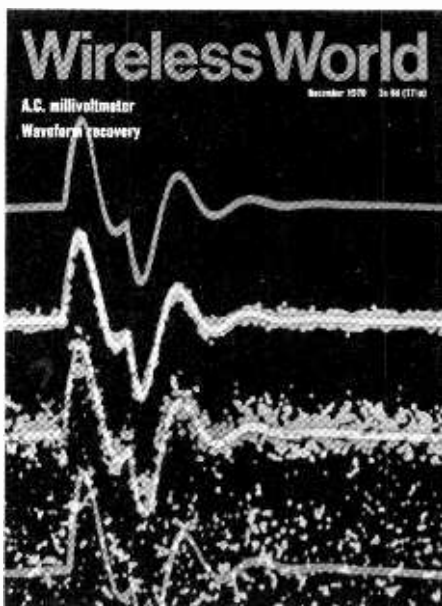
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December 1970

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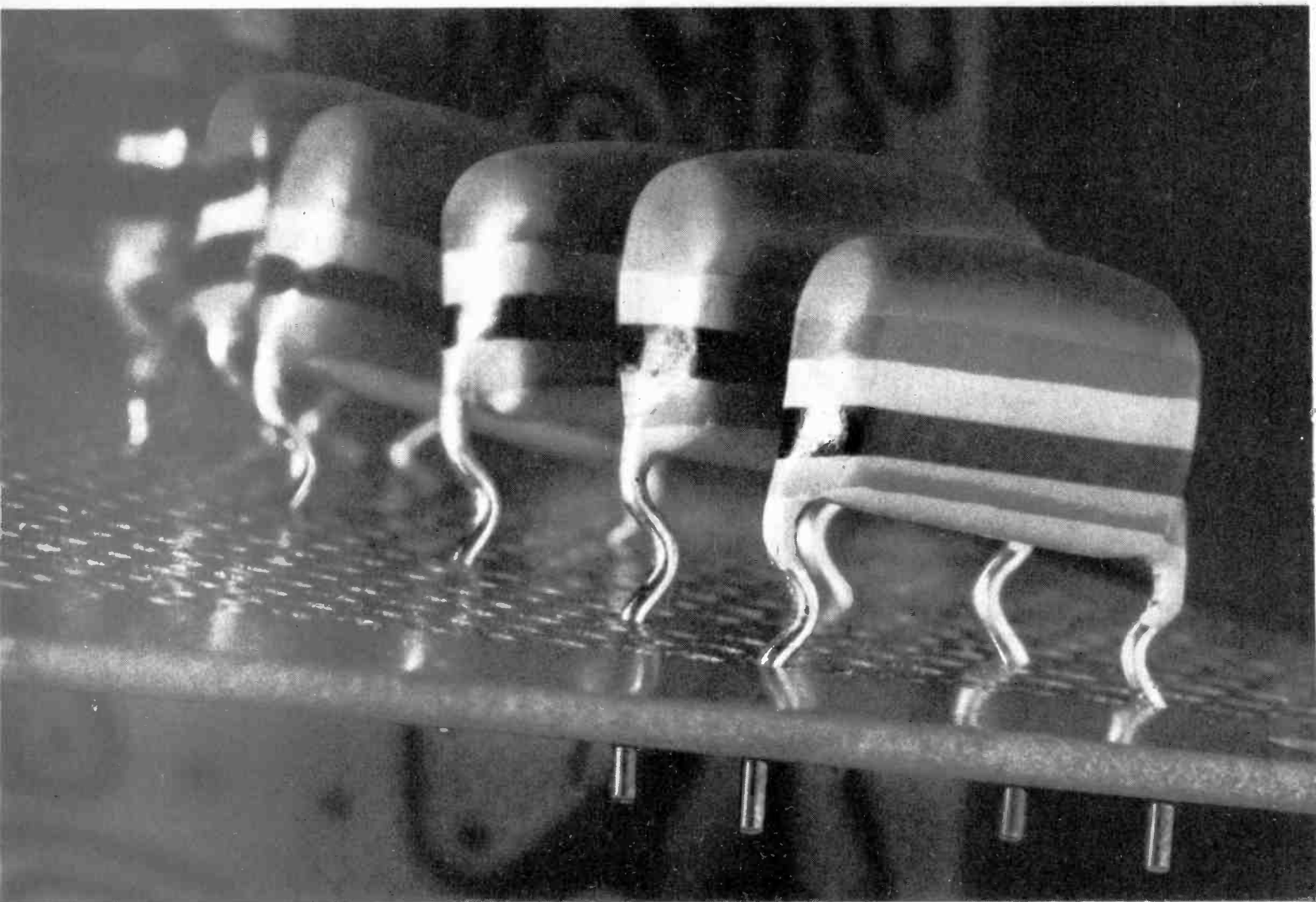


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Sound Sense

Elsewhere in this issue we review the annual London Audio Fair held at the end of October. The Fair has certainly come of age and can now take its place, in its own right, in the calendar of events in the capital. It has, of course, developed as a public show and makes no pretence of assuming the position once held by the annual show of the British Sound Recording Association. This was the Mecca of audio engineers and audiophiles who were concerned primarily with the finer points of sound recording and reproduction techniques. The emphasis now is on the resulting audio output, with the listener's ears as the final arbiter rather than the technical specification. In saying this we have no intention of belittling either the efforts of the Fair organizers or of the manufacturers participating; it is a 'fair' and as such displays a wide variety of wares and includes sideshows to delight the visitors.

There are, however, rumours that it may develop into a 'home entertainments' show. The word 'music' was added to its title this year but this bait did not bring the big fish of the music instrument world into the Fair. Our opinion is that the introduction of equipment not coming under the umbrella of 'audio' (which now has a definite connotation) would in fact weaken rather than strengthen the position of the Fair.

Inevitably the question was asked by participants and visitors to Olympia "is there room in London for two audio shows a year—Sonex in the spring and the Fair in the autumn?" Our unequivocal answer is "yes". Each has a definite role to fill. Sonex is primarily a trade show—with sales offices adjacent to dem. rooms—whereas the Audio Fair is a public spectacle—*son et lumiere!*

Neither show, however, has taken the place in the affections of the audio fraternity once held by the old B.S.R.A. Show. The addition this year of lecture demonstrations at the Audio Fair has, we believe, gone a long way to remedying one of the major deficiencies. If the standard of the lectures can be maintained in coming years they could provide a very worthwhile meeting place for the audio fraternity of this country.

The mention of the audio fraternity brings to mind the question of an association catering specifically for this section of engineers and specialists. There is, of course, the British Kinematography, Sound & Television Society which added the word 'sound' to its title and took the audio engineer into its fold at the demise of the B.S.R.A. This addition, and subsequently that of the word 'television', tended to broaden the scope of what was a kinematographers' association into an all-embracing organization rather than one of specialization. As will have been seen from the titles of lectures included in our meetings section month by month B.K.S.T.S. has covered a very wide variety of audio topics over the years. So far this association has been the main meeting place for audio engineers.

Some time ago we announced that a move was afoot to form a U.K. section of the American Audio Engineering Society. It is understood that plans are now well advanced.

Any move which will encourage the growth of interest in, and development of, audio techniques is to be applauded but is the proliferation of 'learned societies' the best way to achieve this? Certainly the position attained by the A.E.S. in America is very high. It remains to be seen what will happen in this country if the A.E.S. gets underway here.

The Boxcar Detector

A synchronous detector which is used to recover waveforms buried in noise

by J. D. W. Abernethy, B.A.

The boxcar detector* is a signal recovery instrument which is used either to retrieve the waveform of a repetitive signal from noise or to measure the amplitude of a repetitive pulse buried in a noise. It has two modes of operations, 'scan' and 'single-point', the former being used for waveform retrieval and the latter for pulse measurement. Waveform retrieval applications include on-line impulse testing of electro-mechanical systems, measuring the decay time of phosphors, time-of-flight mass spectroscopy and the measurement of magnetic field penetration in superconductors. In the single-point mode the boxcar will measure noise-obscured pulses such as those found in applications concerned with pulsed radar, acoustic paramagnetic resonance, Q-switched lasers, non-linear optical effects etc. The boxcar can also be used to make accurate permanent records of waveforms, even when

noise is not a problem, because of its ability to readout on X-Y plotters.

As an example of an impulse testing application, the boxcar may be used to determine the response of a loudspeaker and its surroundings without taking any precautions to screen the system from outside noise. In a typical experiment (see Fig. 1) a repetitive pulse is applied to the loudspeaker and the resulting signal, together with extraneous noise, is picked up by the microphone and fed into the boxcar. The boxcar is also supplied with a synchronous reference pulse. The output of the boxcar, an internally generated ramp and the modified microphone signal, is applied to an X-Y plotter which reproduces the input waveform with the noise reduced.

The boxcar is basically a sample-and-hold system the sampling time of which is determined by a reference pulse which is

related to the signal of interest. In its single-point mode the boxcar so arranges the timing and width of the sampling 'window' that it coincides with the signal pulse. Thus only the pulse and the noise occurring during the same period of time contribute to the output. See Fig. 2(a). In the scan mode, see Fig. 2 (b), the operation is somewhat similar except that the sampling window, instead of having a fixed time relation to the signal of interest, is slowly swept across it, thus producing at the output a lengthened replica of the input waveform.

Waveform averaging

The boxcar plots out the waveform of a repetitive signal buried in noise by correlating successive parts of the input waveform with a reference pulse and averaging the result with a low-pass filter.



James Abernethy was born in 1936 and was educated at Eton from 1950 to 1955 and at Balliol College, Oxford, where he gained a first in engineering. Before joining Brookdeal Electronics Ltd (where he is now chief engineer) two and a half years ago he served an apprenticeship with B.A.C. at Weybridge, subsequently spending two years in the Dynamics Department. Following this, he spent two years with Rolls Royce and Associates at Derby, working on nuclear submarine power plants.

*The origin of the word boxcar is somewhat obscure but it is thought to be derived from the similarity in outline of an American boxcar train and a train of pulses viewed on an oscilloscope.

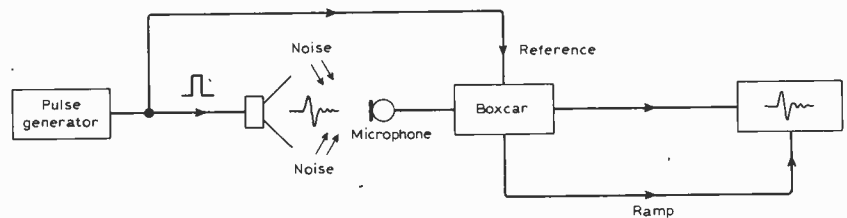


Fig. 1. A typical application. The measurement of acoustic response.

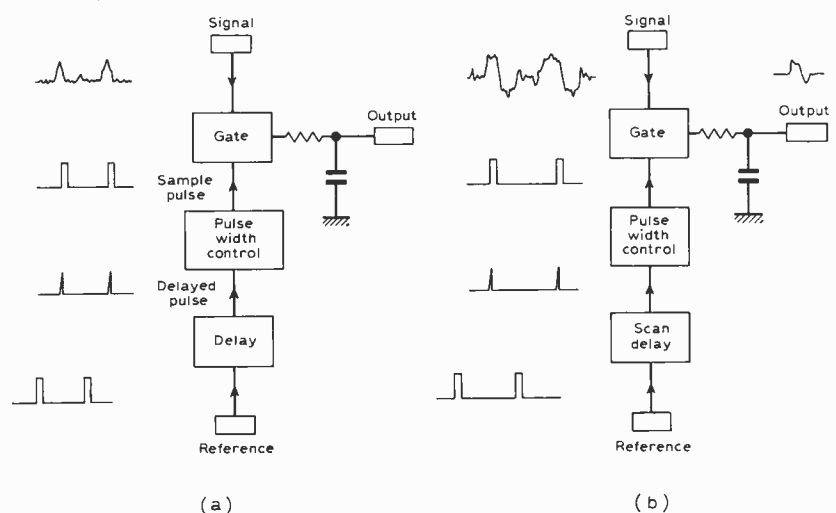


Fig. 2. Block diagrams (a) single point operation (b) scan mode.

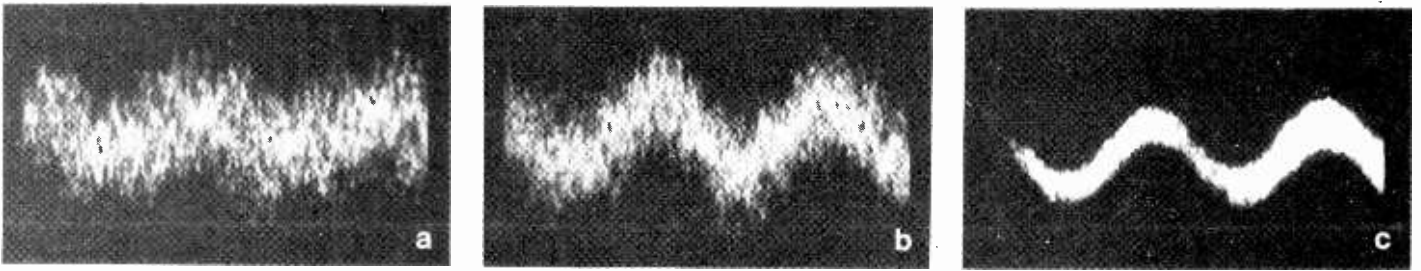


Fig. 3. A demonstration of correlation and averaging as a means of waveform recovery: (a) signal plus noise displayed on a short persistence oscilloscope using the internal timebase; (b) the same display with the timebase externally triggered by a reference waveform which is coherent with the signal; (c) this is the same as (b) with the c.r.t. replaced with one of long persistence.

As a simple example of waveform recovery by correlation and averaging, three oscilloscope traces of a repetitive signal and noise are shown in Fig. 3. In (a) the signal and noise are displayed on a short persistence oscilloscope with the timebase triggered internally. In addition to vertical errors caused by the noise there are also horizontal noise errors. In (b) the timebase is triggered externally by a reference which is coherent with the signal. It can be seen that this simple correlation process considerably improves the observed signal/noise ratio. In (c) the timebase is also triggered externally, but the tube persistence is increased. The improvement from (b) to (c) is due to the increase in the optical time constant.

The example is given to illustrate the principle of correlation and averaging as applied to waveform recovery, and it is not intended to be a serious solution to the problem. However, the principle, somewhat extended, leads to the idea of the multi-channel averager and from there to the functioning of a boxcar.

Multi-channel averager

The multi-channel averager is another correlation instrument which is used specifically for recovering the waveform of a repetitive signal buried in noise. It operates by using a series of sampling gates (see Fig. 4) to divide each signal cycle into segments, the number depending on the resolution required. Storage capacitors connected to the gates charge, on each cycle, towards the mean voltages of the appropriate segments. The gates are triggered by a reference which is synchronous with the signal of interest and thus noise inputs which are not coherent with the reference are averaged towards zero. Because of its complexity (200 channels or more may be required to give sufficient resolution) the multi-channel averager is generally very expensive. A further consequence of its complexity is that the sampling gates need to be fairly simple and therefore are not always capable of the required sampling speed. The boxcar operated in the scan mode provides a much simpler and therefore less expensive instrument with fast-sampling capability.

Boxcar-scan mode

When the boxcar is operated in the scan mode its function is to sample and average each 'point' on the waveform in turn,

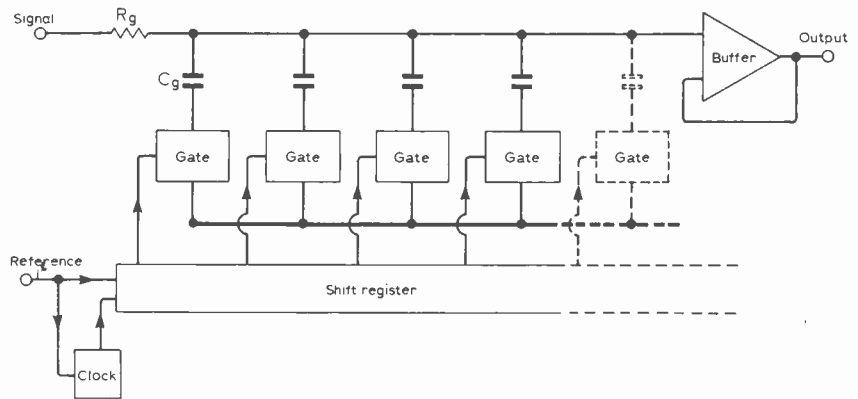
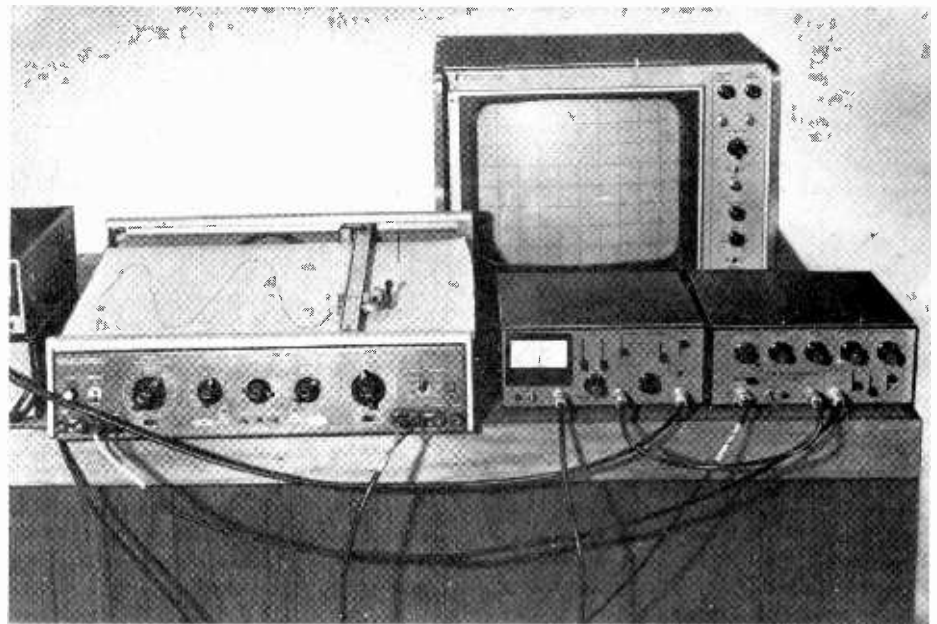


Fig. 4. Block diagram of multi-channel averager.



The boxcar detector in use.

moving to the next when the output has reached the voltage of the waveform at that point with sufficient accuracy. Scanning is achieved by changing the delay of the gate at a predetermined rate; called the scan rate. At the same time as the delay is scanned an output voltage is produced called the 'scan ramp output' which is proportional to the change in delay and provides a drive for the time axis of an X-Y plotter. A complete schematic diagram of the scan mode boxcar in operation is shown in Fig. 5. In practice the scan is continuous but, in order to make the illustration clearer, the scan is shown here as being in discrete steps, as if

the delay and X-Y plotter were controlled by a staircase ramp.

It can be seen that the boxcar wastes information, that is to say it ignores most of the waveform, sampling only a small section per cycle. Therefore by comparison with a multi-channel averager operating with a resolution equivalent to *n* gates the boxcar will take *n* times as long to plot out a waveform with equal noise reduction. In normal practice this is a disadvantage only in low-speed experiments and in particular in medical experiments with low repetition rates: where an averager with 100 channels might take, for example, one minute to produce a result the boxcar

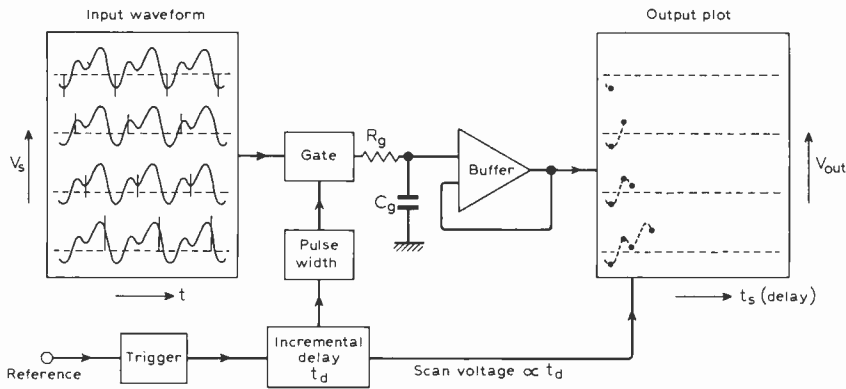


Fig. 5. (above) Point-to-point waveform averaging using the boxcar detector.

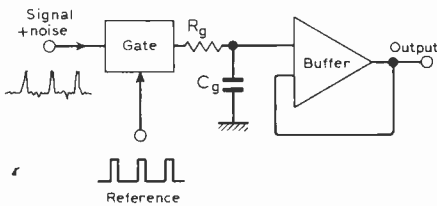


Fig. 6. (left) Sample and hold circuit.

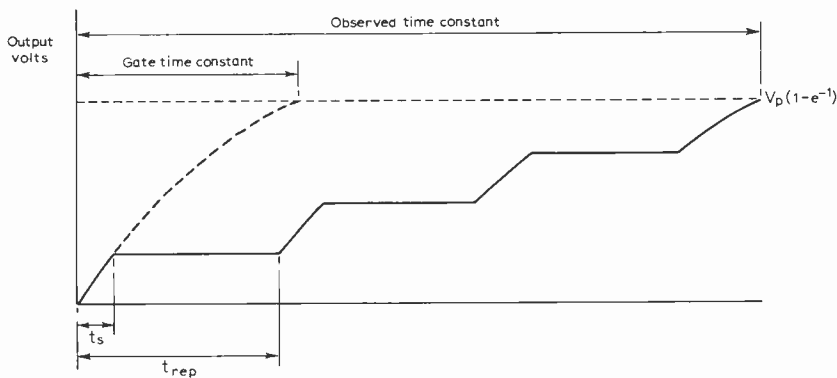


Fig. 7. (below) The charging curve of C_g .

would take two hours. However in high-frequency experiments, for which the boxcar is ideally suited by virtue of its fast sampling capability, the result is usually plotted out in a matter of seconds.

Sample-and-hold system

The sampling section of the boxcar has to perform two functions: it must average the section of the waveform which occurs during the sampling period and it must hold the result until the next sample is taken. In order to do this the sample-and-hold configuration shown in Fig. 6 is used. When sampling, the gate is opened and the capacitor C_g charges towards the mean

value of the sampled signal-plus-noise at a rate dependent on the product $R_g C_g$ (called the 'gate time constant' t_g). Thus the time taken for the capacitor voltage to approach the signal voltage will be greater as t_g is increased; correspondingly, the voltage fluctuations on the capacitor due to noise inputs will be reduced and thus, although the measurement will take longer, the signal/noise ratio will be improved by increasing t_g .

Clearly the factor by which t_g is related to the overall response time is important since this establishes the terms of the trade-off between noise reduction and observation time. This relationship can be seen in Fig. 7 which shows the charging curve of

C_g when a step-input (V_p) is applied and the reference consists of pulses giving a 'gate time' t_s occurring at intervals t_{rep} . It can be seen that during the time the gate is open the charging time constant is $R_g C_g$ whereas at the output there is a longer response time called the 'observed time constant', due to the fact that the charging circuitry is inactive for a large proportion of the time. The observed time constant (t_{obs}) is given by:

$$t_{obs} = t_g \times t_{rep} / t_s$$

The noise averaging properties of the gate time constant are illustrated in Fig. 8 where oscilloscope traces show the voltage on C_g when a step function plus noise is applied to the signal input for (a) $t_g \ll t_s$ and (b) $t_g > t_s$. As one would expect the noise in (b) is reduced by the longer gate time constant. It might be thought, however, that equal noise reduction might be obtained by taking the output of (a) and smoothing it with a filter of time constant equal to the observed time constant of (b). The result of this is shown in (c) and it can be seen by comparison with (b) to be a less efficient noise averaging system. It should be pointed out that with certain types of noise spectra no significant difference between systems (b) and (c) can be detected, though it is obvious that from considerations of general purpose use (b) is to be preferred.

Operation in the scan mode

On first acquaintance a boxcar can seem to be difficult to use because of its apparent complexity especially when used in the scan mode whereas, in fact, its operation is remarkably simple. An artificial experiment set up in the laboratory demonstrates this.

A waveform (see Fig. 9) buried in noise has a known repetition period of $10\mu s$ and is suspected of having components as short as $200ns$. The boxcar of Fig. 10 is set up as follows:

t_i , initial delay, 0s: allows the sampling to start at the beginning of the waveform.

t_b , timebase, $10\mu s$: gives a full scan across one cycle of the waveform.

t_s , sample width, $100ns$: in order to resolve $200ns$ the next lowest time available is used.

t_c , output time constant, 0.3s: this is set

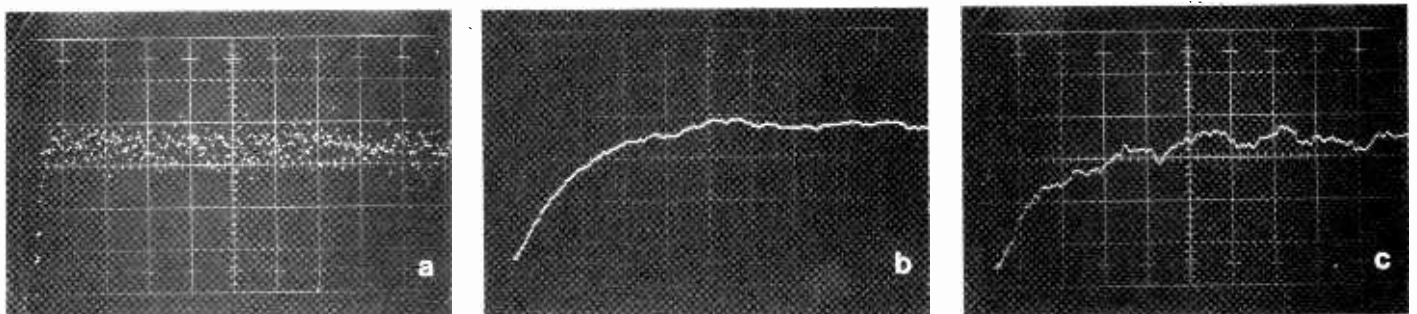


Fig. 8. (a) Voltage on C_g for $R_g C_g \ll t_s$, $t_{obs} = 10\mu s$ (b) voltage on C_g for $R_g C_g > t_s$, $t_{obs} = 10ms$ (c) voltage on C_g for $R_g C_g \ll t_s$ but smoothed by 10ms filter.

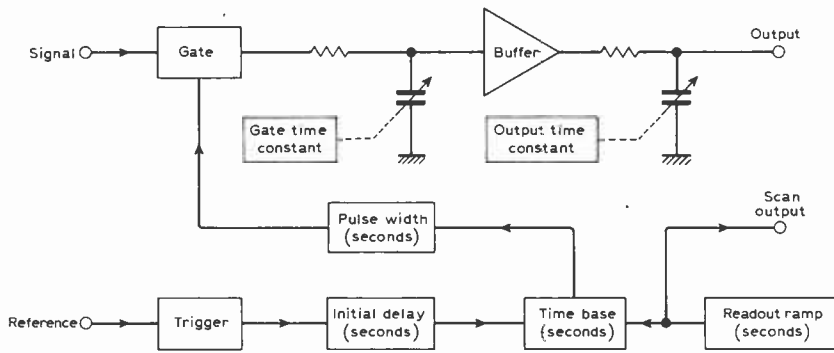
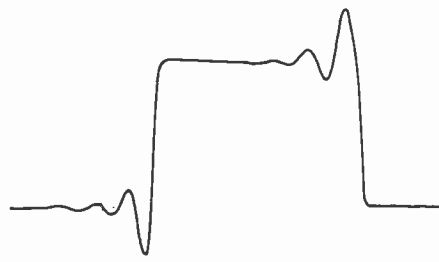
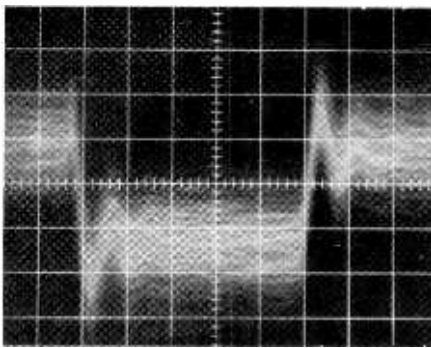


Fig. 9. (top left) a waveform buried in noise. Fig. 10 (lower drawing) Boxcar set-up for scan operation. Fig. 11. (top right) Resulting plot when the waveform of Fig. 9 is applied to the boxcar of Fig. 10.

empirically and in this case gives the required noise fluctuation of 1% of the waveform height.

t_g , gate time constant, 1ms: this is sufficiently long to provide efficient noise averaging within the gate while allowing the overall response time to be set by t_c .

t_s , scan readout time, 100s: the scan readout time needs to be sufficiently long to allow for the response time of the averaging system. It must be greater than $t_c \times t_b / t_g$.

The resulting plot is shown in Fig. 11.

Single-point mode

The boxcar, in its single-point mode in which it measures the mean amplitude of repetitive pulses buried in noise, bears a close resemblance to the phase-sensitive detector which is used to recover the mean amplitude of a simple periodic signal from noise. Both techniques employ a gate (or gates) followed by a low-pass filter and both are used in experiments where the signal is stimulated by a repetitive input which is also used to act as the reference. See Fig. 12. However, whereas the phase-sensitive detector is always gated at half-cycle intervals, the boxcar is gated so that it samples only the signal pulse plus any noise occurring during the sampling time. Since the pulse width may be less than 1% of the repetition period the sample-and-hold configuration shown in Fig. 6 is necessary to carry the information over from one cycle to the next. It can be seen that for pulse measurement, the boxcar is operated with fixed delay so that the sampling window 'encloses' the signal pulse and also with a gate time constant somewhat longer than the gate width so that the output is proportional to the mean amplitude of the pulse.

In Fig. 13 are shown the essential parts of a boxcar used in the single-point mode. The trigger input senses the reference voltage and fires a delay circuit which is manually adjustable so that the gate can be opened just before the signal pulse. This is followed by a pulse width control circuit which is set to close the gate just after the signal pulse.

Conclusion

The existence of high performance boxcar detectors now enables many new measure-

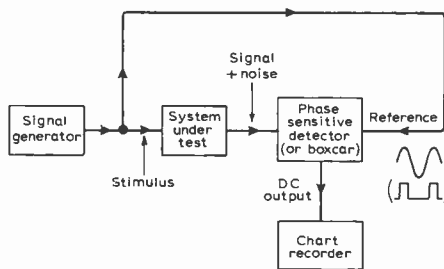


Fig. 12. The recovery of a repetitive signal from noise.

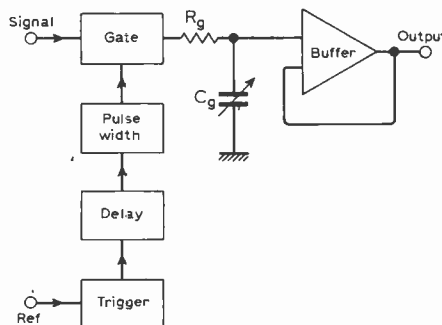


Fig. 13. Boxcar circuit in the single-point mode.

ments to be made on systems, some yielding entirely new information, some eliminating lengthy experimental procedures. The ability to recover short pulses from noise has led to investigations of non-linear effects in systems where there is a limit on mean power dissipation but not on peak power. In addition, the ability to recover noise-obscured waveforms has resulted in the increasing use of impulse testing to characterize the response of systems, where previously the response was measured by swept frequency techniques. Although this article has not mentioned the subject of design criteria it should be added that the use of sampling circuits designed for other purposes will almost certainly give inferior results. In particular oscilloscope samplers are quite unsuitable partly because they have a fixed gate width which does not allow optimum signal recovery conditions to be attained and also because their design, quite reasonably, is aimed at speed of sampling rather than zero stability or linearity. As with any signal recovery system the dynamic range of the signal handling circuits is a most important parameter and worthwhile results will be obtained only if the gate and associated circuits are designed with signal recovery in mind.

Do you know . . .

- The formula for the notch frequency of a bridged-T filter?
- What frequency bands are allocated for industrial, scientific and medical equipment?
- Which transistor manufacturer uses the prefix "SE" for "in house" numbering? The address of the Society of Electronic and Radio Technicians?
- What type of transmission is officially designated P3E?
- What is the name of the sub-multiple 10^{-18} ?
- The SI unit for luminance?
- The significance of the letters in the Pro Electron system of classifying transistors?
- On which television system stations in East Germany operate?
- You cannot be expected to know all the answers (there are no prizes even if you do;) but it gives one a sense of satisfaction to know where the answers can be found. The owner of a *Wireless World* Diary has the answers to all these questions and very many more. The 60-page information section includes formulae, abacs, frequency allocations, circuit building bricks and a host of other technical data.
- The 1971 Diary, which has a week at an opening, is now available price 10s 9d in leather or 7s 6d in Rexine. (postage 6d).

Circuit Ideas

Oscillator circuit for measuring β and f_T

The frequency capability of a transistor may be assessed by plugging it into an oscillator circuit; and by suitable choice of component values β or f_T may be directly measured.

Fig. 1 shows the variation of β with frequency, and Fig. 2 a suitable circuit. The oscillation frequency is chosen to be on

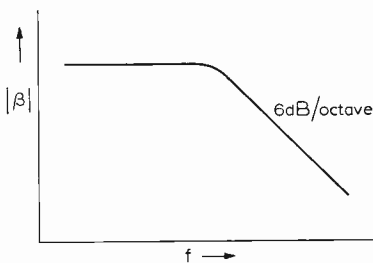


Fig. 1. Variation of β with frequency.

the 6 dB/octave portion of Fig. 1. Here β is not merely reduced from the l.f. value, but has a 90° phase shift as well, so that the base signal feed must be via a small capacitor C_b to enable the transistor to oscillate at the resonant frequency of the tuned circuit. A base signal current i_b produces a current βi_b in R_c ; and the voltage drop $\beta i_b R_c$, with a phase reversal, is fed back, and must equal $i_b/j\omega C_b$ to produce oscillation. Hence if the transistor is just oscillating, $\beta = 1/\omega C_b R_c$.

But $f_T = \beta f = 1/2 \pi C_b R_c$, so that f_T can

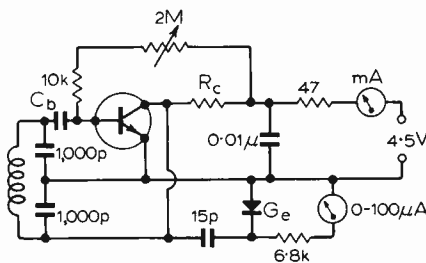


Fig. 2. Suitable test oscillator.

be measured directly, and the oscillation frequency does not matter, provided it puts the transistor on the sloping portion of Fig. 1.

If C_b is in pF, R_c in Ω , and f_T in MHz,

then $f_T \approx 160\,000/C_b R_c$.

In an actual circuit, R_c is 100Ω and C_b is varied in twelve switched steps as follows:
 C_b (pF): 200 120 82 56 39 33 22 15 10 8.2 5.6 3.3
 f_T (MHz): 10 15 20 30 40 50 70 100 150 200 300 500

The two largest values of C_b have been adjusted to allow for their effect on the tuned circuit tapping ratio. The oscillation frequency is 10MHz. The switch is marked with the twelve values of f_T above, and the value to be taken is the highest for which oscillation occurs, as shown by the diode detector. The transistor current is adjusted by means of the $2M\Omega$ base bias resistor, about 10 mA usually giving the best f_T .

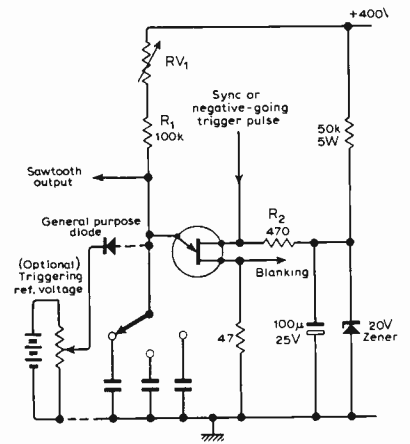
This circuit is intended to give a rough idea of f_T , but by using accurately known components precise measurements would be possible with the method. It would also be possible to measure the small-signal i.f. current gain on the level portion of Fig. 1. In that case C_b would be replaced by a resistor R_b , with a blocking capacitor in series, and β would be R_b/R_c .

Four OC170s tried in this circuit oscillated at the 70 MHz setting, and some showed 100 MHz. A number of unmarked silicon planar transistors showed f_T values from 15 to 300 MHz, and such information would have been difficult to obtain without this very simple and compact test unit.

J. P. HOLLAND,
 London S.W.15

Linear timebase with a unijunction transistor

The unijunction transistor oscillator, when operated from a high-voltage supply, can give a very linear sawtooth sweep; the circuit is both compact and inexpensive. A blanking pulse is available, and the sweep is easily synchronized. It can also be used in the triggered mode. The circuit given will operate reliably with almost any device. For operation with a particular type the values can be adjusted to consume less power; also, stability can be improved by an adequate choice of

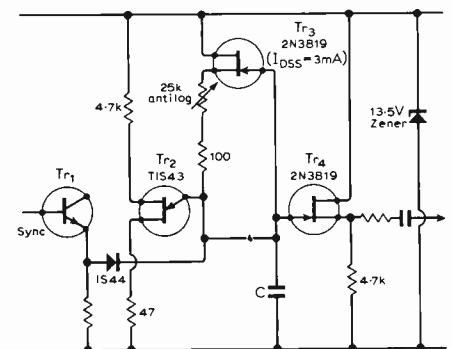


R_2 ; this was not considered necessary for a general purpose oscilloscope timebase as stability was satisfactory. Note that synchronization must be provided from a low-impedance source; and the effect of a load resistance R from emitter to earth is equivalent to a decrease of h.t. by a factor $R/(R+R')$, where $R' = R_1 +$ resistance set on RV_1 , and also causes a frequency change which is not very large if $R \gg R'$. The time constant changes from $R'C$ to $RR'C/(R_1+R)$. The minimum frequency, less than 1Hz, is limited by leakage; the maximum frequency, by device characteristics and stray capacitance. The peak output voltage is about 15V, and flyback is very short. The linearity is excellent, as is evident when it is considered that a capacitor is charged to 15V from a 400V supply.

M. SALEM,
 Peru,
 Argentina.

Linear sawtooth generator

An exceptionally linear sawtooth can be generated by the circuit shown. It uses an f.e.t. as a current source for the timing circuit of a unijunction relaxation oscillator and with unselected components



$C = 1\mu/0.1\mu/0.01\mu$
 to cover range 25Hz-30kHz

will perform over a 10:1 frequency range with a chosen value for C . To avoid loading the oscillator an f.e.t. source follower is used.

S. F. WEBER,
 Rugby.

News of the Month

S.R.C. space activities

The Science Research Council has granted £690,000 over three years to University College, London, to continue the programme of geophysical and astrophysical research in space being carried out at the Mullard Space Science Laboratory (M.S.S.L.) under the direction of Professor R. L. F. Boyd, F.R.S.

Over the past three years, M.S.S.L. has been supported by an S.R.C. grant of £519,000 which allowed the laboratory to continue its very successful programme of ionosphere and magnetosphere research, and of ultra-violet and X-ray astronomy using experiments flown in rockets and satellites. During this period, ten experiments were flown in seven European Space Research Organization (E.S.R.O.) and N.A.S.A. satellites and 53 rocket experiments launched in 37 U.K. and E.S.R.O. rockets.

The new grant of £690,000 is for continuation and expansion of this work. Studies of solar and stellar X-ray emission and solar u.v. emission will continue. Particular emphasis will be placed on the study of cosmic X-ray sources using a reflecting X-ray telescope array in a N.A.S.A. satellite OAO-C, two new instruments in a cosmic X-ray satellite now being planned as a U.K./U.S. co-operative project (to be launched in 1973) and several instruments in U.K. and E.S.R.O. rocket payloads. Studies of the ionosphere and magnetosphere will similarly continue. Satellite experiments under preparation at the present time include a large ion probe for the E.S.R.O.-4 satellite, due to be launched in 1972, and low-energy particle detectors which it is hoped will be carried on a geostationary satellite. Experiments in U.K. and E.S.R.O. rockets will study the coupling between the magnetosphere and the polar ionosphere which involves the same low-energy particles.

The Science Research Council's Radio and Space Research Station at Datchet, Bucks, has taken delivery of an ICL 1904A computer that will be used to assist scientists at R.S.R.S. in processing data and solving problems associated with their tasks. The station's main involvements are with tropospheric and ionospheric research including their applications to communications. Other activities include space research both by designing experiments

for rockets and satellites and by providing services for experimenters in British Universities and elsewhere.

The R.S.R.S. has operated an ICL 1905 computer during the last two years. The 1904A replacement with additional equipment brings the total value of the installation to nearly £400,000.

R.S.R.S. works closely with N.A.S.A. and E.S.R.O. Data from E.S.R.O. is already being processed by the station which has two experiments in each of the E.S.R.O. 1A and 1B satellites launched in 1968 and 1969.

The more fundamental work of R.S.R.S. is aimed at understanding the nature and behaviour of the troposphere (the lower stratum of the atmosphere, lying below the stratosphere, where radio waves are affected by meteorological conditions) and of the ionosphere (the upper stratum of the atmosphere where radio waves are affected by charged particles). In addition

to international collaboration, the station maintains one of the world data centres for ionospheric and space science information.

An interesting experiment in the troposphere undertaken by R.S.R.S. jointly with the Post Office is concerned with the influence of precipitation on microwaves. Microwaves tend to fade in rainfall and to measure this, 47 rain gauges have been set up over an area of 144 square kilometres in Suffolk. Radio beams are transmitted across the area and should these fade, the rain gauges are interrogated to determine the amount and the distribution of rainfall. Data from this experiment will eventually be analysed on the 1904A at Datchet. Information gained from this experiment will aid advanced planning and siting of terminals in communications networks which could provide the basis for high-speed data transfer for computer networks of the future.

New radar for R.A.F. missile range

The obsolete radar in use on South Uist, one of the St. Kilda group of islands 40 miles from the Outer Hebrides, is to be replaced by a more powerful AR1 radar. The present radar system, a type 54, is to be transferred to the Royal Air Force Museum at Hendon. The radar will be used to search for ships and aircraft in the missile range when a firing is to take place.

The AR1 was originally designed for airfield surveillance where it would

The south east Berkshire scouts were very well equipped for the scouting movements 13th annual Jamboree of the Air which was held in October. The Racal amateur radio club GH3RAC and the Racal Electronics Group loaned an impressive array of equipment. The scouts were able to set up four transmitting/receiving stations, five monitoring stations, a news and weather bureau with all the necessary sound and video equipment, a teleprinter and a Mufax picture receiver.



normally be mounted fairly close to the ground. In the missile range installation the radar will be mounted at an altitude of about 1,200 ft and technicians are interested to find out what this will mean in terms of increased range.

C.A.D. for all

All you do is hire a computer terminal consisting of an electric typewriter and a G.P.O. modem, pay Racal £50, and providing you have access to a telephone line, you have a powerful computer-aided-design service at your disposal. Your £50 will buy you the necessary tuition, instruction manuals and a user number. Your first two-hundred pounds worth of use will cost you only one-hundred and fifty pounds so you quickly recoup your down payment.

The service is called Redacal and it might be used as follows. An engineer wants to design a three-transistor amplifier. He sketches the design and assigns component values and reference numbers. He gives each connection point, or node, in his circuit a number. A punched tape is then prepared by the engineer which describes the circuit to the computer or the same information can be fed in by using the keyboard. A typical keyboard statement might be R1, 6.8k 12, which would mean that resistor R₁, which has a value of 6.8kΩ, is connected between nodes one and two. In this way all the components in the circuit are described and the computer is able to simulate the complete circuit.

The designer may then wish to specify

an ambient temperature, or range of temperatures, and request the computer to print out gain at several frequencies or he may ask what is the emitter current in Tr₂ and so on. In this way the design can be toleranced and optimized. In our example a designer may be 'conversing' with the computer, using normal electronic terminology, for 45 minutes and would be presented with a bill for around £50. For his money he has a properly toleranced and optimized design that may have taken him several days using normal methods—and nobody has touched a soldering iron yet.

The Redacal service offers the following programmes. Linear circuit analysis, non-linear d.c. analysis, non-linear transient analysis and databank. The last of these is a computer file containing the parameters and mathematical models of a huge selection of active and passive components that is being continuously updated. Any of these can be placed in the designer's circuit, instead of the components originally specified, if required.

'Snap-shot' and storage TV camera tube

A vidicon type of television pick-up tube that can take, and store for later reproduction, single 'snap-shots' or 'stills' from continuously moving scenes has been developed by RCA Laboratories in the U.S.A. The still picture stored in the tube can be transmitted to a monitor immediately after it is taken, or kept within the device for

several days. Up to 200 pictures per minute can be stored and transmitted, so the rate of operation can be largely determined by the signalling speed requirement of the associated communication system.

Described as a silicon storage vidicon, the tube has a target, which is a silicon wafer, about an inch in diameter containing 600,000 photosensitive diodes. Normally these diodes are scanned sequentially by a low-velocity electron beam and the output from them is amplified to provide a video signal. The diodes are isolated from each other by a silicon dioxide insulating layer which has 600,000 holes to permit the electron beam to pass through to the diodes.

The process of storing a picture is begun by scanning the entire target with a high-velocity electron beam, which erases the previous picture. Then, with the next picture optically focused on the tube, the target is scanned by a somewhat lower velocity electron beam for several television frames, building up electric charge in the silicon dioxide. As a result the picture is stored as a pattern of charge.

To 'read out' the stored picture, the target is scanned by a beam of still lower velocity electrons and at the same time a small lamp outside the tube illuminates the target. The current flowing through each diode during the 'read' cycle is determined by the amount of charge stored at each point on the silicon dioxide during the 'write' cycle.

Giant astronomical satellite

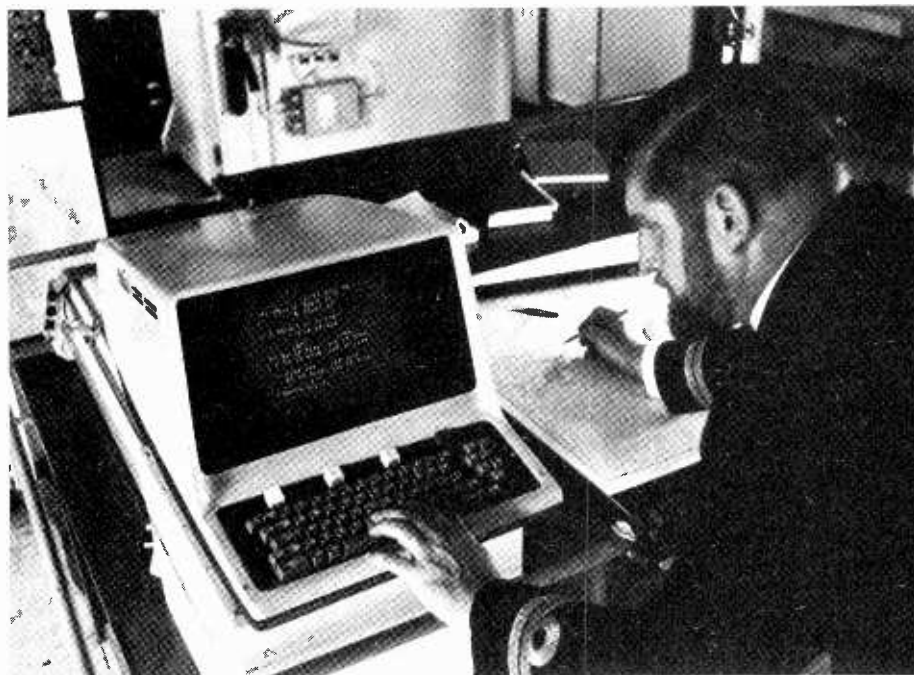
By the time you read this the most complex satellite to be built by the U.S.A. should be in a circular orbit at an altitude of 466 miles (period 100 mins). The satellite in question, OAO-3 (orbiting astronomical observatory), was scheduled to be launched 'no earlier than November the 17th'. If the launch is successful OAO-3 will study by means of the on-board 36-inch telescope, distant stars and galaxies (down to 12th or possibly even 13th magnitude) with detail previously unobtainable due to the masking effects of the atmosphere. The satellite weighs 4,680 pounds and stands ten feet tall; it has 21-foot solar panels, which, when unfolded, gives the structure the appearance of a giant bat.

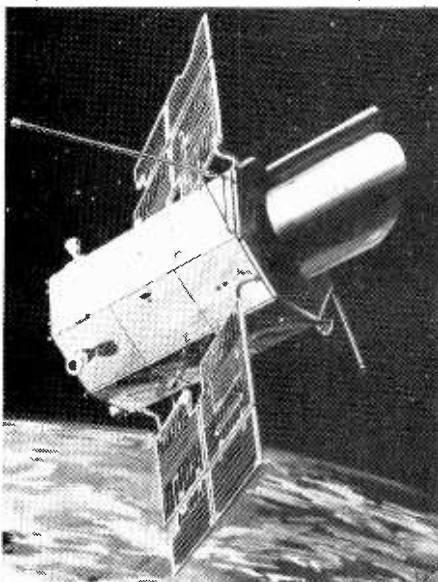
OAO-3 has two sister satellites. OAO-1 was launched in 1966 but failed shortly after going into orbit due to a malfunction in the power supply system and probably high-voltage arcing in the star tracking system. OAO-2, launched in 1968, has far exceeded its design objectives and still continues to operate successfully.

One of the most important things to be studied by OAO-3 is interstellar dust which many authorities agree holds the key to the origin of the universe.

The accuracy with which the telescope on board OAO-3 can be pointed is the equivalent of a rifle sight being held on a two-inch diameter target at a range of six

United Nations' survey work on fish off the north-west coast of Africa is to be carried out with the aid of an on-line data processing system. The Naval Division of Marconi Space and Defence Systems has received a £74,000 contract to equip a new fisheries research trawler with Hydroplot, a computer system which will collate, analyse and present, in real-time, a mass of data derived from navigation aids and oceanographic sensors.





A sketch of OAO-3 in orbit.

miles for several minutes.

The next satellite in this series will also carry a large telescope, and an X-ray experiment from University College, London and the University of Leicester.

Audio Engineering Society Convention

The 39th Convention of the A.E.S. was held at the Hotel New Yorker, New York, from 12th to the 15th of October where some 70 papers were read during 13 sessions. The sessions covered the major disciplines in audio engineering and in particular four-channel technology, multi-channel recording techniques and electronic music. Undoubtedly the greatest interest was shown in four-channel disc and tape recording and reproduction, and at least two disc recording systems showed considerable promise, those of the Victor Company of Japan, and the matrixing system developed by Peter Scheiber of the Audiodata Co. Although a considerable amount of four-channel hardware was to be seen in the exhibition and many demonstrations were heard, there was a shortage of either disc or tape software, and not until some internationally agreed standards are formulated can one expect an upsurge in the domestic field.

The general acceptance of the tape cassette has created a new field for high-speed duplicating equipment, and demonstrations of the latest Dolby "B" system showed the closing of the gap between disc and tape reproduction, and it may not be very long before twin-track, four-channel cassettes become available.

The highlight of the annual Awards Banquet was the presentation of the Emile Berliner Award to Arthur C. Haddy, the technical director of the Decca Record Co. Ltd, "for pioneering development of wide-range recording and playback heads and for

his significant part in the international adoption of the 45°-45° stereo disc recording system". In his speech as guest of honour, Arthur Haddy made particular reference to his team of engineers who had so ably supported him in more than forty years in the industry. The John H. Potts award was presented to R. T. Bozak for his work in loudspeaker development, and among the recipients of the Fellowship of the A.E.S. was Willi Studer and Percy Wilson, the first technical editor of *The Gramophone*.

Hitachi make PAL television receivers

An agreement has been reached between AEG-Telefunken, of Germany, and Hitachi, of Japan, enabling Hitachi to manufacture PAL system colour television receivers under licence. According to Stanley Fenner, general sales manager of the newly formed company Hitachi Sales (UK) Ltd, sets will be available in the spring of 1971. This arrangement has also been agreed with E.M.I., (co-licenceses for PAL system receivers in the U.K.). It is subject to the approval of the Japanese government, who will no doubt want such an agreement extended to other Japanese manufacturers.

U.S. space programme costs

These are the figures as submitted to us by N.A.S.A. If you add them up you will find that money is indeed no object!

1959	330.9 × 10 ⁶
1960	523.6 × 10 ⁶
1961	966.7 × 10 ⁶
1962	1,825.3 × 10 ⁹
1963	3,674.1 × 10 ⁹
1964	5,100.0 × 10 ⁹
1965	5,250.0 × 10 ⁹
1966	5,175.0 × 10 ⁹
1967	4,968.0 × 10 ⁹
1968	4,588.9 × 10 ⁹
1969	3,953.0 × 10 ⁹
1970	3,696.6 × 10 ⁹
Total	\$40,051,600,000

Public Address Exhibition

Main theme of the next annual exhibition of the Association of Public Address Engineers will be Sound Techniques for the '70s. The exhibition, the only one of its kind in Europe devoted exclusively to public address equipment, will be held 16-19 March 1971 in Camden Town Hall—opposite St. Pancras Station—following last year's successful change from the old Harrow venue. Lectures will be given

by Ralph West, on loudspeakers, R. Jones, on amplifiers, Hayden Warren on microphones and by a representative from B & K on acoustic measurement. Further information is available from the Exhibition Office, A.P.A.E., 3 & 4 Northolt Road, South Harrow, Middlesex HA2 8EY (telephone 01-864 3405).

An exercise in circuit maximization

Under the above heading, in our August issue, we had a playful dig at a lamp flasher circuit released by Motorola in which we suggested that the designer perhaps had his eyes on the sales figures. This was because the circuit used a very large number of components. We have since heard from Motorola that the circuit was designed to meet a very stringent specification and is now in use in defence equipment in the United States. For security reasons Motorola are unable to release all the whys and wherefores of the circuit but one of the requirements was that it should consume no standby power. While apologizing to Motorola for our little bit of fun, and accepting their explanation completely, we would like to point out that our comment was prompted by the suggestion in the original circuit description that the device was intended for use in cars for direction indicators, amongst other things.

Conferences and Exhibitions

Further details are obtainable from the addresses in parentheses

LONDON	
Dec. 6-9	Olympia
Opportunity 70—Careers Exhibition (R. W. Boardman (Exhibitions) Ltd, 8 Leicester St., London W.C.2)	
OVERSEAS	
Dec. 2 & 3	New York
Display Devices (I.E.E.E., 345 East 47th St., New York, N.Y. 10017)	
Dec. 2-4	Washington
Vehicular Technology Conference (I.E.E.E., 345 East 47th St., New York, N.Y. 10017)	
Dec. 6-9	Chicago
National Electronics Conference (Nat'l Elec. Conf., Oakbrook Exec. Plaza 2, 1121 W. 22nd St., Oak Brook, Ill. 60521)	
Dec. 7-9	Texas
Adaptive Processes: Decision & Control Symposium (D. G. Lainiotis, Elec. Res. Ct., University of Texas, Austin, Texas 78712)	
Dec. 9-11	New York
Applications of Simulation (I.E.E.E., 345 East 47th St., New York, N.Y. 10017)	
Dec. 14-16	Atlanta
Symposium on Circuit Theory (I.E.E.E., 345 East 47th St., New York, N.Y. 10017)	

London's Audio Fair

A new quest for realism?

The organizers of this year's International Audio and Music Fair (let's give it the honest title London's Audio Fair, LAF for short), seem to have made it a complete success. A large hall such as Olympia, although possessing some of the grim and grimy desolation typical of a railway station, is certainly the best place to hold an Audio Fair open to the general public.

One of the most striking features of this year's Fair was the evidence arising from the better demonstrations, and the listeners' reactions, that naturalness in sound reproduction really is considered to matter. Also the attendance at the programme of lectures, and the response to what was said, showed that people are keen both to learn and to be guided.

The common theme of all the *Wireless World* sponsored lecture demonstrations was that the proper aim of audio engineering is the recreation of the original sound—all the speakers confessed that correct reproduction of music was central to their design efforts. Both Arthur Bailey and Ted Jordan demonstrated that speakers with flat response curves (measured in an anechoic room) can sound very wrong indeed. Also, the subject of class A versus class B arose (inevitably) and present opinion seems to be that a class A output stage is easily set up to work consistently well but has a peak-power availability severely limited by the amount of allowable continuous dissipation, whereas a class B system with no crossover distortion (!) needs to have very tightly controlled quiescent current (not varying with rail fluctuations) and symmetry of the upper and lower halves independent of temperature variations.

The overall message of the lectures seemed to be "listen carefully and let your ears be the final judge".

The shortcomings of last year's Fair were few but fairly serious. The walls of the listening rooms were flimsy and gave rise to highly objectionable colouration as they resonated with the programme material. Ventilation was also a problem.

This year the listening rooms were substantial structures, well braced, and each ventilated by two fans. Good quality demonstrations were possible, and provided the room was not crowded the fans could be turned off while the equipment was being put to the test.

What were the demonstrations like? Before reporting on this in any detail, a few general points can be usefully made. No session need last longer than fifteen minutes, and the less technical talk the better.

Also, demonstrators of amplifiers and loudspeakers would do well to prepare a top-quality demonstration tape and play it through a good tape deck. In this way a programme can be prepared to run efficiently with an unvarying standard of excellence. With the limitations inherent in even the best magnetic cartridges tape is the better medium anyhow.

Loudspeakers

K.E.F. switched between speakers during moments of musical silence—between phrases and movements—and there was no serious distraction of the listener. It was possible to make a valid choice of preferred speaker performance from this session. The new speaker on display was the Cadenza which employs a bass diaphragm driven from the rear of the 8-in B200 bass and mid-range unit.

Bell & Howell, representing Acoustic Research, showed a very honest face when demonstrating the AR-4x, the new AR6 and the top class AR3 loudspeakers. In the catalogue they declared "when listening, please bear in mind that it is believed that the term 'high fidelity' means fidelity to the sound of live music. If you agree, then your basic comparison should not be whether speaker A sounds better than speaker B, but rather how each compares to the sound of live music". We agree entirely, of course. The new AR6 employs a passive radiator like the K.E.F. Cadenza. It is a truism that amongst acoustic suspension systems from really experienced manufacturers you get what you pay for.

New loudspeakers were demonstrated by Leak and Wharfedale. The leaflets describing the new speakers—the Triton and Aston from Wharfedale and the sandwich 200, 300 and 600 from Leak—are quite a revelation. Wharfedale say "The pure clean sound of the Aston will excite the most critical music lover", and Leak speak of "an amazing frequency response of 60 Hz to 18,000 Hz" and make a virtue of using a 12in bass driver in a 3 cu ft enclosure.

Two slim-line speakers from Sansui, the SL-5 and SL-7 each employ a "drone cone"—a passive radiator of the type mentioned earlier. We are informed by the handout that "A massive ferrite magnet assembly, specially processed urethane edge and linear-efficiency damper combine to overcome the anti-resonance problem. . .". Also, the treble horn-type tweeter of the SL-5 "reproduces high-frequency tone with mirror-like clarity". The crossover frequency between the two active units is 6500Hz, and bass response falls off below 70Hz.

A kit for a new bookshelf speaker is available from Heathkit. The speaker, called the Trent, uses an 8in bass driver and a 4in cone-type treble unit, the enclosure measuring 10×19×8½in. Crossover is by a single capacitor. The price for the kit is £14.

The JBL range of speakers certainly demonstrates the difference between American and British requirements of size and appearance. Large items of furniture were yet examples of complex engineering excellence. The sound level from the speakers in the demonstration room was too high (we were assured that this was to prove their superiority) and bass and treble lift seemed a permanent amplifier setting. Technical specifications for the "domestic" speaker range were unobtainable but we did get details of the model 4320 studio monitor (price £215). The 15in bass driver (2215) has a 4in voice coil. Total flux in the voice-coil gap is 450,000 maxwells! The high-frequency driver (2420) is coupled to an aluminium exponential horn with a slant-plate acoustic lens. Sound from the horn is then spread through 120° laterally but only 40° vertically. This provides excellent dispersion for two-channel stereo. The cabinet has a pair of ducted ports giving increased loading between 30 and 50Hz. The system is rated at 120W maximum input power.

Now we turn to a different phenomenon—the availability of speakers designed with omnidirectional diffusion over almost the full frequency spectrum. Such designs approximate to spherical sources but dispersion is modified by the nearness of floor and walls.

Ideally, the mid-range and treble spectrum should radiate through a wide angle laterally but a narrow one vertically, and

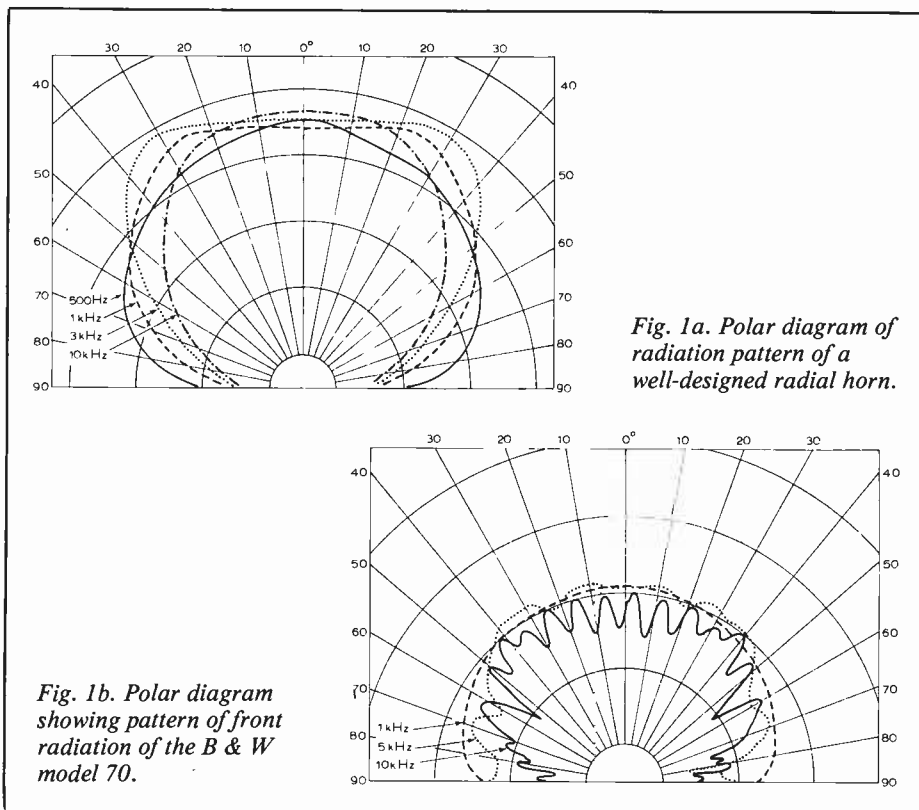


Fig. 1a. Polar diagram of radiation pattern of a well-designed radial horn.

Fig. 1b. Polar diagram showing pattern of front radiation of the B & W model 70.

there should be no significant treble emphasis on the driver axis. Now it is unusual for medium-price speakers to provide anything like the correct dispersion. A cone-type tweeter will almost always deliver frequencies above 6kHz in such a way that on axis they are too loud but a few degrees off axis they are too soft. (Those accustomed to looking at polar diagrams would see this situation as a narrow lobe projecting far forward of the rest of the constant-power frequency lines.) The result of a poor polar response with heavy axial emphasis is the virtual impossibility of obtaining a satisfactory stereo image. If you move a few feet from the 'ideal' position the image collapses or severely distorts. Excellent polar response is given by a radial horn driven from a single diaphragm (Fig. 1a). A multicellular horn can give high-frequency 'fingering' because the outputs of each section are not finally integrated, and this also occurs in the Bowers and Wilkins model 70 (see Fig. 1b) due to the spacing of the eleven electrostatic modules.

Returning then to the omnidirectional types—the Sonab series, the Harman Kardon AK50 and Citation speaker, the JBL Aquarius 4, the ITT KB 'radial sound' speakers, and the Nivico spheres—it must be said that these cannot provide the geometrical basis for sharp-image stereo. They provide a diffuse source where singers have gigantic mouths and musical instruments are free to move about as they produce different frequencies.

Notes on turntables and cartridges

Garrard displayed two new models—the AP76 'transcription quality' single record playing unit, and the SP25 Mk III with calibrated bias compensation.

Goldring have a new turntable unit model

705 with belt drive. Rumble level is given as -35dB and the auto/manual arm mechanism is hydraulically damped. The price is £19 8s.

Two 'transcription quality' pickup arms and a range of magnetic cartridges (£7-£21) were displayed by Micro Seiki. The MA100 12in dynamic balance arm gives 'the highest degree of tracking accuracy yet devised' (£75), and the MA77/11 allows coarse and fine stylus pressure adjustment up to 6gm (£25). The MC4100/E, requiring a booster amplifier or step-up transformer, is a magnetic cartridge costing £21. Frequency response appears to be creditably flat and crosstalk very satisfactory.

B.S.R. displayed the McDonald range of record turntables comprising the MP60 single player, with plinth and cover (£23 16s), and automatic changers types 610, 510 and 310. The 310 has a low-mass arm with manual cue/pause lever.

Five new 'x' range ADC cartridges are now available from K.E.F.; and a new low-price Audio Development magnetic cartridge, the AD96K, is available from B. H. Morris & Co.

Reduced-noise tape units

Hiss reduction system for cassette tape recorders is used on the Bell & Howell DES 1700. Cassette equipment using noise reduction techniques first appeared on the American market in 1969—by Fisher, Advent, Harman Kardon—and in the U.K., Kellar Electronics (see p.488 October issue), were first to produce equipment. Like the Kellar DCR1 recorder deck, the Bell & Howell deck is designed for both recording and playback, the noise reduction system lifting the level of quiet high-frequency signals above the tape noise level during recording and reducing them—and

the noise—during playback (Fig. 2). As well as catering for cassettes using this technique—such as those being produced by Decca—these machines allow recordings to be made at 1½in/sec with less noise than previously possible. Use of this system, called Dolby 'B' being derived from a simplification of the Dolby 'A' system announced some time ago (p.632 December 1966 issue), is not limited to cassette recorders, of course, and Kellar also have a unit—type KDB1—which can be used with any good tape recorder. The Bell & Howell unit—available February 1971 at about £100—needs separate power amplifiers. The Harman Kardon CAD5 deck was also shown. Specifications of all these decks are similar (see table) as the basic Japanese-made mechanism (Nakamichi) is common.

Decca, producer of cassettes recorded with the Dolby 'A' system, are not happy with this deck, especially over repeatability of performance due to tape wander, and are working on an alternative type. Conventional cassette decks do not have good enough wow figures and with one at least the improvement in noise level is an embarrassment because of noise picked up from the motor. When a machine is finalized it will probably use the new chromium dioxide tapes made by Dupont or Agfa.

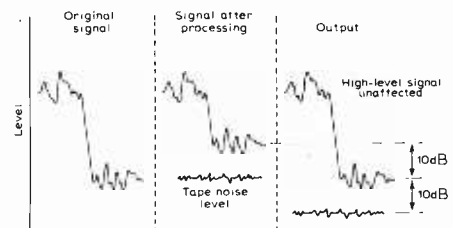


Fig. 2. In Dolby 'B' noise reduction system low-level high-frequency signals are raised above noise level prior to recording and reduced together with the noise on playback.

Signal-to-noise ratio	52dB	(0 VU)
Tape speed fluctuation	1%	
Wow and flutter	0.15%	
Crosstalk	< 35dB	(1kHz)
Bias frequency	105kHz	
Erase	> 60dB	
Harmonic distortion	2.5%*	
3dB roll off	about 12kHz	

*1.5% is quoted for Harman Kardon CAD5

Prices:

Kellar DCR1 (cassette)	£95
KDB1 (open reel)	£45 (no tax)
DTA50 (with amplifier)	£148
Harman Kardon CAD5	£185
Bell & Howell DES 1700	£100
Rank (to be announced 1971)	£115

New Sanyo three-head tape deck incorporates a novel circuit to reduce tape hiss called an 'actual noise silencer'. When playing low-level passages, where noise would be most obvious, high frequency signals are attenuated. Circuit operates by having the collector load of a transistor effectively across the amplifier output.

Variation of the transistor output impedance, by altering its collector current, causes attenuation of the output. This transistor is normally biased on, presenting a relatively low impedance at the collector, and is cut off when l.f. signals reach a certain level increasing the output impedance and hence reducing the shunting effect of the transistor. This is clearly not a compensated system, like the Dolby system, because there is no pre-conditioning of the recorded signal. Snag is that your 'actual noise silencer' seems to be your actual signal silencer! Circuit is switchable but we haven't yet had an opportunity to hear a comparison test. This recorder incorporates several other features including superposing 'sound on sound', monitoring during record, echo effects, beat eliminator, and different switchable bias currents for standard and low-noise tapes. Output level is around 0.7V. Price is just under £100.

New tape heads

Another development in tape recording is the use of new heads made from crystalline ferrite and glass by both National and Akai. The Akai GX head, developed for video tape recorders and used on the new tape recorder type GX-365, claims to reduce head wear by at least an order of magnitude because of its hardness. At the time of going to press details were not available of the National heads, used on model RS735 tape recorder (price £96).

Thick-film power amplifiers

A trend in power amplifier design which emerged this year is the use of hybrid power integrated circuits as opposed to low-level monolithic i.c.s. First British amplifiers to use hybrid i.c.s. are those designed by Stanley Kelly of Kellar Electronics—in the amplifier used in the DTA50 cassette tape recorder and in a 15+15 watt amplifier marketed with the name Nova. The Nova amplifier at about £1 per watt offers very good value for money, made possible by the use of a Bendix hybrid amplifier. (Circuit of these amplifiers is similar to that shown on p.448 of the October issue.)

Rotel amplifiers and tuner

It is gratifying to know that much imported Japanese equipment is specially tested and modified for the U.K. market. Rotel amplifiers, for example, have been put through their paces at Rank and as a result various changes have been made to designs. The RA610 amplifier, for instance, has had its output transistors changed as well as the push-button switches. For 30+30 watts price is £65. Amplifier RA310 gives 15+15 watts at £42 10s and RT320 tuner sells for £54 10s.

Graphic control unit

Audix Ltd, manufacturers of professional sound systems, have produced a stereo graphic control unit for the enthusiast market. Designed by David Rees with

techniques used in studio sound mixers it has an outstandingly low noise level of -80dB for high-level inputs and an equivalent input noise of about -115dB for the magnetic pickup channel. There are a number of circuit techniques not normally used in pre-amplifiers and details will be released by the company shortly. As well as five controls of the 'graphic' kind, giving a control range of ± 12 dB at 80, 250 and 800Hz and 2.5 and 8kHz, there is a noise cancelling circuit which also raises the level before equalization is applied. Power amplifiers are available for use with this control unit with outputs of 80 and 35 watts per channel (into eight ohms). Both have tone correction controls which give a few decibels lift or cut at high and low frequencies. Control unit costs £162 and power amplifiers £97 and £83.

Akai tuner-amplifier

New Akai products include two tuner-amplifiers. Model 8500TA is the most powerful and expensive Akai receiver. Giving 85 watts per channel into four ohms, it has an impressive specification. Stereo crosstalk is 40dB (1kHz), capture ratio 1.5dB and f.m. signal-to-noise ratio is 70dB. Using up-to-date circuit techniques, it has a 'graphic' control unit and separate a.m. and f.m. signal-strength meters. It naturally has the Akai protective circuitry which safeguards the output stage against wrong connection—even to the mains supply—by switching off for intervals of five seconds until the fault is cleared. Price is £228. For more modest needs the 6000 series of tuner-amplifiers now has three variants—the new 6200: 15+15 watts for £95, 6300: 20+20 watts for £128 and 6600: 37.5+37.5 watts for £142.

Nikko tuners

New 'budget-price' a.m./f.m. tuner made by Nikko (Japan) is without any a.f.c. circuitry. Details of the circuit were not available at the time of going to press, so how the necessary stability is achieved isn't clear. Priced at £48 10s this tuner has a sensitivity of $2\mu\text{V}$ on f.m. measured by I.H.F.M. standards, a capture ratio 4.5dB and crosstalk of 32dB (1kHz). At the other end of the price scale the Nikko STA-1101 costs £246. For this you get 2×37 watts into 8 ohms, damping factor of 30 (8 ohms), a speaker compensator giving 10dB boost at 30Hz, 0.6% intermodulation distortion, power response of -1dB at 30kHz, crosstalk on f.m. of 40dB (1kHz), capture ratio of 1.5dB, distortion on stereo reception of 0.8% and a sensitivity of around $1\mu\text{V}$ on f.m. Using the current circuit components of f.e.ts. in the front ends (five), integrated circuits (12) and solid-state i.f. filters, this receiver is claimed to be the first to use an f.e.t. for the a.m. detector. Stereo decoder also uses an i.c. for A-B demodulation.

Test equipment

Audio test gear, intended mainly for manufacturers, is made by Sign Electronics Ltd

and marketed by Rogers Developments. Available at present are a low-distortion oscillator (£63), a.f. millivoltmeter (£70), and distortion factor meter (£140). Due for release early in 1971 are a tone burst generator (£85), monitored attenuator (£47), followed by sweep oscillator (£180), and 100-watt power meter later in 1971. Also on the cards are a stereo encoder, a wow and flutter meter, and an auto-nulling distortion factor meter.

Short items

- . . . Highgate Acoustics market a range of equipment in agreement with Rank—who now have a stake in Arena—under the brand name Alpha-Arena.
- . . . New Sony a.m./f.m. tuner—type ST 5600—has stereo crosstalk of 38dB at 400Hz, capture ratio 1.6dB, harmonic distortion of 0.7% stereo for 100% modulation at 400Hz. Price £64.
- . . . Uher have produced their first stereo amplifier, model CV140, priced at £175.
- . . . Goldring-Lenco ST3000 amplifier uses germanium AD 166 transistors in output stage and is rated at 15+15 watts. (Difficult to see, this, as consumption is quoted as 20 watts!)

Addresses of Agents and Manufacturers

- Acoustic Research International—see Bell & Howell
- Akai—see Rank Audio Visual
- Arena—see Highgate Acoustics
- Audix Ltd, Stanstead, Essex.
- BSR Ltd, Monarch Works, Cradley Heath, Warley, Wors.
- Bell & Howell Ltd, Alperton House, Bridgwater Road, Wembley, Middx.
- Bosch Ltd, Rhodes Way, Radlett Road, Watford, Herts.
- Decca Radio & Television, Ingate Place, Queenstown Road, London SW.8.
- Denham & Morley Ltd, 453 Caledonian Road, London N.7.
- Feldon Recording Ltd, 126 Great Portland Street, London W1N 5PH.
- Garrard Engineering Ltd, Newcastle Street, Swindon, Wilts.
- Goldring Manufacturing Co. Ltd, 10 Bayford Street, London E8 3SE.
- Grundig (Gt. Britain) Ltd, Newlands Park, London S.E.26.
- Hacker Radio Ltd, Norrey's Drive, Cox Green, Maidenhead, Berks.
- Harman Kardon—see Highgate Acoustics
- Heathkit (Gloucester) Ltd, Gloucester, GL2 3EE.
- Highgate Acoustics Ltd, 184 Great Portland Street, London W.1.
- Howland-West Ltd, 2 Park End, London N.W.3.
- ITT Consumer Products (UK) Ltd, Footscray, Sidcup, Kent.
- JBL—see Feldon Recording
- KEF Electronics Ltd, Tovil, Maidstone, Kent.
- Kellar Electronics Ltd, Maryland Works, London E.15.
- Leak, H. J. & Co. Ltd,—see Rank Wharfedale
- Micro Seiki Co.—see B. H. Morris & Co.
- Morris, B. H. & Co., 84 Nelson Street, London E.1.
- National—see Unamec
- Nikko—see Howland-West
- Nivico—see Denham & Morley
- Rank Audio Visual, P.O. Box 70, Great West Road, Brentford, Middx.
- Rank Wharfedale Ltd, Idle, Bradford, Yorks.
- Rogers Development (Electronics) Ltd, 4 Barmston Road, London S.E.6.
- Rotel—see Rank Audio Visual
- Sansui—see Vernitron (UK) Ltd, Thornhill, Southampton SO9 1QX.
- Sign Electronics—see Rogers
- Sonab Ltd, 136 Mansfield Road, London W.3.
- Uher—see Bosch
- Unamec Ltd, United Africa House, Blackfriars Road, London S.E.1.

High-quality Tape Recorder

2. Construction

by J. R. Stuart, B.Sc.

Signal monitoring takes two forms in this recorder—A-B signal switching, and a peak-programme meter to read signal and bias levels.

Experience has shown that the most effective meter has a fast attack and slow decay time. This stabilizes the pointer and gives a pessimistic reading, reducing the risk of overload while maintaining wide dynamic range.

Probably the most difficult transient a music signal could provide would be of 25 μ s duration, being one half cycle at 20 kHz! Although this is unlikely the meter described was designed to attempt to cope with this. A circuit diagram is shown in Fig. 11.

For simplicity only positive-going peaks are read. It is a simple matter to extend the circuit to read positive and negative peaks, but no instances have been found to suggest that this should be necessary. The author has used a meter of this type for several years without problems.

As the input signal to the recorder is balanced, either from a programme source or mixer, it is not necessary to have two expensive meter movements. Instead a ganged record gain control is used and the meter circuit of Fig. 11 indicates on a logarithmic scale the peak value of whichever channel is the greater at any instant.

Two i.c. amplifiers, A1 & 2, are used to raise the OdB input signal to 1V r.m.s. These low output impedance amplifiers charge

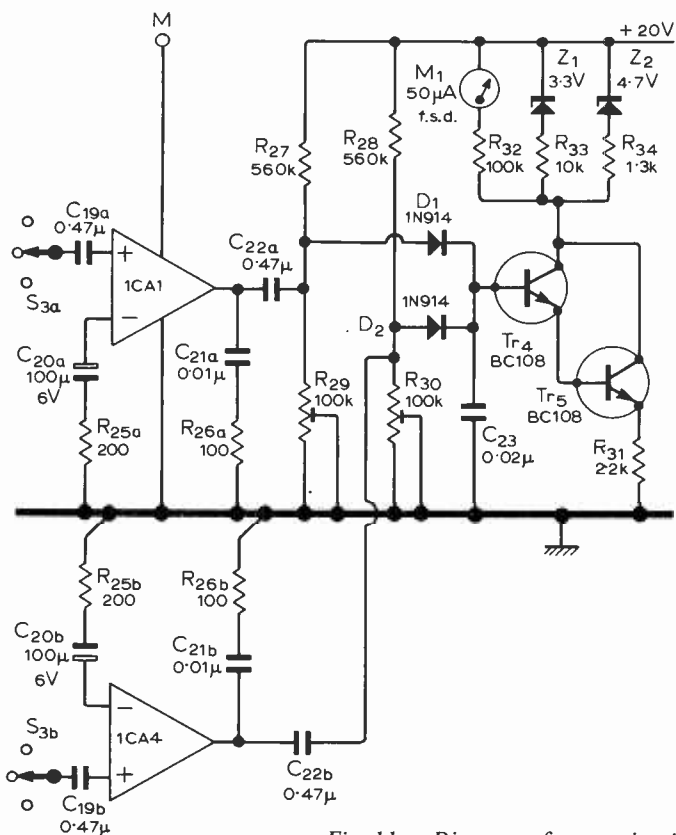


Fig. 11. Diagram of meter circuit.

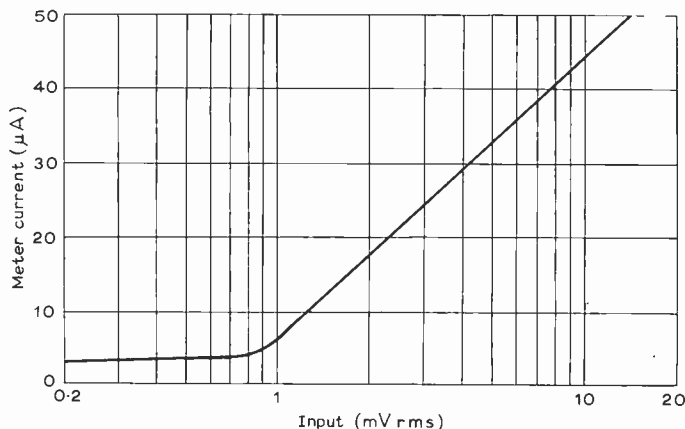


Fig. 12. Response of meter circuit.

capacitor C_{23} to the peak positive value via D_1 or D_2 and as the forward resistance of the diodes is current dependent, the attack time will depend upon the amplitude difference between successive peaks, with a minimum value of around 20 μ s. The decay time is determined by the rate at which C_{23} is discharged by the high input impedance of the amplifier formed by Tr_4 and Tr_5 . This impedance is nearly $\beta_4 \beta_5 R_{31}$ and the measured decay time is around 2s. Two zener diodes are included in the collector network to give a three-slope approximation to a logarithmic response; this approximation is good as can be seen in Fig. 12. Table 3 gives the meter calibration. The linearity is good and the meter response to continuous sine-wave is within ± 0.5 dB from 10 Hz to 10 MHz.

To set-up the meter, switch on with the wipers of R_{29} and R_{30} set to ground, slowly rotate R_{29} until 2 μ A flow in the meter and then rotate R_{30} until the meter reads about 3 μ A. The circuit is now set and the calibration is determined by R_{25} and R_{31} .

As would be expected the super-alpha pair Tr_4 and Tr_5 exhibit some temperature sensitivity but the author has found no problems; R_{29} and R_{30} need never be touched after the initial set-up, provided $C_{22a,b}$ are paper or polyester capacitors and D_1 and D_2 silicon planar diodes. Small drifts in the standing current through the meter do not affect calibration, which is highly predictable.

Two schemes have been included in the basic recorder for A-B monitoring and the one chosen will be determined by the pre-amplifier with which the recorder is used. The essential difference is that one scheme includes the A-B switch in the recorder, in the other this switching is a pre-amplifier function.

TABLE 3 Meter calibration details

calibration current μ A	calibration dB
50	+6
38.5	0
27	-6
16	-12
4.5	-18

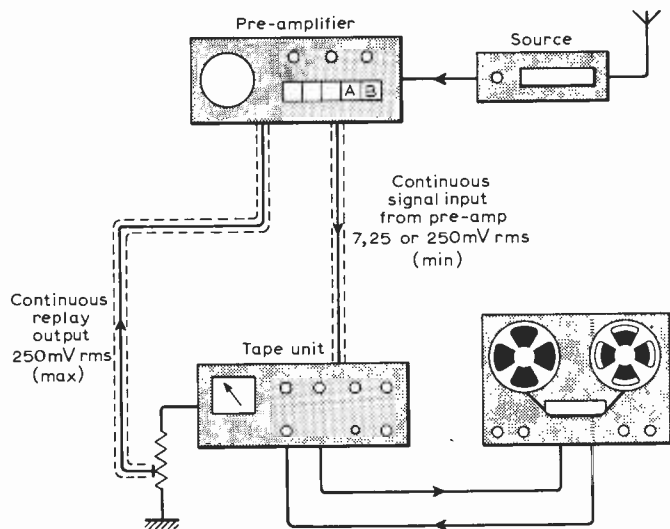


Fig. 13. A-B monitoring arrangement type 1. Input to top of RV_1 is 7mV. For output of 250mV r.m.s. link A + B, C + D; A' + B, C' + D'. For output of 25mV r.m.s. link A + D, A' + D'; connect R to A.

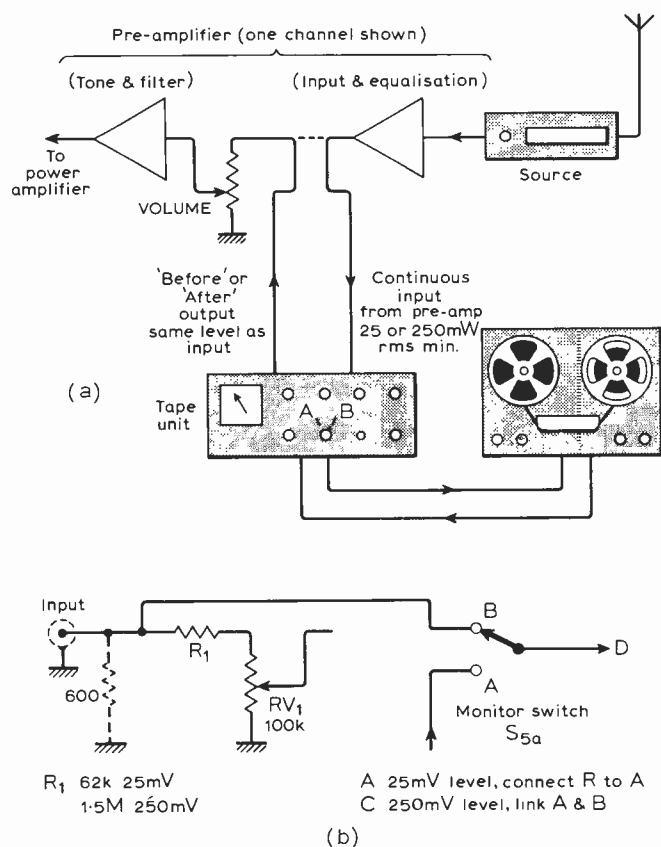


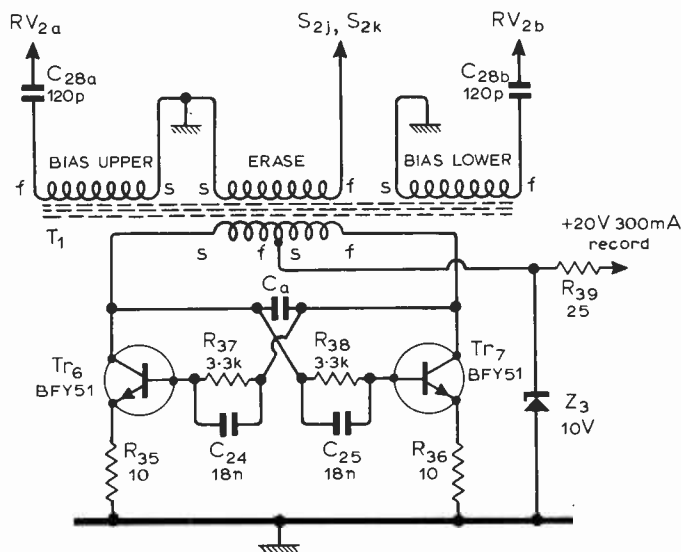
Fig. 14. A-B monitoring arrangement type 2. (a) The pre-amp is modified by breaking the lead shown dotted and re-routing as indicated. (b) Wiring for one channel.

Some amplifiers (Quad 33 and Leak for example) provide tape A-B switching. The scheme is shown in Fig. 13.

If this facility is not available then the scheme shown in Fig. 14 may be used; the input signal is passed to the recorder, which then either routes it back to the pre-amplifier 'B', or replays the signal on tape 'A'. The levels are arranged to be the same in this design, and the A-B function may be carried out at a 0dB level of 25 mV r.m.s. or 250 mV r.m.s. as shown in Figs 13 and 14, with reference to Fig. 3.

Bias and erase oscillator

As the bias network and erase heads present a reactive load to an erase oscillator it is a difficult problem to ensure that the



$C_{2a} = C_{26}$ Mono 0.13 μ s = Start of winding
 $= C_{27}$ Stereo 0.33 μ f = Finish " "

Fig. 15. Circuit diagram of erase oscillator.

calibration of the bias and erase current are retained on switching from stereo to single track recording. The most direct method of ensuring well defined currents is to use a master oscillator and two current mode output stages, this is described in Part 3; however the increased complexity and cost were not considered worthwhile in this recorder, in view of the excellent results which can be obtained, with care, from a multivibrator.

Fig. 15 shows the circuit diagram of the oscillator. Although it does not operate in Class D⁹ it is of a current switching type, the current being determined by the reflected load on the transformer primary and the supply voltage. Amplitude of oscillation is determined by Z_3 . The Brenell Mk6 deck is fitted with Bogen UL290 erase heads. These require 70 V r.m.s. at a current of 45 mA (100 kHz) the equivalent loss resistance is 10 k Ω , giving a power of 500 mW per channel. Bias requirement for the recording head is 24 V r.m.s. at 9.7 mA for each channel.

In order to minimize interference from Droitwich transmissions, or harmonics of 38 kHz from stereo decoders, the oscillator was designed to run at 107 kHz. This frequency is set by the total effective primary inductance of T_1 and C_{2a} ; which is switched to keep the frequency constant for mono or stereo recording. On mono a 10 k Ω load is switched across the erase head to keep the bias and erase currents unchanged, see Fig. 2. L_{2a} and L_{2b} form high impedance tuned rejectors to keep the h.f. bias out of the recording stage, some 100 mV pk-pk of bias waveform appears at C_{2a} .

The bias current is set by the capacitors C_{28a} and b , and the ganged potentiometer RV_2 gives +3, -5 dB variation in the bias current, around 0.7 mA. As the loaded Q of the tuned circuit is around 30 the third harmonic distortion should be less than 0.4%. In fact because the driving waveform on the transistor bases is not square the distortion is about 0.1%. This is quite adequate for a bias waveform and the author can detect no increase in the noise on virgin tape when the oscillator is switched on, in fact the increase is just less than 1 dB. C_{30} ensures that the bias decays slowly to demagnetize the recording heads. Winding details for all coils are given in Table 4.

Power supply

During playback the quiescent power requirement is about 60 mA at +20V and this rises to about 500 mA during recording. Fig. 16 shows the circuit of a regulator which will accept 30-60V d.c.

TABLE 4 Winding details for inductors

L_1	100 turns 36 s.w.g. enamel covered wire.
L_2	275 turns 36 s.w.g. enamel covered wire.
T_1	Primary 5 + 5 turns 26 s.w.g. enamel covered wire, bifilar wound. Erase winding 100 turns 34 s.w.g. enamel covered wire. Bias winding 100 + 100 turns 36 s.w.g. enamel covered wire, bifilar wound. Separate each winding with one layer of Sellotape or similar.

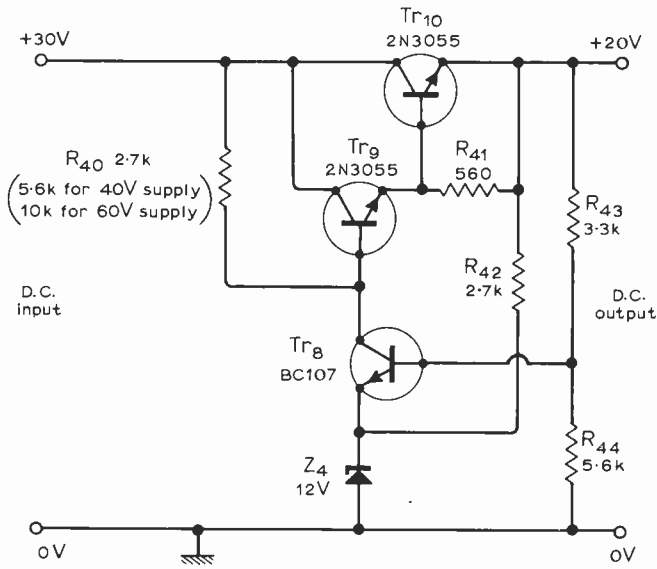


Fig. 16. Power supply input regulator. R_{40} must be altered as shown if higher supply rails are employed.

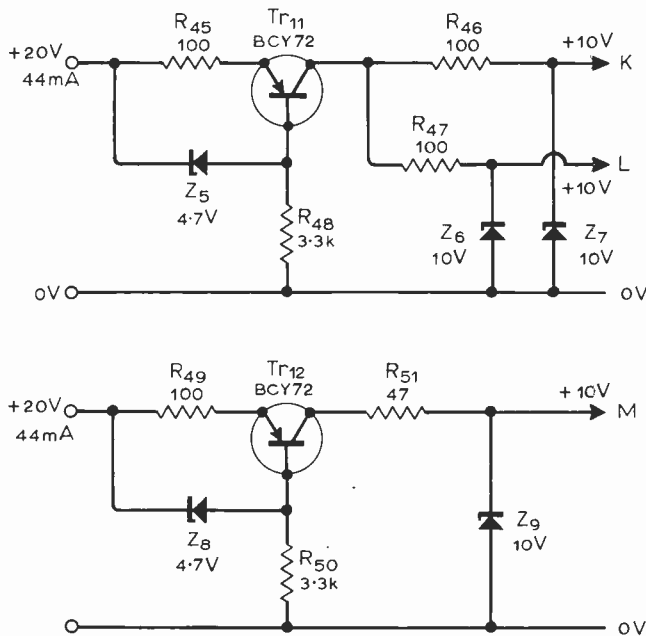


Fig. 17. 10V supply regulators.

input to give a 20V supply for the recorder. Tr_{10} should be mounted on a heatsink, or bolted to the chassis, with mica insulation to allow a safe operating temperature.

The integrated circuits use three separately derived +10V supplies to ensure stability and low crosstalk: the circuit is shown in Fig. 17. Tr_{11} and Tr_{12} are arranged as current sources. This affords protection to the i.c. and the effect of the low zener slope resistance with the high impedance of the current source is to reduce ripple by more than 66 dB. Tr_{11} and Tr_{12} should be fitted with cooling clips as they dissipate 160 mW each. Fig. 18 shows a simple power supply which may be used if d.c. is not available from the power amplifier.

Construction

The recorder unit circuits are built into an enclosed metal case. This affords rigidity and adequate screening. If a mains transformer is required it is recommended that this is not mounted in the case, but away from the unit and deck to maintain the low hum figures. A front panel carries all the controls detailed in Fig. 3 and the recording-level meter.

The circuits are built up on three Lektrokit boards 4in x 4 1/2in. The first board, which is shown in Fig. 19, carries the replay

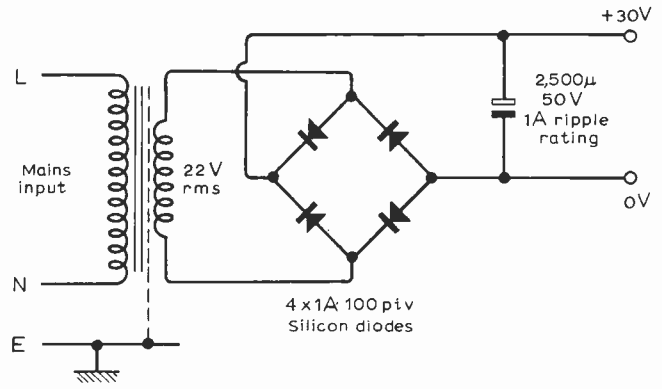


Fig. 18. Main supply circuit.

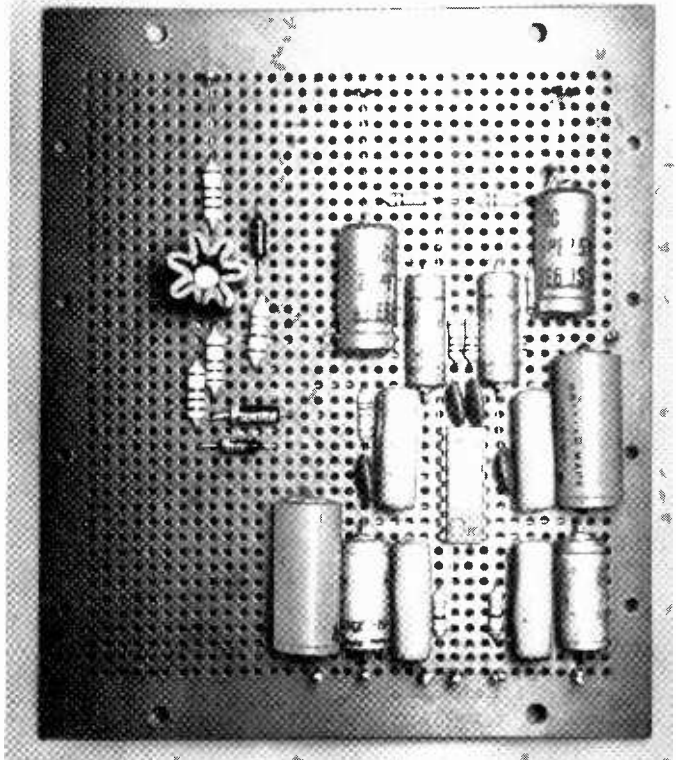


Fig. 19. Replay amplifier board.

circuits including i.c.B and the regulator Tr_{11} which derives the +10V supplies for K and L in Fig. 8. (Pt. 1).

No special precautions need to be taken with the integrated circuits; the networks C_{14} , R_{21} and C_{17} , R_{23} reduce the gain above 200 kHz. Stability is independent of layout.

Fig. 20 shows the second board which carries the meter circuits, record equalization and output stages, including i.c.A. Layout is again non critical.

The third board shown in Fig. 21 holds the erase oscillator, bias rejection and the regulator Tr_{12} to power i.c.A. In the author's version the 20V regulator of Fig. 17 was external, however there is plenty of room to mount it on this third board. All equalization components are mounted on the switch S_1 . To obviate any earthing or hum-loop problems the course adopted was to common on each board all earth and supply points and arrange only one connection to be made to these two rails. This means that the earth and supplies are common to the two channels but only one earth and supply lead is used per board. All signal leads in the unit are screened as indicated in Fig. 3 and the braid is earthed at only one end, that being the source. A wiring diagram is shown in Fig. 22.

It is not necessary to screen the circuits from each other. The erase board and the replay amplifiers should be arranged to be as

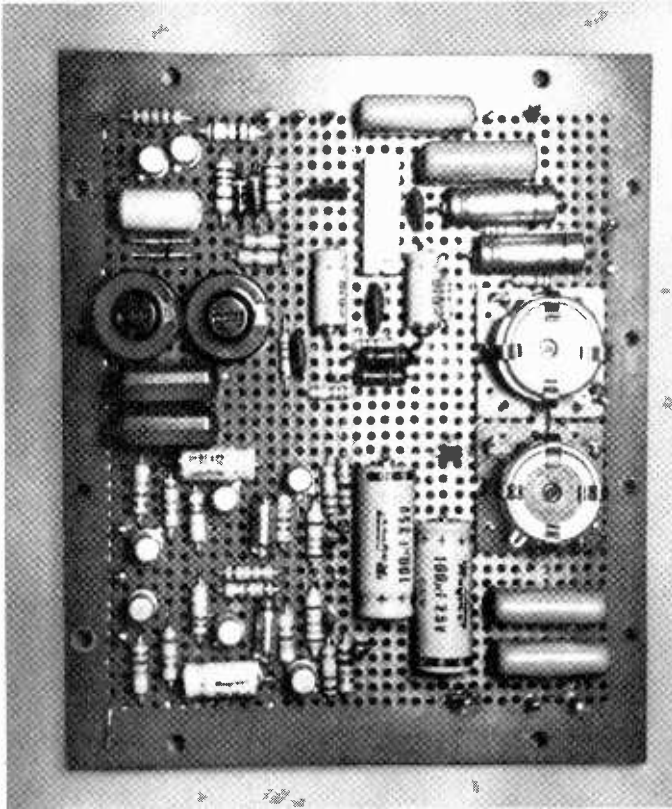


Fig. 20. Recording and meter drive circuits.

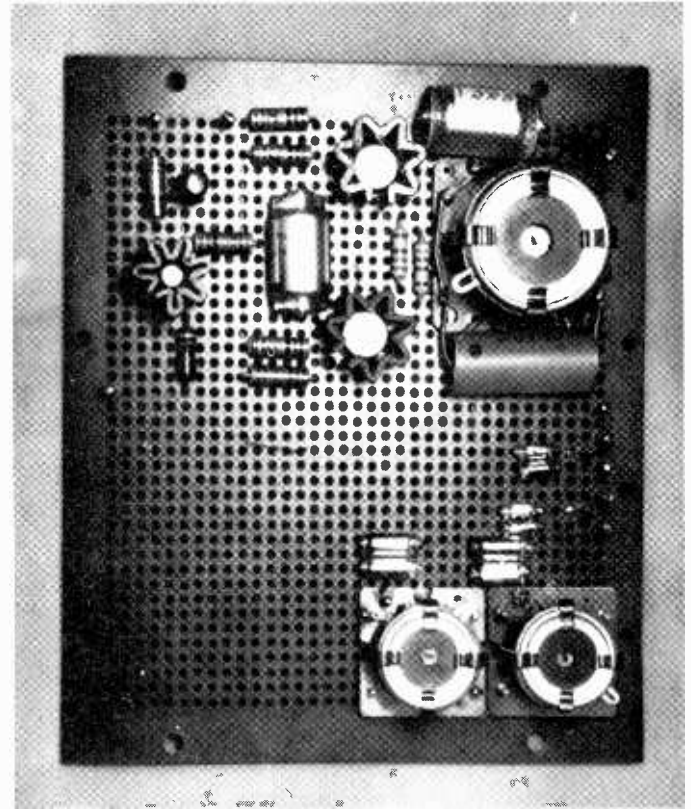


Fig. 21. Construction of erase oscillator.

far apart as possible but otherwise the layout is a matter of convenience. Inputs and outputs use Belling-Lee coaxial sockets for economy and good reliability. The Brenell deck supplies coaxial plugs on the record and replay heads, however the erase leads will have to be fitted with coaxial plugs.

Switching

Switching is carried out at the front panel and on the deck, the only switch which is repeated is the speed change. This was done to avoid the inevitable hum pick-up which would occur switching signals at the deck, and also it would not be possible to cater for the four speeds as the Brenell Mk6 deck uses a three-way switch with two capstans to provide $1\frac{7}{8}$, $3\frac{3}{4}$, $7\frac{1}{2}$ or $3\frac{3}{4}$, $7\frac{1}{2}$ and 15 i.p.s.

The record-play function is only required to switch the inputs to the recording equalization and to provide a 20 V rail to the recording circuits. For simplicity the Brenell deck is modified to give this +20 V rail when it is switched to record—which has an interlock button; the signal switching is then performed by a small low current relay, S_4 .

This modification is necessary because in standard form the record-off-play switch on the deck is arranged to short the erase heads on all but the record position. This is not desirable as it would inhibit the decaying bias waveform. Fig. 23 illustrates the modification. The spare wafer is used to switch the +20V supply.

Testing and setting-up

It is recommended that the circuits are tested individually before final power supply connections are made. D.C. levels should be checked using a good multimeter; check all supplies, and the output of each i.c. amplifier, which should be around +5.6V in each case. The d.c. conditions of the recording output stage can be checked by measuring the potential on the emitter of Tr_3 . This should be about +9.4V. If any discrepancy is noted, switch off immediately and check the circuit. Set up the meter as described. This may now be used when adjusting the bias rejection; otherwise use a high input impedance millivoltmeter or an oscilloscope. The bias rejectors L_2 allow a 5% adjustment of inductance and this should be more than enough to accept the expected uncertainty of oscillator frequency. Loose wiring or winding on L_2a or L_2b may move the resonant frequency, in which

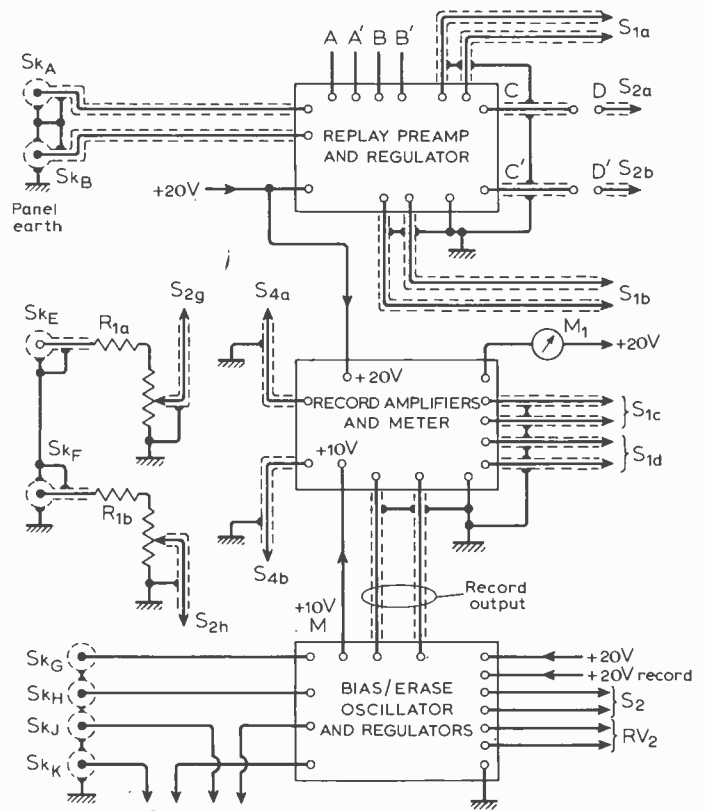


Fig. 22. Overall wiring diagram.

TABLE 5 Recording pre-emphasis component values

speed	boost frequency	amount of boost	C_1 nominal	R^*	overload margin of boost
ips	kHz	dB	μF	k Ω	dB
15	28	8.6	0.01	2.4	+5
7.5	15.5	18.25	0.02	22	-4
3.75	9	19.0	0.05	27	-5
1.875	6.5	19.0	0.08	27	+1

* depends on final Q of L_1

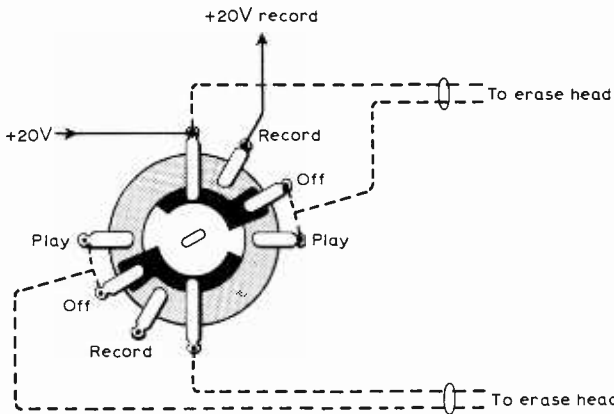


Fig. 23. Switch modification for Brenell Mk 6 deck. Remove wires shown dotted and add the wires shown solid.

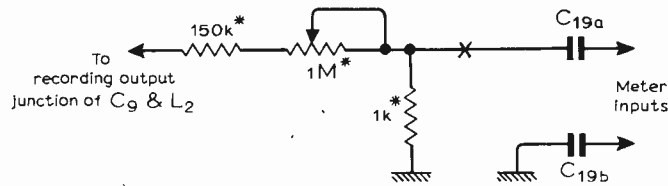


Fig. 24. Arrangement of meter to set up bias rejection. Begin with the potentiometer set to maximum, then reduce as null is approached. Components marked * are required only for setting up.

case C will have to be adjusted to find the null. Switch to record and using the arrangement of Fig. 24, or by another means, adjust L_2 and b for a minimum bias voltage at the recording amplifier—with all erase and record heads connected. The meter may now be wired, according to Fig. 3. Check that bias is present at the correct level and that about +3 -5dB variation is available. The recorder is now set up and ready to operate.

Choice of recording standard

Several replay standards can be used and these are shown in Fig. 10. The choice will depend upon the tapes to be replayed. All the measurements given here use either the C.C.I.R. standard or C.C.I.R. with high frequency lift. On replay some high-frequency boost can be applied by connecting C_x and R_y of the values given; note that this network must be switched for each speed requiring an extra two poles on S_1 . If tapes recorded to a different standard are to be used this lift must not be applied; however for recordings made on this machine the boost will compensate replay head losses at high frequencies and enable a true 'constant-induction' recording characteristic to be used, with less chance of high frequency overload.

It must be understood that having chosen a replay characteristic, then although a good approximation can be made, the amount of recording pre-emphasis which must be applied depends upon the tape used, the bias level—which incidentally will be different for each speed—and the form of response which is required, see Part 1.

Two examples are given here, maximally flat response with C.C.I.R. playback, maximally flat response with C.C.I.R. and high frequency lift. These two give low intermodulation distortion at high frequencies.

The component values are given in Table 5. The first response is shown in Fig. 1 and the dotted curve gives the response with high frequency lift on playback. With careful construction there is no reason why these specifications should not be met.

REFERENCE

9. Baxandall, F. J., "Transistor LC Sine-wave Oscillators". I.E.E. paper 2 978 E, February 1960.

Corrections to Part 1 on p. 595

Component List

Capacitors

- | | |
|------------------------------------|------------------------------------|
| C_1^* 0.47 μ paper† | C_{16}^* 64 μ 6.4V |
| C_2^* 25 μ 25V | C_{17}^* 0.01 μ disc ceramic |
| C_3^* 1 μ paper | C_{18}^* 150 μ 6V |
| C_4^* 12 μ 6V | C_{19}^* 0.47 μ paper |
| C_5^* 0.01 μ disc ceramic | C_{20}^* 100 μ 6V |
| C_6^* 0.02 μ disc ceramic | C_{21}^* 0.01 μ disc ceramic |
| C_7^* 0.05 μ disc ceramic | C_{22}^* 0.47 μ paper |
| C_8^* 0.08 μ disc ceramic | C_{23}^* 0.02 μ disc ceramic |
| C_9^* 1 μ paper | C_{24} 18n polystyrene |
| C_{10}^* 1000 μ 6.4V | C_{25} 18n polystyrene |
| C_{11}^* 0.47 μ paper | C_{26} 0.13 μ 2% paper |
| C_{12}^* 250 μ 6V | C_{27} 0.33 μ 2% paper |
| C_{13}^* 0.33 μ paper | C_{28}^* 120p polystyrene |
| C_{14}^* 0.01 μ disc ceramic | C_{29}^* 270p polystyrene |
| C_{15}^* 0.47 μ paper | C_{30} 500 μ 25V |

*two required

†for plastic in each case

Semiconductors

- | | |
|-------------------------------|---------------------------------|
| ICA CA3052 } supplied by | Tr_{11} BCY 72 } with cooling |
| ICB CA3048 } A. Marshall | Tr_{12} BCY72 } clips |
| Tr_1^* BC108 | $D_{1,2}$ 1N914 |
| Tr_2^* BCY72 | Z_1 3.3V 400mW |
| Tr_3^* BC108 | Z_2 4.7V 400mW |
| Tr_4 BC108 | Z_3 10V 1W |
| Tr_5 BC108 | Z_4 12V 400mW |
| Tr_6 BFY51 } with 5 cool- | Z_5 4.7V 400mW |
| Tr_7 BFY51 } ing clips | Z_6 10V 400mW |
| Tr_8 BC107 | Z_7 10V 400mW |
| Tr_9 2N3055 } plus | Z_8 4.7V 400mW |
| Tr_{10} 2N3055 } insulation | Z_9 10V 400mW |

* two required

Resistors All $\frac{1}{4}$ W 5% carbon unless otherwise stated.

- | | | |
|------------------------|---|------------------------|
| R_1^* see Fig. 3 | R_{21}^* 100 | R_{41} 560 |
| R_2^* 750 | R_{22}^* 5.6k | R_{42} 2.7k |
| R_3^* 120 | R_{23}^* 100 | R_{43} 3.3k |
| R_4^* 2.4k | R_{24}^* 10k | R_{44} 5.6k |
| R_5^* 22k | R_{25}^* 200 | R_{45} 100 |
| R_6^* 27k | R_{26}^* 100 | R_{46} 100 |
| R_7^* 27k | R_{27} 560k | R_{47} 100 |
| R_8^* 2.7k | R_{28} 560k | R_{48} 3.3k |
| R_9^* 18k | R_{29} 100k pre-set | R_{49} 100 |
| R_{10}^* 16k | R_{30} 100k pre-set | R_{50} 3.3k |
| R_{11}^* 910 | R_{31} 2.2k | R_{51} 47 |
| R_{12}^* 470 | R_{32} 100k | R_{52}^* 18k |
| R_{13}^* 10k | R_{33} 10k | R_{53}^* 4.7k |
| R_{14}^* 2.4k | R_{34} 1.3k | R_{54}^* 1M |
| R_{15}^* 47 | R_{35} 10 $\frac{1}{2}$ W metal oxide | R_{55}^* 330 |
| R_{16}^* } | R_{36} 10 $\frac{1}{2}$ W metal oxide | R_{56} to suit bulb |
| R_{17}^* } | R_{37} 3.3k | RV_1 100k log 2-gang |
| R_{18}^* see Fig. 8. | R_{38} 3.3k | RV_2 50k lin. 2-gang |
| R_{19}^* } | R_{39} 25 5W w/w | * two required |
| R_{20}^* } | R_{40} 2.7k | |

Inductors

- L_1 Plessey 905/1/01581/006 μ e 18mm pot core with base and clips, or Mullard LA2532. Two required.
- L_2 Plessey 905/1/01581/009 μ e 18mm pot core with base and clips, or Mullard LA2538. Two required.
- T_1 Plessey 905/1/01613/008 μ e 26mm pot core with base and clips, or Mullard LA2332. One required.

Switches etc.

- S_1 4-pole 4-way (minimum, see text) make-before-break
- S_2 9-pole 3-way make-before-break
- S_3 2-pole 3-way break-before-make
- S_4 low-current relay 2-pole change over.
- S_5 required for A-B monitoring arrangement type 2. 2-pole 2-way break-before-make
- Bulb (6V 40mA: R_{56} =330)
- Lektrokit board \times 3 plus pins.

Linear Scale Millivoltmeter

Using an output transistor with a constant current load improves an earlier design

by A. J. Ewins

Many circuit designs for a.c. millivoltmeters have been presented in this and other journals. At some stage in these circuits the a.c. signal is rectified in order that it may be measured by a d.c. instrument. The diodes used for rectifying the a.c. signal are extremely non-linear devices (i.e. their impedances increase as the current flowing through them decreases) and, if a linear scale is to be obtained, must be "swamped" with a large series resistance. The technique of placing the rectifying circuit in the negative feedback line of the amplifier is well known. In this type of circuit the output current through the measuring instrument is not entirely independent of the diodes' varying impedance.

The circuit to be described goes one step further than the negative feedback type by using a constant current load for the output transistor which makes the output current virtually independent of the rectifying circuit impedance.

Circuit operation

The circuit of the a.c. millivoltmeter is shown in Fig. 1. Observant readers of this journal will notice the close similarity between the writer's circuit and that of D. E. O'N. Waddington (*Wireless World*,

March, 1966). Acknowledgement must, indeed, be given to Mr. Waddington, for it was his circuit which inspired the present design.

The circuit of Fig. 1 can be considered in two separate sections; the first consisting of the input stages up to and including the range attenuator, and the second of the main amplifier and the meter circuit.

The first section provides a high input impedance to the millivoltmeter and a low output impedance for the range attenuator. It has an overall voltage gain of unity. The design of this stage is virtually a copy of Waddington's but with some component changes, to cater for the different supply voltage, and some additions to give good low-frequency stability, low noise and improve the frequency response.

The second section of the millivoltmeter provides the main amplification and has an output sensitivity of 1 mV (r.m.s.) for an output current of 50 μ A, d.c. The input impedance is of the order of 100 k Ω . It is this section of the millivoltmeter that is somewhat unconventional. The corresponding stage of Waddington's design is shown in Fig. 2 as an example of the more usual type of circuit.

A simplified equivalent circuit diagram (at 1 kHz) of the second section of Fig. 1 and Fig. 2 is shown in Fig. 3. The resistor R_c is

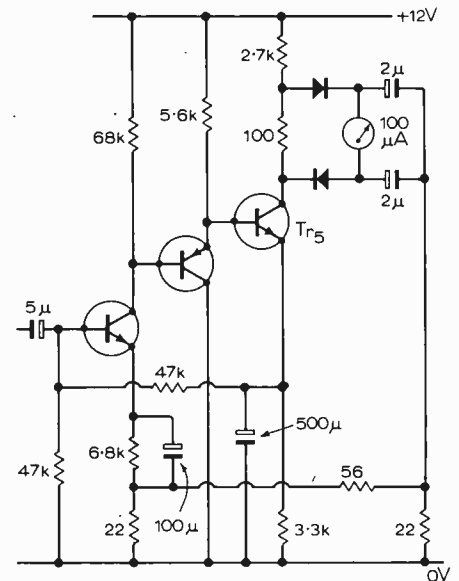


Fig. 2. The meter stage of Waddington's millivoltmeter.

the collector resistor of transistor Tr_5 in Waddington's circuit and is not present in the equivalent position in the circuit of Fig. 1. The first stage of the main amplifier (i.e. transistor Tr_3 in Fig. 1) is shown in detail but the second and third stages are simply shown as a current amplifier with a

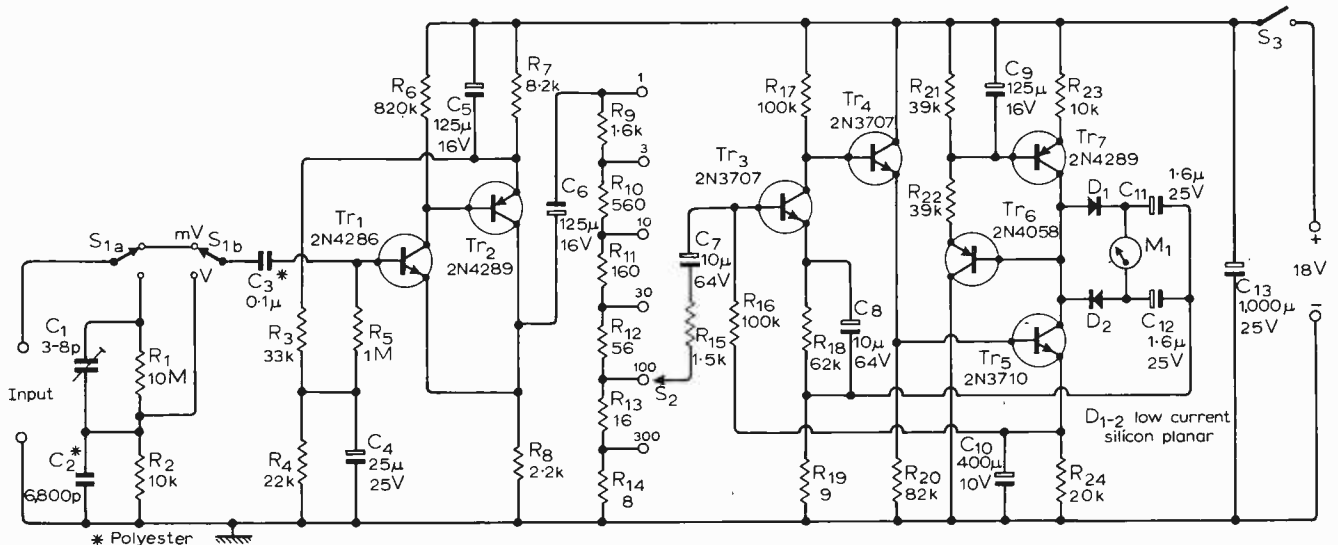


Fig. 1. The complete circuit diagram of the instrument. A components list appears on the facing page.

gain of *A*. From the detailed analysis of the equivalent circuit (see Appendix) the following relationship is obtained:

$$V_i = R_f \cdot i_L + \dots + \frac{(R_c + R_L + R_f)[r_b + (\beta + 1)(r_e + R_f)]}{A \cdot \beta \cdot R_c - (\beta + 1)R_f} i_L$$

If the last term of the above expression is made equal to *dR* · *i_L* then:

$$dR = \frac{(R_c + R_L + R_f)[r_b + (\beta + 1)(r_e + R_f)]}{A \cdot \beta \cdot R_c - (\beta + 1)R_f}$$

and

$$V_i = R_f \cdot i_L + dR \cdot i_L$$

Thus *i_L* = *V_i* / (*R_f* + *dR*), and if *dR* is very much less than *R_f*, *i_L* = *V_i* / *R_f*, which is the ideal relationship.

It will be seen from the above expression for *dR* that the larger the value of *R_L* the greater is the value of *dR*. Thus, it is possible that, as the current *i_L* falls and *R_L* increases due to the characteristics of the rectifying diodes, the term *dR* may increase from a negligible amount to a significant value. The effect of an increasing value of *dR* at low values of *i_L* is to modify *i_L* to some lower value, thus cramping the scale of the meter towards its zero end. It would not be fair to suggest that for the conventional type of negative feedback voltmeter this cramping is very serious. Indeed, the whole purpose of the negative feedback is to reduce this effect. From the expression for *dR* it will be seen that the greater the value of the open loop current gain, *Aβ*, and hence the greater the amount of negative feedback that can be applied, the smaller becomes the value of *dR*, reducing the effect of a varying *R_L* upon the linearity of the millivoltmeter. In a well designed millivoltmeter, such as Waddington's, the cramping effect at 10% of full scale deflection (f.s.d.) should not be more than about 1% of f.s.d. However, having said that, the relationship between *i_L* and *V_i* would be more nearly perfect if *i_L* were made independent of *R_L*. This is what the writer has attempted to achieve.

From a study of Fig. 1, it will be seen that the resistor, *R_c*, of Fig. 3 does not exist but is replaced by a constant current load

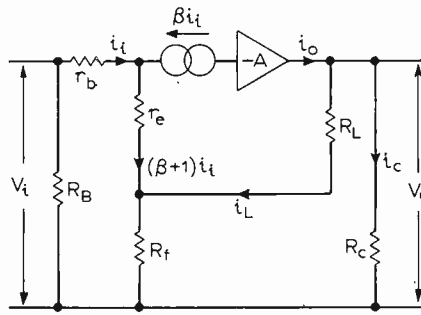


Fig. 3. Simplified equivalent circuit, at 1 kHz, of Figs. 1 (second stage only) and 2.

provided by transistor *Tr₇* whose output impedance is extremely high. Thus the equivalent value of *R_c* tends to infinity with the result that the value of *dR* tends towards the value *dR'*, where

$$dR' = [r_b + (\beta + 1)(r_e + R_f)] / A \cdot \beta$$

which is independent of *R_L*.

Now, it is not the intention to "pull the wool over the reader's eyes" by suggesting that the equivalent value of *R_c* is infinite; it is not. The load on the collector of *Tr₅*, in addition to the rectifying circuit, is the collector output impedances of transistors *Tr₅*, *Tr₆* and *Tr₇* and the input impedance of *Tr₆*, all in parallel. For the operating currents of these transistors the collector output impedances can be anything in the order of 1 to 3 MΩ. The input impedance of *Tr₆*, assuming a value for *h_{fe}* of not less than 100, will be approximately 3.9 MΩ. Thus, at best, the equivalent value of *R_c* may be considered to be a little less than 1 MΩ.

To demonstrate the effect of using a constant current load in place of the conventional resistor, the two simple circuits of Figs 4 and 5 were constructed. The sensitivity of these two circuits is determined by the un-decoupled emitter resistors of transistor *Tr₁*. With the value of this resistor set nominally at 9.1 kΩ the sensitivities are approximately 1 V for an output current of 50 μA through the d.c. measuring instrument. An alternating voltage of approximately 1 V at a frequency of 1 kHz was applied to the inputs of these two circuits,

individually, and adjusted for f.s.d. The input voltage was then attenuated in 10 dB steps and the meter's deflection, as a percentage of the f.s.d., noted. Table 1 shows the results obtained.

For the constant current loaded circuit (Fig. 5) the reading for an input of 10% of f.s.d. is in error by -4%, which is only -0.4% of f.s.d., while that for the conventionally resistor-loaded circuit is in error by -42%. The constant current circuit is thus as linear, without placing the rectify-

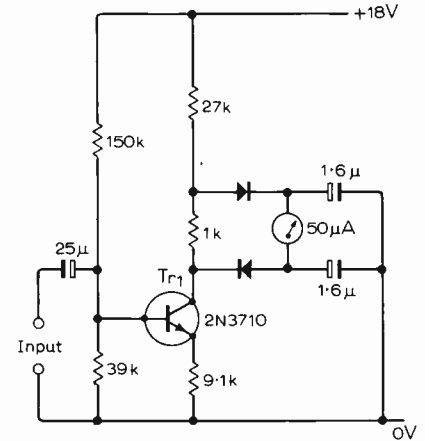


Fig. 4. Test circuit. Meter stage with resistive load.

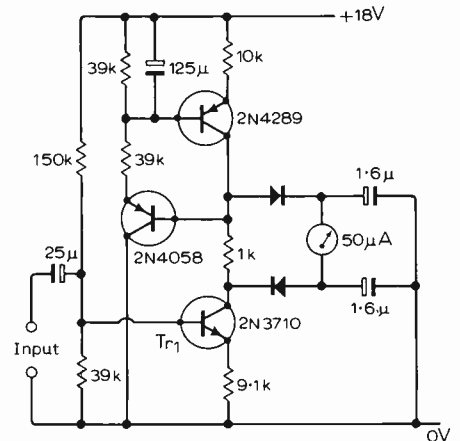


Fig. 5. Test circuit. Meter stage with constant current load.

Component list

Resistors

The prefix *R* and the suffix *Ω* have been omitted from this list.

1—10 M*	9—1.6 k*	17—100 k
2—10 k*	10—560*	18—62 k
3—33 k	11—160*	19—9Ω
4—22 k	12—56*	20—82 k
5—1 M	13—16*	21—39 k
6—820 k	14—8†	22—39 k
7—8.2 k	15—1.5 k	23—10 k
8—2.2 k	16—100 k	24—20 k

* 1% tolerance for maximum accuracy.

† 2 × 16 Ω in parallel.

§ 2 × 18 Ω in parallel.

All resistors 0.5 W. 5% tolerance unless otherwise specified.

Capacitors

The prefix *C* and the suffix *F* have been omitted from the list below.

- 1—3—8 p. beehive trimmer
- 2—6800 p, polyester, 160 VW
- 3—0.1 μ, polyester, 250 VW
- 4—25 μ, 25 VW*
- 5—125 μ, 16 VW*
- 6—125 μ, 16 VW*
- 7—10 μ, 64 VW*
- 8—10 μ, 64 VW*
- 9—125 μ, 16 VW*
- 10—400 μ, 10 VW*†
- 11—1.6 μ, 25 VW*
- 12—1.6 μ, 25 VW*
- 13—1000 μ, 25 VW

* Mullard type C426.

† 2 × 200 μF in parallel.

Semiconductors

The prefix *Tr* has been omitted from the list below.

- 1—2N4286
- 2—2N4289
- 3—2N3707
- 4—2N3707
- 5—2N3710
- 6—2N4058
- 7—2N4289
- 8—2N4289
- 9—*D₁*, *D₂*—Any low current silicon planar diode.

Miscellaneous

- Die-cast box
- S1—2 pole, 2 way
- S2—Single pole, 6 way
- S3—Single pole make
- 4 terminals
- Veroboard 0.15 in pitch
- 2 × type PP6 battery
- Avo model-9 50 μA meter movement

open circuit equivalent noise voltage reliably. Short circuited, the noise level was equivalent to about 5 μ V as indicated by the 50 μ A meter. With a 100 k Ω resistor connected across the input terminals the equivalent noise voltage rose to a little more than 10 μ V.

APPENDIX

Analysis of the equivalent circuit, Fig. 3: From the equivalent circuit diagram of Fig. 3, the following equations can be formed:

$$i_0 = A \cdot \beta \cdot i_i = i_L + i_c \quad (1)$$

$$V_0 = R_c \cdot i_c = R_L \cdot i_L + R_f [i_L + (\beta + 1)i_i] \quad (2)$$

$$V_i = [r_b + (\beta + 1)r_e]i_i + R_f [i_L + (\beta + 1)i_i] \quad (3)$$

From eqn. 2 we have:

$$R_c \cdot i_c = (R_L + R_f)i_L + R_f(\beta + 1)i_i \quad (4)$$

From eqn. 1 we have:

$$i_c = A \cdot \beta \cdot i_i - i_L \quad (5)$$

Substituting for i_c from eqn. 5 and inserting in eqn. 4.

$$R_c(A \cdot \beta \cdot i_i - i_L) = (R_L + R_f)i_L + R_c(\beta + 1)i_i$$

thus,

$$i_i = \left\{ \frac{(R_c + R_L + R_f)}{A \cdot \beta \cdot R_c - (\beta + 1)R_f} \right\} \cdot i_L \quad (6)$$

If i_i from eqn. 6 is substituted into eqn. 3 the following relationship is obtained:

$$V_i = R_f \cdot i_L + \dots + \frac{(R_c + R_L + R_f)[r_b + (\beta + 1)(r_e + R_f)]}{A \cdot \beta \cdot R_c - (\beta + 1)R_f} i_L \quad (7)$$

If the base bias resistor, R_b , is ignored the input impedance of the amplifier is given by V_i/i_i . From eqn. 3, by rearrangement:

$$V_i = [r_b + (\beta + 1)(r_e + R_f)]i_i + R_f \cdot i_L \quad (8)$$

Substituting for i_L from eqn. 6 in eqn. 8 and rearranging:

$$V_i/i_i = r_b + (\beta + 1)(r_e + R_f) + \frac{R_f[A \cdot \beta \cdot R_c - (\beta + 1)R_f]}{(R_f + R_L + R_c)} \quad (9)$$

and if R_c is very much greater than R_L and R_f , then V_i/i_i is approx. equal to $R_f \cdot A \cdot \beta$.

appointed sole U.K. agents for **Lorlin Industries Inc.**, of Danbury, Connecticut, manufacturers of a range of programmable automatic transistor and diode test sets.

The M.E.L. Equipment Co. Ltd., of Manor Royal, Crawley, Sussex, have been awarded a **£1M contract** for the supply of 100-W s.s.b. transmitters, receivers and associated equipment for use in the 1.5-30 MHz band by the British Royal Navy.

In order to strengthen their marketing facilities in Europe, and in Germany in particular, **Dynamco** have announced a new company called **Sintrom-Dynamco GmbH** with headquarters in Frankfurt. Sintrom's sales director David Hatfield will have overall control of the company and Fernie de Weerd of Dynamco will be manager.

Books Received

Sun Earth and Radio, by J. A. Ratcliffe, F.R.S. This book, in the World University Library series, is about the two layers of the upper atmosphere—the magnetosphere and the ionosphere—which affect the behaviour of radio waves. Two introductory chapters describe the investigation of these atmospheric layers by radio waves from ground level and from space vehicles, and the remainder of the book examines the facts we have. Two appendices provide an analysis of some fundamental terms and ideas, and gives an outline of very recent advances. There is a bibliography and an index. Pp. 256. Price 35s for hardback, 16s for paperback. Weidenfeld and Nicolson, 5 Winsley Street, Oxford Circus, London W.1.

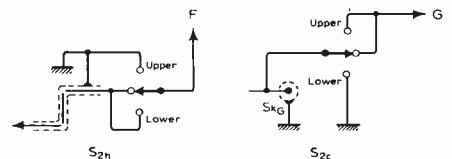
Dictionary of Telecommunications, by R. A. Bones. Technical terms which are specific to telecommunication theory and practice are given clear and concise definition, along with a wide range of terms met in general electronics. This is a very useful reference work, Several appendices tabulate frequency bands, spectra, and codes used in communications. Pp.200. Price 45s. Butterworth & Co. (Publishers) Ltd, 88 Kingsway, London W.C.2.

Corrections

High-quality Tape Recorder: The following amendments should be made to Part I (November issue).

In Fig. 1 the bias levels given in the upper graph should have minus signs in front of them and refer to the 0dB = 700 μ A in the lower graph.

In Fig. 3 S_{2h} and S_{2c} should be drawn as below.



The mid position of S_2 is the stereo mode. On page 526 line 28 should read '... the just noticeable difference is greater than 1 phon.'

In Fig. 4 the i.c. is viewed from above.

In Fig. 7 the final equation should read

$$\frac{I_0}{V_{in}} = \frac{h_{21}}{1 + h_{21}} \cdot \frac{1}{R} \cdot \frac{1}{R}$$

In Fig. 9 the ordinate should be $\sqrt{2}/\text{Hz}$.

Active Filters—15: In Fig. 19 (g) of Part 15 (November issue) the polarity of the input terminals of the amplifier should be reversed, as in Fig. 16 (b) of Part I (August 1969).

Announcements

In future the B.B.C. World Radio Club will be broadcast in the World Service on Thursdays at 12.45 G.M.T. with repeats on Fridays at 23.45 and Sundays at 08.15.

In line with changes in terminology brought about by international standardization and the move to the metric system, the British Fractional Horsepower Electric Motor Manufacturers' Association has been renamed the **Small Electric Motor Manufacturers' Association**.

At a meeting in October of interested parties it was decided that a **new importers' association** should be formed which would start operations on January 1st 1971. The organization will be called International Radio and Electrical Distributors Association (I.R.E.D.A.).

A commercial television station, ZFB-TV, is the first to transmit **colour in Bermuda**. The station, which is owned by the Capital Broadcasting Co. Ltd, employs the N.T.S.C. standard and was set up by Thompson Television (International) Ltd., of London.

The Boards of **Plessey** and **BSR** have announced that it was not possible to agree on terms for the acquisition of BSR by Plessey. Discussions have therefore ceased.

A new company, called **Sonicstore Ltd** of Cornhill, Ilminster, Somerset, is to undertake research, development and manufacture of sonic and ultra-sonic **delay lines**. Sonicstore has concluded an agreement with Mullard for the transfer of their technical and commercial activity in the field of custom-built solid-state delay lines for industrial applications.

Competa International Products, a division of A. F. Bulgin & Co., have been appointed sole U.K. agents for Chicago Switch Inc., manufacturers of precision,

miniature, slide, leaf, rocker and multi-pole modular switches.

Semicomps Ltd (5 Northfields Estate, Beresford Avenue, Wembley, Middx.) have been appointed sole U.K. agents for capacitors manufactured by **Rubycon** the capacitor manufacturing division of Shin-Ei Tushin Kogyo Co. Ltd. of Tokyo.

R.C.A. entertainment receiving valves can now be obtained from the following newly appointed distributors: Semicomps Northern Ltd (Kelso), REL Ltd (Hitchin), and ECS Ltd (Windsor).

Mullard semiconductor products can now be obtained from Celdis Ltd, 37-39 Loverock Road, Reading, Berks, RG3 1ED, who have been appointed as a Mullard distributor. Later the agreement will be extended to cover other Mullard components.

Decca audio equipment will now be available to non-franchised dealers through A.C. Farnell Ltd in London, Leeds and Sheffield. London address is, 25a Hermitage Road, London N.4.

The Plessey Company Ltd has signed an agreement with Oki Electric Industry Company of Japan, whereby **Oki** will supply to **Plessey** specification certain communications equipment.

ITT Components Group Europe has signed an agreement to sell the 5170 **modular connectors** manufactured by Ultra Electronic (Components) Ltd.

Granger Associates Ltd, of Weybridge, have been appointed exclusive sales representatives for **Sparta Electronic Corporation**, of Sacramento, California. Sparta products include the Bauer range of f.m. and a.m. broadcast transmitters and a line of studio equipment.

Euro Electronic Instruments Ltd, of Shirley House, 27 Camden Road, London N.W.1, have been

Electronic Building Bricks

7. Controlling electron flows in circuits

by James Franklin

If we want to represent and process information electronically we must be able to control electron flows in circuits. In some cases it may be necessary to set a current at a fixed value; in others we may wish to vary a current in proportion to some quantity conveying information, such as the air pressure variations in speech or music sounds. In yet other situations we may wish to make a current vary with time in some characteristic fashion (e.g. to conform to a mathematical law such as a square law or logarithmic law); or, if we already have a current varying in a particular fashion, we may wish to modify this to something else (e.g. to some other mathematical law).

There are two possible ways of controlling electron flows. One of them is to vary the source of electromotive force that is making the electrons move in the first place (Part 5). The second is to set up an *opposing* electromotive force, which is variable, in the path of the electrons. In practice the e.m.f. source is usually fixed—a battery or the mains supply. Our main concern is therefore with methods to produce opposing e.m.f.s which can be varied in order to control the current given by the fixed e.m.f. source.

One method of setting up an opposing e.m.f. is to introduce what is called *resistance* in the flow path of the electrons (Fig.1(a)). This form of hindrance to electron flow is analogous to mechanical friction, and the way it controls the

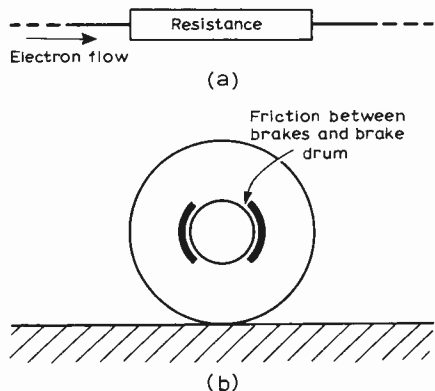


Fig. 1. A component introducing resistance to electron flow is shown symbolically at (a). A mechanical analogy of this property is friction, as used in the brakes of a car (b).

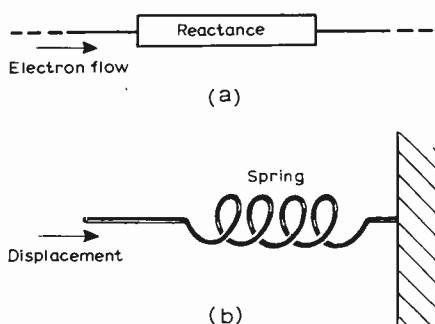


Fig. 2. Electron flow can also be controlled by reactance (a); mechanical analogies are the inertia of a mass or the stiffness of a spring (b).

movement of electrons is roughly comparable with the way friction in the brakes of a car controls the rotation of the wheels (Fig.1(b)). Just as with friction some of the mechanical energy is converted into heat energy, so with resistance some of the electrical energy is converted into heat energy.

Another method of producing an opposing e.m.f. is by introducing *reactance* in the electron flow path (Fig.2(a)). This is analogous to the mechanical inertia of a mass (e.g. the initial "unwillingness to move" of a vehicle when power is first applied to its wheels) or to mechanical stiffness (e.g. the "pushing back" property of a spring (Fig.2(b)). In reactance the electron flow energy is not converted into heat but into electric or magnetic field energy (Part 5). Because this field energy will at some point automatically change back into electron flow energy one can think of a reactor* as an energy store.

Let us see in more detail how a flow rate of electrons can be controlled in a circuit. For this purpose we will take the elementary circuit Fig.3 of Part 5, which is reproduced here as Fig.3(a). The rectangular symbol not explained in Part 5 is here an electrical component which introduces resistance to electron flow—a resistor. We have also added a meter which measures volts (see Part 5) and which can be connected across the resistor.

An hydraulic analogy of Fig.3(a) is shown at (b). This is a water circulating

system containing a wad of some fibrous or particulate material (e.g. a filter) which provides mechanical friction and is analogous to the resistor in (a). If the wad of material were not present the pump would give a high flow rate of water round the circuit. With the wad inserted, however, the water can only percolate slowly through the interstices of the material and consequently the flow rate is limited and determined by the amount of material. A point to notice here is that on the left-hand end of the wad, because the water is held back by it, a high pressure is built up. On the right-hand side of the wad there is nothing obstructing the flow of water back to the pump so the pressure there is relatively low. Thus, across the wad a pressure difference is set up. And the greater the resistance to flow offered by the wad the greater the pressure difference.

In the corresponding electrical circuit (a) we have a source of e.m.f. causing a flow of electrons in a circuit containing the resistor and a milliammeter to measure the current. The resistor in the circuit behaves in a similar manner to the wad in (b) in that, with a fixed source of e.m.f., it limits the flow rate of electrons (current) in the circuit. The flow rate determined by the resistor is measured by the milliammeter. If we changed the value of the resistor—that is, changed the resistance or "current hindrance"—we would obtain a different current reading on the milliammeter. Thus the electron flow rate can be controlled to make it whatever value we wish.

Another way in which the resistor resembles the wad in (b) is that it causes a pressure difference to be set up across it, though in the electrical form this is called a *potential difference*. This potential difference (normally abbreviated to p.d.) is measured in volts, and so is similar to the electromotive force. Thus the voltmeter in Fig.3(a) measures the p.d. between the two ends of the resistor.

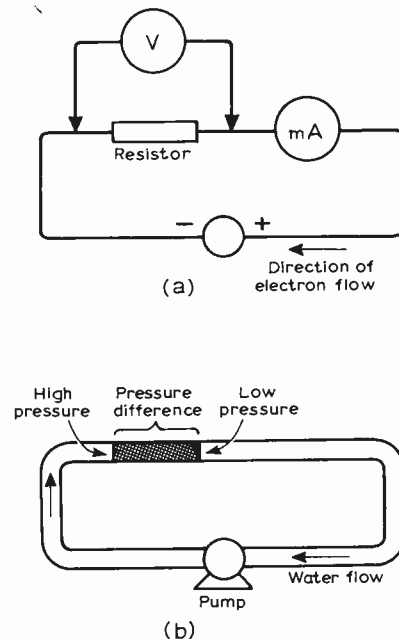


Fig. 3. At (a) is a simple circuit which controls electron flow rate by means of resistance; at (b) is an hydraulic analogy (without the meters).

*Common examples in electronics are coils and capacitors.

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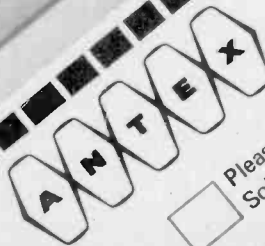
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30-MHz Video Amplifier with Variable Power Law

by A. M. Pardoe*

This article describes a video amplifier which accepts pulses with dynamic range up to 300:1 and compresses to 20:1 dynamic range necessary for intensity-modulated c.r.t. in p.p.i. display. Circuit uses transistor rather than diode as logarithmic element.

In radar work it is important that the dynamic range of the received signal is adjusted to match that of the cathode-ray tube—if it is an intensity modulated display. It is also necessary in television work to be able to match the dynamic range of the scene or recording to be televised to that of the c.r.t. in the receiver.† The ideal method of controlling the dynamic range of the signals, called gamma, and preserving the contrast between signals at different amplitude levels is to use an amplifier having a variable power law characteristic; i.e. where $V_{out} = V_{in}^\gamma$ where γ is variable. As this can be expressed by taking the logarithm of the input signal, multiplying it by γ and then

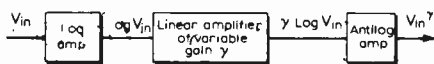


Fig. 1. Basic idea of variable power law amplifier.

antilogging the result, it is possible to perform this as shown in Fig. 1.

The advantage of a γ -controlled amplifier of this kind over a simple logarithmic amplifier is that the ratio between signal levels and thus the contrast is independent of signal level. That is, if two input signals differ in amplitude by a factor of n , then at the output the ratio between the two will be equal to n^γ (see for example Fig. 2).

Transfer functions

From the basic diode equation, the transfer function of a log amplifier (from Appendix A) is

$$\Delta V_{out} = \frac{1}{c} \log \frac{\Delta I_{in}}{I_{01} + I_{b1}} \quad (1)$$

where I_{b1} is the bias current and I_{01} is the leakage current. In the circuits described a transistor has been used as the logarithmic element rather than a diode, because it maintains a logarithmic characteristic over

a greater number of decades.

The log amplifier is followed by a linear amplifier of controllable gain γ . Output from this amplifier will therefore be $\Delta V_2 = \gamma \Delta V_1$. From eqn 1 therefore,

$$\Delta V_2 = \frac{\gamma}{c} \log \frac{\Delta I \Omega}{I_{01} + I_{b1}} \quad (2)$$

The transfer function of an antilog amplifier (from Appendix B) is

$$\Delta V_{in} = \frac{1}{c} \log(\Delta I_{out}/I_{b2})$$

From eqn 2

$$\frac{\gamma}{c} \log \frac{\Delta I_{in}}{I_{01} + I_{b1}} = \frac{1}{c} \log(\Delta I_{out}/I_{b2})$$

If the bias currents are adjusted to be equal and if the leakage currents are small, then $I_{b1} = I_{b2} = I_b$, and $I_{01} \ll I_b$. Also

$$\gamma \log(\Delta I_{in}/I_b) = \log(\Delta I_{out}/I_b)$$

and

$$\Delta I_{out} = (\Delta I_{in})^\gamma \cdot I_b^{1-\gamma}$$

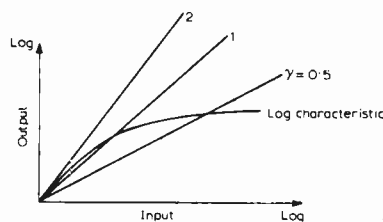


Fig. 2. Transfer characteristic showing that in log amplifier contrast will alter with signal level.

Thus the log amplifier needs to be current driven and the antilog amplifier voltage driven (Figs 3 and 4). Bias current is required to keep the log and antilog transistors on and as these currents appear on both sides of the transfer function, making them equal simplifies matters.

Design considerations

General specification of this circuit was to accept maximum signals of 2 V into 68 Ω with a varying dynamic range from 300:1 to 20:1, to compress them to a dynamic range of 20:1 and to make the final stage capable of producing 2 V into 68 Ω . Therefore a range of varying gamma from 0.5 to 1 was required.

Log amplifier. Because the input dynamic range to the log amplifier was to be 50 dB, and it was found by experiment that the minimum bias current through the log transistor that would maintain the required bandwidth was 30 μ A, maximum driving current required had to be 10 mA.

Bias through the log transistor was set by adjusting the non-virtual earth side of the long-tailed pair in the log amplifier (see Fig. 5). The response of the log amplifier

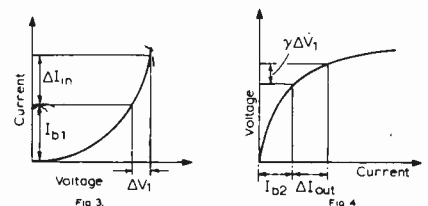


Fig. 3. Log amplifier transfer function.

Fig. 4. Antilog amplifier transfer function.

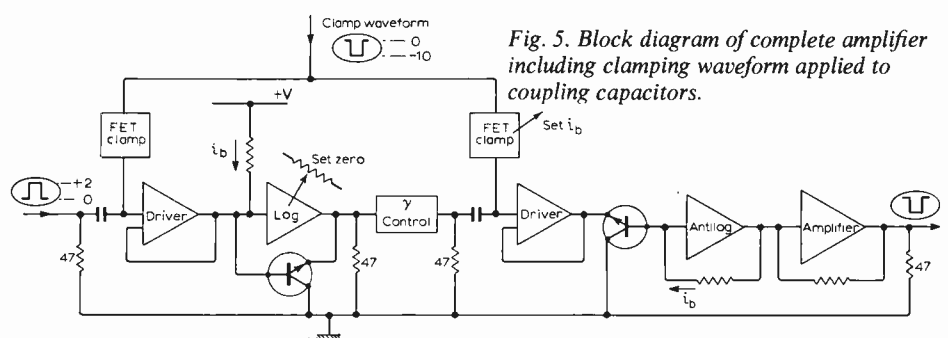


Fig. 5. Block diagram of complete amplifier including clamping waveform applied to coupling capacitors.

*Royal Radar Establishment, Malvern, Worcs.

†Kundu, P. and Banerji, S. 'Transistorized non-linear function generator', *Industrial Electronics*, vol. 2, no. 1, Jan. 1964, pp. 35-41.

needed to be adjusted so there was virtually no positive-going overshoot, as this would tend to cut off the feedback transistor and the amplifier would become unstable. To set up the bias current in the log amplifier, a sensitive centre-zero microammeter was switched into the live end of the 220 Ω resistor (R_1) feeding the amplifier and the control adjusted for zero reading; the bias being set by the 330 kΩ resistor (R_2) to the virtual earth (Fig. 6). In the antilog case, the meter was inserted in the collector to virtual-earth line and the potentiometer controlling the clamping voltage adjusted for 30 μA bias. Signal input had to be disconnected when setting the bias.

As any random change in current into the log amplifier will produce an output voltage, a very stable current driver is required. It was decided to use a voltage amplifier with a stable output and to drive the log amplifier through a resistor to its virtual earth, thus defining the current accurately. The value of this resistor is determined by the required response of the amplifier and by the voltage swing available.

If the quiescent output of the log amplifier is not to move by more than 10%, the output of the current driver must be stable to better than ±1 mV. As the driver was to be fed from other amplifiers whose outputs were not particularly stable, the input was a.c. coupled. This raised a further problem because the coupling capacitor acquired a charge depending on the p.r.f. and pulse length of the applied signals.

As the incoming signals were at a fixed p.r.f. there was a short dead time when no signals appeared and during this time the capacitor could be discharged. This was done by applying a clamping waveform to an f.e.t. connected between the capacitor and earth so discharging it once a cycle.

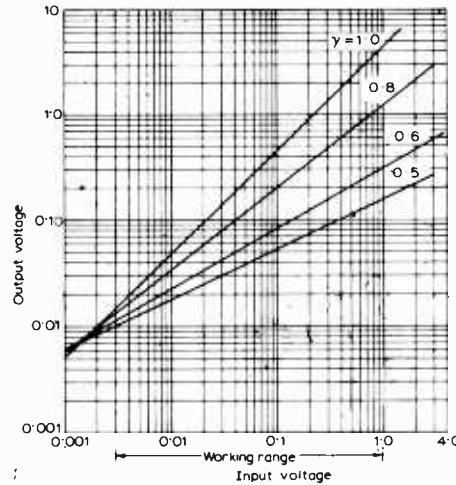


Fig. 8. Overall transfer characteristic using 500-ns pulse and bias of 30 μA.

Antilog amplifier uses a.c. coupling and clamping between the log and antilog amplifiers as the log amplifier output was still not stable enough to drive the antilog stage. In addition, so that temperature variations in the antilog transistor did not affect the output level, the voltage to which the coupling capacitor was clamped was determined by the voltage drop across an identical transistor in thermal contact with the antilog transistor. The voltage applied to the equalizing transistor was variable and determined the bias current through the antilog transistor, as shown in Fig. 7. Thus any variation in the voltage required to pass a particular current through the antilog transistor was automatically controlled by the input quiescent voltage.

Overall rise time of circuit is 25 ns and transfer function is shown in Fig. 8. The

gamma control took the form of a simple π-network switched attenuator so the dynamic range could be reduced. **Alternative log amplifier.** Towards the end of this work a hybrid 100 MHz operational amplifier—a Zeltex type 147—became available through Advance Electronics. One of these was used in place of the designed log amplifier, as shown in Fig. 9 and performed equally well.

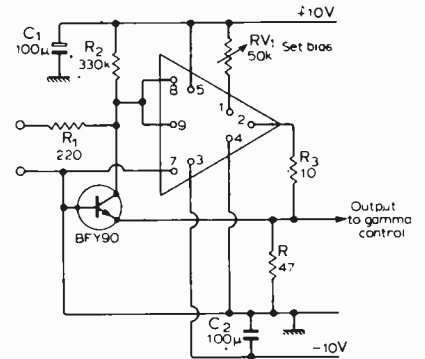


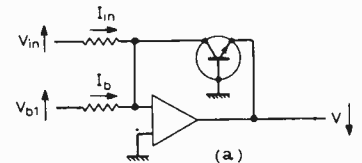
Fig. 9. Log stage using hybrid operational amplifier with external log transistor. Similar i.c. was tried in the antilog stage, but bandwidth could not be obtained.

Acknowledgement

The author thanks W. M. Nixon, who wrote the Appendices, and R. C. Bowes and T. Snowball for their valuable advice and suggestions in this work.

APPENDIX A

Log circuit transfer function



The basic diode equation is $I = I_0 (\exp cV - 1)$, where I is the signal current, I_0 the leakage current and V is the voltage across the diode. If $I = I_{in} + I_{b1}$, then

$$\frac{1}{c} \log \left(\frac{I_{in} + I_{b1}}{I_{01}} \right) + 1 = V$$

Assume $V = V_0$ at $I_{in} = 0$ and $\Delta V = V - V_0$ which is the output voltage pulse for input pulse I_{in} , then

$$\begin{aligned} \Delta V &= \frac{1}{c} \log \left(\frac{I_{in} + I_{b1} + I_{01}}{I_{01}} \right) \\ &\quad - \frac{1}{c} \log \left(\frac{I_{b1} + I_{01}}{I_{01}} \right) \\ &= \frac{1}{c} \log \left(\frac{I_{in}}{I_{b1} + I_{01}} + 1 \right) \end{aligned}$$

If $I_{in} \gg I_{b1} + I_{01}$, then

$$\Delta V = \frac{1}{c} \log \left(\frac{I_{in}}{I_{b1} + I_{01}} \right)$$

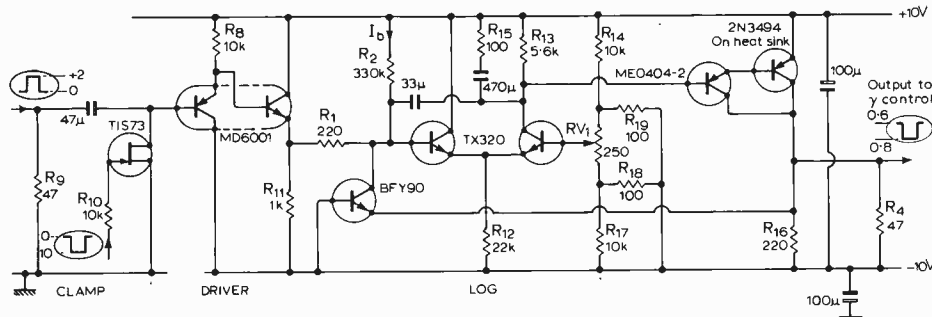


Fig. 6. Log amplifier circuit showing bias current setting arrangement.

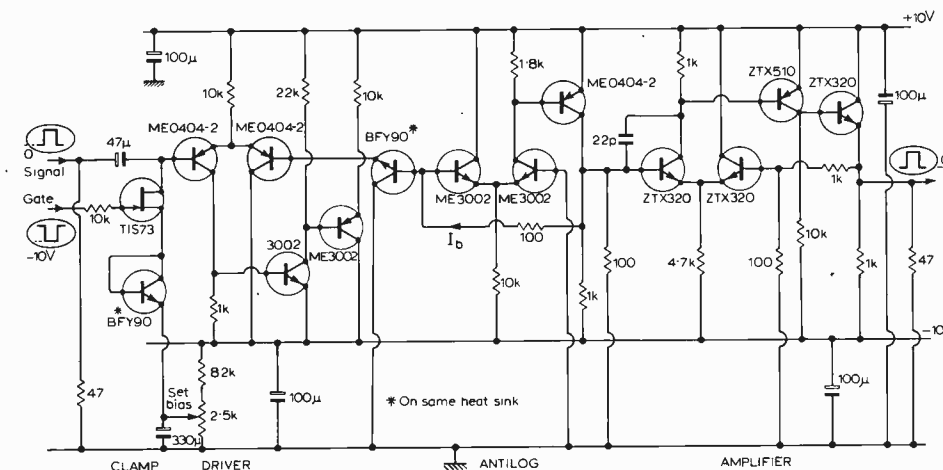
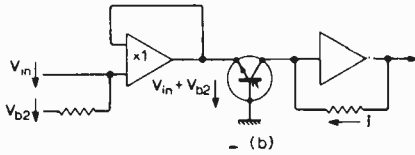


Fig. 7. Antilog stage circuit. Matched pair of antilog transistors minimises thermal drift.

APPENDIX B

Antilog circuit transfer function



Diode equation is $I = I_{02} (\exp cV - 1)$. If $V = V_{in} + V_{b2}$, where V_{in} is the pulse voltage and V_{b2} is the bias, and if ΔI_{out} is the pulse output current, then

$$\Delta I_{out} = I_{02} [\exp c(V_{b2} + V_{in}) - 1 - \exp cV_{b2} + 1]$$

$$= I_{02} \exp cV_{b2} (\exp cV_{in} - 1)$$

$$V_{in} = \frac{1}{c} \log \left(\frac{\Delta I_{out}}{I_{02} \exp cV_{b2}} + 1 \right)$$

Let $I_{b2} = I_{02} (\exp cV_{b2} - 1)$ and if $I_{02} \ll I_{b2}$, then $\exp cV_{b2} - 1 \approx I$ and $I_{02} \exp cV_{b2} \approx I_{b2}$. Therefore

$$V_{in} = \frac{1}{c} \log \left(\frac{\Delta I_{out}}{I_{b2}} + 1 \right)$$

and if $\Delta I_{out}/I_{b2} \gg 1$, then

$$V_{in} = \frac{1}{c} \log (\Delta I_{out}/I_{b2})$$

December Meetings

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

LONDON

1st. I.E.E./I.E.R.E.—Colloquium on "On-line data capture and verification" at 14.30 at 9 Bedford Sq., W.C.1.

1st. I.E.E./I.Meas.Control/I.Mech.E.—"Feedback control—the evolution of an idea" by Prof. A. G. J. MacFarlane at 17.30 at Savoy Place, W.C.2.

2nd. I.E.R.E.—"Systems using integrated circuits" by K. G. Nichols at 18.00 at 9 Bedford Sq., W.C.1.

2nd. B.K.S.T.S.—"A film-man's guide to lasers, holography, digits and bits" by C. B. B. Wood and E. W. Taylor at 19.30 at the I.T.A., 70 Brompton Rd., S.W.3.

3rd. R.T.S.—"Topics" by Eric Rout at 19.00 at the I.T.A., 70 Brompton Rd., S.W.3.

7th. I.E.E. Grads.—"Microwaves" by Thereza Macnamara at 18.30 at Savoy Place, W.C.2.

8th. I.E.E.—"Solid state devices for microwaves" by C. A. P. Foxell at 17.30 at Savoy Place, W.C.2.

9th. I.E.R.E.—"Earth's station installation projects" by W. B. Hopkins and N. Wheatley at 18.00 at 9 Bedford Sq., W.C.1.

10th. I.E.E./R. Signals Inst.—"Technology of mobile trunk communications in the 1970s" by Lt. Col. L. Ball at 17.30 at Savoy Place, W.C.2.

10th. I.E.R.E.—"Recent trends in oscilloscope technique" by B. Ellison at 18.00 at 9 Bedford Sq., W.C.1.

10th. R.T.S.—"Extracts of educational TV programmes" by Dr. P. D. Whitaker at 19.00 at the I.T.A., 70 Brompton Rd., S.W.3.

11th. I.E.E.—"Selectivity measurement for radio receivers" by H. Evans at 17.30 at Savoy Place, W.C.2.

11th. I.E.E.—"Standard cells as standards of e.m.f." by P. M. Clifford at 17.30 at Savoy Place, W.C.2.

15th. I.E.E./I.Mech.E.—"Financial analysis for engineers" by D. Murphy at 17.30 at Savoy Place, W.C.2.

16th. Inst. of Navigation—"Developments in airborne navigation equipment" by W. O. Broughton and J. W. McIvor at 17.00 at the Royal Inst. of Naval Architects, 10 Upper Belgrave St., S.W.1.

16th. I.E.E./I.P.P.S.—Colloquium on "Magnetic recording" at 17.30 at Savoy Place, W.C.2.

16th. I.E.E.—"Technological planning" by G. Siddall at 17.30 at Savoy Place, W.C.2.

16th. I.E.R.E.—"The design of amplitude-phase modulated by pulse sequences for radar" M. H. Ackroyd at 18.00 at 9 Bedford Square, W.C.1.

16th. B.K.S.T.S.—"Noise suppression in domestic sound reproduction" by Dr. Dolby at 19.30 at the World-Wide Films Theatre, 25 Breems Buildings, E.C.4.

30th. B.K.S.T.S.—"The B.K.S.T.S. colour test films" by Ray Knight at 19.30 at the I.T.A., 70 Brompton Rd., S.W.3.

ABERDEEN

10th. I.E.R.E.—"Microelectronics" by M. Newnham at 19.30 at Robert Gordon's Inst. of Technology, Schoolhill, AB9 1FR.

BATH

2nd. I.E.E.—"Recent developments in digital logic" by S. L. Hurst at 18.00 at the University, Claverton Down.

2nd. I.E.R.E.—"M.O.S.T. arrays—i.s.i." by M. Griffiths at 19.00 at the Technical College.

BOURNEMOUTH

16th. S.E.R.T.—"Solid state ignition" by D. Perkins at 19.30 at the College of Technology, Lansdowne.

CAMBRIDGE

2nd. I.E.E.—"Mintech and the electronics industry" I. Maddock at 19.30 at the College of Arts & Technology.

10th. I.E.E.—"Quartz crystals for electronic applications" by W. S. Metcalf at 18.30 at the Eng'g Labs, Trumpington St.

CARDIFF

9th. I.E.R.E./I.E.E.—"High-fidelity sound reproduction" by R. L. West at 18.30 at the University of Wales Inst. of Science & Technology.

16th. S.E.R.T.—"Modern radio control of models" by W. J. O'Neil at 19.30 at the Llandaff Technical College, Western Avenue.

CHELTENHAM

10th. I.E.R.E.—"Audio reproduction" by Dr. A. R. Bailey at 19.00 at the Government Communications Headquarters, Oakley.

COLCHESTER

9th. I.E.R.E.—"Printed circuit boards for micro-electronics" by J. Scarlett at 18.30 at the University of Essex.

COVENTRY

3rd. I.E.E.T.E.—"Storage of sight and sound" by J. E. Shepherd at 18.45 at the Technical College, The Butts.

CRAWLEY

1st. I.E.E.—"Developments in data communications" by M. B. Williams at 18.30 at the College of Further Education.

EDINBURGH

8th. I.E.E./I.E.R.E.—"M.O.S. integrated logic" by W. R. Betts at 18.00 at the Carlton Hotel, North Bridge.

EVESHAM

7th. I.E.E.—"Some aspects of TV reception from satellites" by K. G. Freeman at 19.30 at the B.B.C. Training Centre, Wood Norton.

FARNBOROUGH

8th. I.E.E.—"Microwave acoustics" by Prof. K. W. H. Stevens at 18.30 at the Technical College.

GLASGOW

7th. I.E.E./I.E.R.E.—"M.O.S. integrated logic" by W. R. Betts at 18.00 at the Instn. of Engrs. & Shipbuilders, 183 Bath Street, C.2.

HUDDERSFIELD

10th. I.E.E.T.E.—"Design and operation of electronic organs" by R. Paynter at 19.00 at the Lecture Theatre, College of Technology.

INVERNESS

16th. I.E.E.—"Telemetry" by R. E. Young at 19.30 at the Technical College.

LEICESTER

9th. I.E.R.E.—"Latest developments in oscilloscopes" by M. Davies at 18.30 at the Physics Lecture Theatre, the University.

LIVERPOOL

9th. I.E.R.E.—"Traffic control and the computer" by A. Davidson at 19.00 at the Dept. of Electrical Engineering, the University.

MAIDSTONE

7th. I.E.E.—"Invention as part of education" by Prof. M. W. Thring at 19.00 at the Royal Star Hotel, High Street.

MANCHESTER

10th. I.E.R.E.—"Electroluminescence" by C. N. W. Litting at 19.15 at U.M.I.S.T., Altrincham St.

10th. S.E.R.T.—"Medical Electronics" by L. T. Henry at 19.30 at Room J17, U.M.I.S.T., Sackville St.

NEWCASTLE-UPON-TYNE

7th. I.E.E.—"Pattern recognition" by J. R. Parks at 18.30 at the Polytechnic, Ellison Place.

9th. I.E.R.E.—"Electronics in meteorology" by C. V. Else at 18.00 at the Lecture Theatre, Ellison Bldg., the Polytechnic.

NOTTINGHAM

8th. I.E.E.—"Electromagnetic levitation" by H. Bolton at 18.30 at the TI Lecture Theatre, the University.

15th. I.E.E.—"Microcircuits, noise, sequential control and mechanised assembly" by A. Pugh at 18.30 at TI Bldg., the University.

READING

3rd. I.E.E./I.E.R.E.—"Automatic control of road vehicles" by S. Penoyre at 19.30 at the J. J. Thomson Laboratory, the University, Whiteknights Park.

RUGELEY

1st. I.E.R.E.—"Railways, computers, and signalling" by M. Birkin at 19.00 at the Shrewsbury Arms Hotel.

SOUTHAMPTON

1st. Brit. Computer Soc.—"Weather prediction by computer" by G. R. P. Benwell at 19.15 at the University, Mathematics Department.

2nd. I.E.R.E./I.E.E.—Colloquium on "Thick film circuits" at 18.30 at the Lanchester Theatre, the University.

SWANSEA

10th. I.E.E.—"Some applications of medical electronics" by D. W. Hill at 18.15 at the University College, Singleton Park.

SWINDON

1st. I.E.R.E./I.E.E.—"Numerical control of machine tools" by D. F. Walker at 18.15 at the College.

THURSO

17th. I.E.E.—"Telemetry" by R. E. Young at 19.30 at the Technical College.

Elements of Linear Microcircuits

3: Handling and safety precautions

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

Microcircuits (especially monolithic) are physically robust but electrically fragile. This article outlines precautions careful engineers commonly take in using them. The precautions apply equally to linear and digital circuits. Old hands may cut corners but beginners should observe all the precautions, initially at least.

Whatever kind of microcircuit is being used, and however experienced you are, always get hold of the device data sheet and study it till you understand it in detail, before you set to work.

Storing

If you do not use a microcircuit immediately you receive it, see that you store it carefully. This means avoiding extremes of temperature and humidity, and chemicals; or more simply, keeping it cool, dry and clean. If the microcircuits arrive in special protective package frames, keep these on until just before use.

When you do get out a microcircuit to fit into circuit, do not just leave it lying loose around the bench, where the pins or leads can get distorted. A useful trick is to keep a sheet of expanded polystyrene on which you can temporarily store the unit in safety by pushing its leads into the polystyrene.

Handling

As remarked earlier, microcircuits are physically robust. Even so do not handle them roughly. Generally units have leads designed to withstand pulls up to 1 lb for 0.020in diameter, 2 lb for 0.030in, 4 lb for 0.050in and 10 lb above this. Even so do not pull on the leads unnecessarily.

Most microcircuit leads have diameters between 0.015 and 0.030 inch and you should take care when you bend them. A good rule is not to bend the lead nearer than 1/32in to the body of the device. This is particularly important where the lead issues from the body through a glass-to-metal seal.

If the leads accidentally get twisted out of shape, you should dress them carefully back into shape with a pair of small, long-nosed half-round pliers. Where you are handling large numbers of units with displaced leads, you can buy a

lead-straightening jig to do the job safely without skilled hands.

Manufacturers aim for indelible code markings on their packages. However, if you are using a large number of different types, test the printing with your finger tip. If you find it liable to rub off, handle the unit so that the markings are not accidentally erased.

Modern microcircuits are so robust that you are unlikely to cause damage if you drop them, except that you may distort the leads. You can therefore handle them with the same care you would resistors, capacitors and other similar encapsulated components.

Mounting

How you mount a microcircuit in circuit depends very much on whether you are just using one experimentally, are making a few hand-assembled equipments, or are involved in large-scale production.

For preliminary bench experiments you can solder the leads to tag strips. But many microcircuits have such short and closely spaced leads that this is impracticable. One alternative is to solder the leads to terminal pins pushed into the selected holes in pre-punched 'Bakelite' boards such as the well known 'Lektrokit' and 'Veroboard'. Another alternative is the 'S-dec' solderless breadboard assembly.

More methodical workers mount their microcircuits with the leads threaded through the grid of holes in Lektrokit or Veroboard. Because the device leads are so closely spaced, the only suitable boards are those with a matrix of holes on a $0.1 \sqrt{0.1}$ grid. Two board thicknesses are commonly available, $\frac{1}{16}$ and $\frac{1}{8}$ in. You will find that the short, $\frac{1}{4}$ and $\frac{1}{2}$ in. leads found in some microcircuits may not allow enough lead to project through to make a satisfactory connection underneath the $\frac{1}{8}$ in board.

Three microcircuit package outlines are virtually standard. These are the dual-in-line, the multilead TO-5 and the flatpack as discussed in part one of this series in the October issue.

Only in the case of the dual-in-line are the leads (spaced at 0.1in on two parallel lines 0.3in apart) capable of being fitted directly through the holes on a 0.1in grid board. The dual-in-line thus mounts easily as in Fig. 1(a), or, if you wish, you can purchase dual-in-line mounting sockets.

When we come to the multilead TO-5, difficulties arise because the leadout positions do not match a 0.1in grid. This means re-dressing the leads as shown for 6-, 8-, 10-, and 12-lead versions in Fig. 1. Experimentally it is usual to thread $\frac{1}{4}$ in lengths of heat-resisting impregnated glass fibre sleeving over the leads to space the package off the board.[†] Some manufacturers supply their multilead TO-5 devices with leads already dressed to a 0.1in grid spacing.

If you do not want to bother with the fiddling adjustment of the TO 5 multileads, you can buy small plastic adapter pads with guide holes which steer the leads into the right configuration. Sockets for multilead TO-5 devices are also available for use with printed circuit boards, but most of these unfortunately feed the leads straight through so that

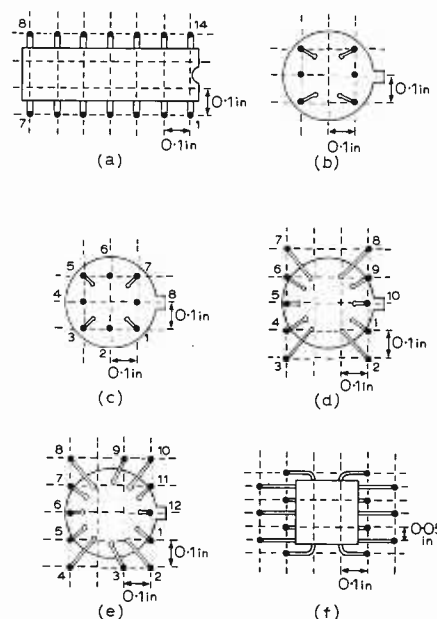


Fig. 1. How the leads of microcircuit packages can be dressed to conform to an 0.1×0.1 in matrix. (a) dual-in-line, (b) to (e) multilead TO-5, (f) flat-pack. Note that additional holes, spaced at 0.05 in. are required.

*Newmarket Transistors Ltd.

[†] See Literature Received this month.

there is still the difficulty of mating the socket pins to the board matrix.

The flat-pack is probably the most difficult package to handle experimentally. It is really designed to be mounted by welding or reflow-soldering the flat ribbon leads direct to metallization on one side of a printed-circuit board.

The difficulty of using it with the leads pushed through a 0.1in grid board is that the leads are on a 0.050in spacing and often only 0.25in long. No adapters exist for flat-pack use. One solution appears to be to bend the leads down at the spacings shown in Fig. 1(f) and drill holes in a blank board on a 0.050in grid to match the resultant pattern.

When a printed circuit board is being laid out to receive microcircuits, the dressed lead spacings indicated above can be replicated exactly. With the TO-5, however, the designer often abandons the 0.1in grid and merely symmetrically expands the pin circle diameter of the leads on the can (which is usually 0.2in) to a diameter of 0.4in on the board to give room for soldering pads round the holes.

Commercial microcircuits always have some index mark (e.g. the "pip" on a TO-5 can, or the coloured dot, or indent, on a flat-pack) to ensure that the device can be fitted in the right way round. The printing on the package is usually orientated in a fixed relation to the index mark, but sometimes you will find units with printing the wrong way round. Therefore, do not rely on the printing to orientate. Always look for the index mark, or you may have a 'dead' microcircuit in your hand.

Soldering

Microcircuit leads are generally designed to withstand temperatures of something like 325°C for ten seconds during soldering into circuit. Even so, always follow good soldering practice. Use a clean well-tinned iron, and heat the lead only long enough to ensure a good joint. Where the lead lengths permit, use one of the plastic pads available commercially for spacing the package off the board.

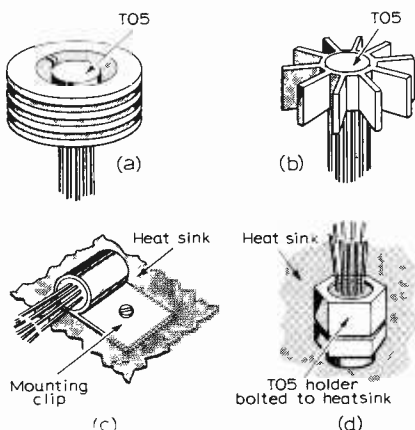


Fig. 2. Heatsinks for increasing the power dissipation of TO-5 microcircuits: (a) 'halo type', (b) 'corona type', (c) P-clip for attachment to a separate heatsink, (d) 'bolted-up' type for attachment to a separate heat sink.

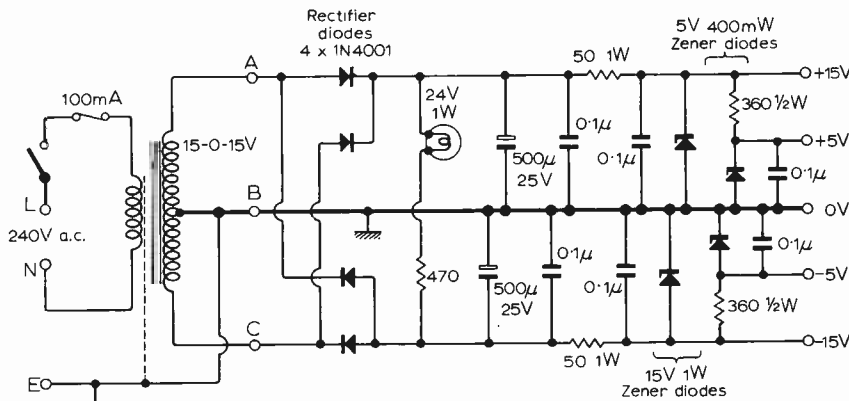


Fig. 3. Suitable power supply for most linear microcircuits.

Use a miniature soldering bit, and do not apply the iron on a lead closer than 1/16in to the body of the package. Many people grip the lead close to the body with a pair of fine-nosed pliers before soldering, to act as a 'heat shunt'.

The precautions detailed above of course would apply to soldering any very small component into circuit. But with microcircuits there is one special additional precaution necessary. The semiconductors in the microcircuit are prone to catastrophic failure if not protected from short high-voltage electrical "spikes". Remember that unearthed soldering irons are a fruitful source of these.

Some engineers find difficulty in unsoldering microcircuits, particularly those with many stout pin leads. Experienced bench hands use a large tinning iron (dressing the bit with a file into a suitable flat face), to melt the solder on all leads at the same time and swiftly withdraw the unit complete. In a production line special solder bits contoured to fit over all the leads at one time can be devised. Probably the most satisfactory way of unsoldering microcircuit leads is, however, to use one of the suction soldering irons now available. This neatly removes all the solder and permits clean easy withdrawal of the defective microcircuit.

Temperature

The soldering precautions discussed above are only one aspect of the thermal precautions to be taken with microcircuits. Generally, the cooler you keep a microcircuit, the longer it will live.

If there are any surfaces of the unit touching a larger mass (e.g. if the unit sits down on a printed circuit board) use silicone grease to mate the two touching faces for better heat removal. A variant of this is to use a thermally conductive epoxy such as Delta Cast 153 which has the additional advantage of holding the microcircuit firmly in position.

If you are mounting the microcircuit on a printed circuit board, leave as much copper on the board as possible to act as an extra radiator.

For supplementary heat removal use one of the 'add-on' heat sinks which are screwed or clipped onto the body of the microcircuit. Fig. 2 illustrates examples for

use with multilead TO-5 cases. Inside equipment, always try to locate your microcircuit as far as practicable away from hot spots round mains transformers, rectifiers, etc. Study your data sheet thermal ratings and operate your microcircuit as far below the ratings as possible. Engineers looking for long trouble-free equipment life, aim to keep dissipation 25 to 50% down on the permissible maximum.

Common microcircuits tend to fall into two categories: industrial with maximum junction temperatures inside the package of about 150°C and entertainment with a maximum junction temperature of about 100°C. To get some idea of the permissible dissipation of the different packages, note that typically a multilead TO-5 has a thermal resistance from junction to ambient of around 150°C/W and from junction to case of around 45°C/W. By comparison, a dual-in-line is about 100°C/W junction-to-ambient and a flat-pack about 185°C/W junction-to-printed-circuit board.

Power

Commercial single-polarity power supplies are often not ideal for working with microcircuits because many require a positive and a negative supply. Typical voltage requirements range from 15V to 5V, and currents up to about 150mA.

The relatively low currents and voltages required often lead experimenters to work with dry batteries. These are convenient and quite adequate, particularly if you use the grid-bias type with tapplings at every 1.5V. A word of caution, however, if there is a danger of the microcircuit drawing damaging excess current from the supply, do not rely on the apparently limited current rating of the grid-bias battery. It may surprise you to discover that it can provide well over an amp for a few seconds, long enough to blow up an overloaded microcircuit.

Fig. 3 gives a circuit for a double power supply suitable for use with linear microcircuits with two convenient zener-established voltages, ±15V and ±5V.

Because of the dangers from fast voltage spikes, power supplies used with linear microcircuits should always be decoupled to r.f. by 0.1 to 0.01 µF ceramic (low-inductance) capacitors, these being in

addition to any electrolytics used.

A danger with mains power supplies is that large smoothing capacitors can hold their voltages for a dangerously long time after switch off. Check that any supply used has some adequate 'bleeder' arrangement. In the version of Fig. 3, the pilot lamp circuit discharges the electrolytics rapidly on switch off.

With current-limited commercial power supplies, it might not seem important how the d.c. voltages vary on switch on and off, because the limited current available would not appear enough to damage the microcircuit. But such a supply may have sufficient output capacitance to provide a lethal spike for the microcircuit.

In work with microcircuits some engineers use an Avometer in series with the power supply and rely on the meter cut-out as a protection against damaging overloads. Apart from being bad for the Avometer, this puts inductance in the supply line which can easily lead to instability.

In working with mains power supplies on microcircuits, a cautious engineer will always turn off the supply before plugging in or unplugging the microcircuit.

Spikes

With microcircuits there are several steps you can take to guard against damaging overloads from external voltage transients. Firstly always operate well within the published ratings of the device. In the power supply, use a screened mains transformer with the screen adequately earthed.

Always guard against reversed polarity in the power supply. Where you are trying different circuit arrangements on the bench, fit some form of diode protection against accidental reversal. A series diode connected as shown in Fig. 4(a) is a common precaution.

Where you install zeners to clip dangerous spikes, use fast devices such as the emitter-base junction of a silicon planar transistor rather than the slower alloy-construction zeners.

Use low-leakage silicon diodes as limiters at sensitive points in the circuit. Fig. 4 gives three examples. In Fig. 4(b) oppositely phased diodes across the inputs of a high-gain operational amplifier restrict to a safe limit the differential voltage that can be applied. In Fig. 4(c) zener diodes are used to limit the common mode input voltage. In Fig. 4(d) a diode in the feedback loop prevents input 'latch-up', i.e. the output overdriving the input so that the negative feedback becomes positive, and the amplifier saturates.

Always use test prods with special care around a microcircuit.

Design

Anyone used to working with transistors knows the increased dangers of instability that arose when they went over from the older germanium low-frequency alloy

devices to typical planar silicon devices with frequency cut-off of hundreds of megahertz. Now in a microcircuit you may find crammed into a 0.05×0.05 in chip of silicon anything from five to fifteen such planar devices each capable of amplifying up to v.h.f.—an obvious invitation to r.f. instability and oscillation.

Additionally the circuit elements of the monolithic microcircuit are all connected through isolating reverse-biased diodes to the basic silicon substrate of the chip which has to be connected to the most negative point of the circuit supply. If instability arises or if supply polarities are accidentally reversed, these isolating diodes can become forward biased and allow damaging currents to flow.

Always look upon a microcircuit as a high-gain v.h.f. amplifier. This means that you use an adequate single-point earth in your surrounding circuitry. Concentrate on the neatest layout possible. Use short supply leads. Also see that your supply leads (and particularly your earth leads) are of low inductance and resistance.

In your coupling and decoupling arrangements, recognize that aluminium electrolytics have a relatively high inductance. Where electrolytics are necessary, use tantalum types (which are low inductance and nowadays only a little

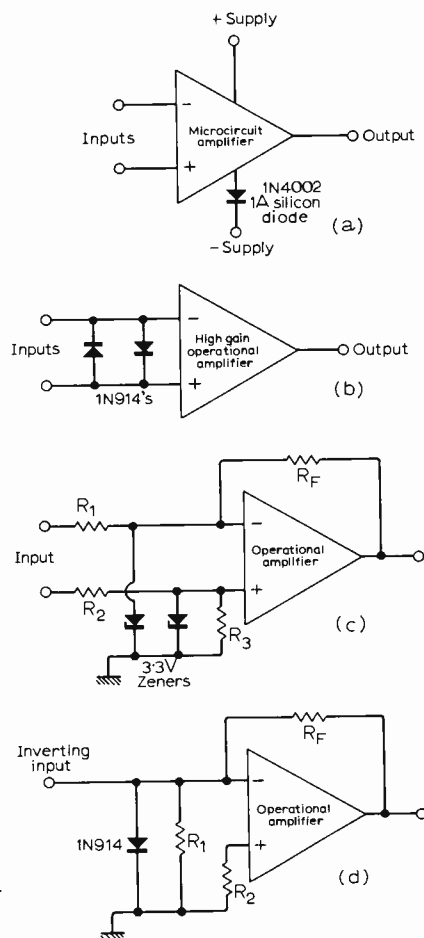


Fig. 4. How diodes can be used to protect op. amps from overloads. (a) preventing damage due to accidental power supply reversal, (b) limiting differential voltage inputs, (c) limiting common-mode voltage on the inputs, (d) protecting against latch-up.

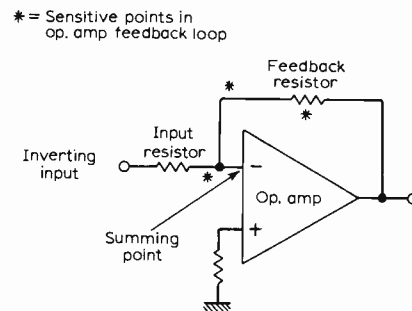


Fig. 5. Points in an op. amp. feedback network which are very sensitive to electrical interference.

more costly than aluminium). In general, connect a ceramic capacitor of about $0.1 \mu\text{F}$ in parallel with an electrolytic. By-pass r.f. to earth at both positive and negative supply rails. Look on every lead with jaundiced eye as an inductive choke and keep it as short as possible. Where you are operating more than one microcircuit from a supply, decouple each individually.

In shortening lead lengths, however, remember the dangers of uncontrolled feedback through interlead capacitances. In particular, always space your input lead as far as possible from the output and consider using shielded leads. With such high gain at v.h.f. you easily run the risk of r.f. interference, and you may on occasion have to go to the extreme of threading a ferrite bead on an input lead to avoid picking up your local taxi radio service!

As most general-purpose linear microcircuits are operational amplifiers, look carefully in your layout to the positioning of the summing point and the feedback resistor (indicated in Fig. 5) which are very sensitive to interference, and parasitics. If you want proof of this, just touch the feedback resistor and see what happens!

Fig. 6 shows a layout recommended for a standard $\mu\text{A} 709$ operational amplifier connected as a $\times 10$ voltage amplifier taking into account the major precautions listed above.

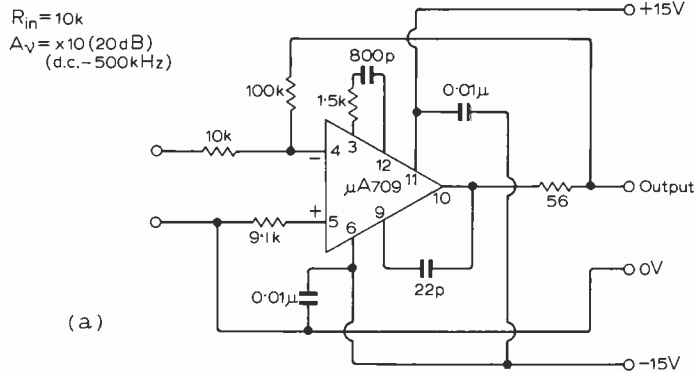
Instability

High-gain linear microcircuit used at d.c. or low frequencies are prone to r.f. oscillation from 5 to 50 MHz. This oscillation can be 'Nyquist' (caused by incorrect design values of the intentional feedback loop components) or 'parasitic' (arising from unintended feedback through interlead capacitances, inadequately decoupled supplies, etc.).

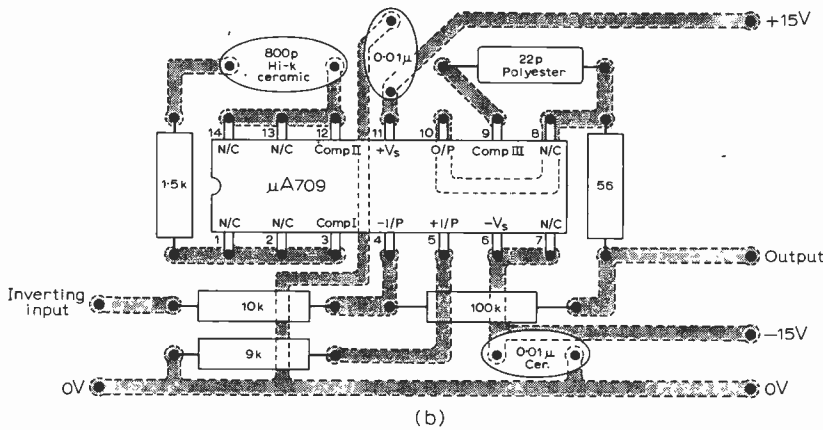
'Motorboating' (low-frequency oscillation) can also occur, but is usually easily cured by adequately large decoupling capacitors in the supply.

Curing Nyquist oscillations is mainly a matter of knowing basic feedback theory and studying the data sheet to see that the correct compensation network is used.

Parasitic oscillation is sometimes very difficult to eliminate because of the close



(a)



(b)

Fig. 6. (a) Circuit diagram of a 709 used as a $\times 10$ (20dB) amplifier, (b) recommended layout.

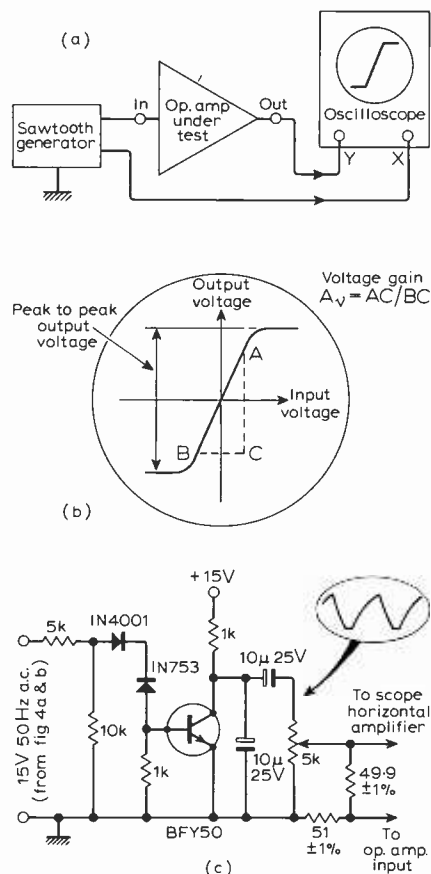


Fig. 7. Displaying the transfer characteristic of an op. amp. (a) general arrangement, (b) details of the oscilloscope display, (c) suggested circuit for the sawtooth generator part of (a).

proximity of the microcircuit leads. But good layout practice and adequate l.f.-cum-r.f. decoupling usually cures it.

Testing

There is no general-purpose instrument on the market for linear microcircuits. A single tester to deal with all the many different kinds of linear i.c.s would be impossibly expensive. So, if you need a linear tester, you have to lash-up a separate circuit for each type you use.

As most linear microcircuits work on the principle of the op.amp however, a useful general-purpose piece of test gear is something to display the transfer function. Fig. 7(a) shows an arrangement for such a display. Fig. 7(b) shows the sort of trace to be expected, while Fig. 7(c) gives details of a circuit for the sawtooth generator to provide drive voltages for the microcircuit input and the scope horizontal sweep input.

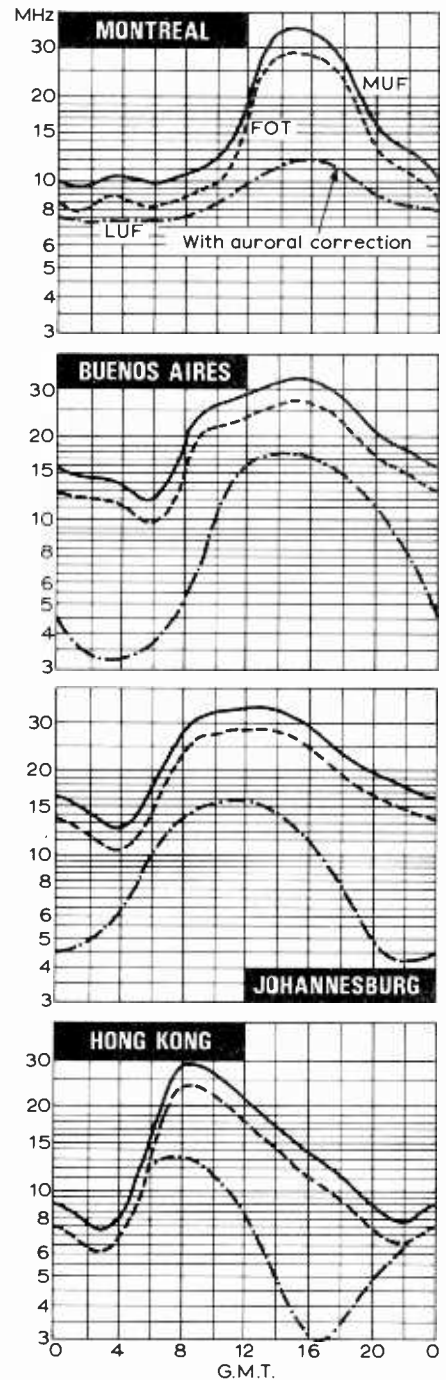
The oscilloscope finds another useful purpose in linear microcircuit work. Earlier we examined steps to avoid oscillation. The usual signs of oscillation are that the microcircuit is taking a heavy supply current or that the circuit operation is erratic. You can also detect oscillations on a scope, but as the oscillations can be occurring at up to 100 MHz, a low-frequency scope may only show you distortion or broadening of the l.f. trace. Because of the difficulties of getting a 'scope of sufficient bandwidth, use a small wideband r.f. detector or field-strength meter on the bench to detect oscillations.

(continued next month)

**H.F. Predictions—
DECEMBER**

Conditions during December have been almost identical for the last three years. The steady decline in solar activity predicted last December did not prove correct as the table of Solar Index values below shows. As a maximum was reached last May we can now expect a decrease at about half the rate of 1966/1967 increase.

	1966	1967	1968	1969	1970
Jan.	15	79	121	95	109
Feb.	21	95	109	104	114
Mar.	33	113	109	127	126
Apr.	37	114	109	122	130
May	46	115	113	118	144
June	55	92	111	119	125
July	55	89	96	114	122
Aug.	53	108	104	122	117
Sep.	42	124	121	115	
Oct.	47	125	133	110	
Nov.	64	106	106	106	
Dec.	66	115	112	108	



Personalities

Professor E. D. R. Shearman, B.Sc., A.C.G.I., F.I.E.E., of the Department of Electronic and Electrical Engineering, University of Birmingham, has been appointed a member of the Astronomy, Space and Radio Board of the Science Research Council. A graduate of Imperial College, London, Professor Shearman, who is 46, spent two years in the Admiralty Signals Establishment followed by 14 years at the Radio Research Station of the D.S.I.R., now the Science Research Council, before joining the staff at Birmingham University as a senior lecturer in 1962.

The appointment of **R. Gordon Williams**, M.A.(Eng.), M.I.E.E., as commercial director has been announced by Marconi Communication Systems Ltd. Mr. Williams, who is 49, took a degree in engineering at Wadham College, Oxford, and then went to the Army Radio School, being commissioned in the R.E.M.E. in 1943. He joined the Marconi Company in 1947 and three years later went into the sound studio equipment department of the Broadcasting Division. He was seconded to the Canadian Marconi Company for a nine-month television studio planning project. In 1957 he was appointed contracts manager in the Broadcasting Division of which he has been manager since May 1968. He is succeeded as manager of the



Gordon Williams



Cyril Teed

division by **Cyril Teed**, B.Sc., M.I.E.E., who has rejoined the company after spending two years as chief engineer of Independent Television News. Mr. Teed, who is 38 and a mathematics graduate of London University, spent two years as a radar lecturer in the Royal Air Force, and in 1952 joined the television camera design group of the Marconi Company. In 1959 he moved to California as the Company's television engineer in the U.S.A. Three years later he returned to the Division as deputy sales manager.

Brigadier E. J. H. Moppett, F.I.E.E., who retired recently from the board of Pye Telecommunications Ltd, of which he had been a member for 17 years, has been retained by Ultra Electronics Ltd as communications adviser. Brigadier Moppett served in Royal Signals throughout his military career and was head of the Government's Electrical Inspection Directorate immediately prior to his retirement from the Army in 1963.

E. A. Faulkner, M.A., Ph.D., M.I.E.E., F.I.E.E., senior lecturer in physics at Reading University, has been appointed to the Professorship of Solid State Electronics recently established in the Physics Department with the support of funds from Brookdeal Electronics Ltd. After National

Service, Dr. Faulkner entered University College, Oxford, in 1948. He obtained Second Class Honours in Physics and became a research student at the Clarendon Laboratory. In 1955 he joined the scientific staff of the G.E.C. Research Laboratories, and in 1957 left to take up a Research Fellowship in the Division of Tribophysics, in the Commonwealth Scientific and Industrial Research Organization, Melbourne. Dr. Faulkner was appointed lecturer in the Department of Physics, University of Reading, in 1960.

Peter C. Goldmark, B.Sc., Ph.D., president of CBS Laboratories, is the 1970 recipient of the Progress Medal of the Society of Motion Picture and Television Engineers awarded annually for outstanding technical contributions to the progress of engineering phases of the motion picture and or television industries. Dr. Goldmark, who was born in Hungary and received his doctorate in physics from the University of Vienna, started his professional career with Pye Radio in Cambridge before going to the United States to join the Columbia Broadcasting System as chief engineer in 1936. The S.M.P.T.E. has awarded the Progress Medal to Dr. Goldmark in recognition of past contributions to the television industry and the record industry, and especially for the development of the electronic video recording system, on which he wrote in our August issue.

Charles B. B. Wood, head of the image scanning section of the B.B.C.'s Research Department, is the only U.K. engineer among the list of 25 new fellows of the American Society of Motion Picture and Television Engineers.

Michael G. Dickson has rejoined E.M.I. after spending the past six years as manager, broadcast and communications equipment, with R.C.A. Ltd. Mr. Dickson has become manager, television and general electronics, with E.M.I. (South Africa) Ltd. Prior to joining R.C.A. he was with the broadcast equipment division of E.M.I. Electronics Ltd.

David Blythe, who is 33, is appointed general manager of Transiron Electronic Ltd. He has spent the past eight years with Texas Instruments, initially as a field sales engineer. He became managing director of T. I. Sweden, which he established, and came back to Britain to become manager of the general products department in T.I., Bedford.

Derek H. Pringle, Ph.D., F.Inst.P., research director of Nuclear Enterprises Ltd, has been

appointed a member of the National Radiological Protection Board. Dr. Pringle, who is 44, graduated in physics at Edinburgh University in 1948. In 1954 he obtained his doctorate in gas discharge physics. From 1948 to 1959 he was attached to the Valve Division of Ferranti, Edinburgh, becoming senior physicist. In 1959 he joined Nuclear Enterprises as chief physicist and was appointed research director and deputy managing director in 1961.

Changes in the management of its manufacturing organization have been announced by Pye Telecommunications Ltd. **N. Kendrick**, B.Sc., becomes works manager of the company's Cambridge Works in succession to **P. S. Westmacott**, B.A. (Cantab), who, as deputy manufacturing manager, has been assigned special responsibilities for central production planning and inventory control. **J. A. Farrell**, M.I.E.E., has been appointed works manager of the Telecoms Haverhill Works, in succession to Mr. Kendrick. He was previously with Royal Worcester Industrial Ceramics Ltd, and has also worked for Marconi Instruments and the General Electric Company.

Ian H. West is appointed assistant export marketing manager of Guest International Ltd with special responsibility for development of the company's Western European activities. Mr. West joined Guest from the Engineering Liaison Division of the British Radio Corporation's television plant at Gosport, following five years with C & N (Electrical) Limited, Gosport.

S. G. Grosmith is appointed sales manager of Kynmore Engineering Company Ltd, of London. Immediately prior to joining Kynmore he was sales manager of Seaelectro Ltd and previously with M.E.C. Ltd as assistant sales manager.

English Electric Valve Company has announced the appointment of **K. C. Ginn** as sales liaison engineer within the G.E.C. Group. He was with the Post Office Engineering Department before joining the Marconi Company, where he has been for the past 20 years, latterly as controller of marketing services in the Radio Communications Division.

Aidan Starkey, who joined Guest International Ltd 15 months ago, has been appointed manager of the Distributor Division. His experience in the electronics field includes 10 years on development work with Plessey followed by four years on R & D with the Company's Marine Systems Division.

Attenuators

Notes on the calculation and uses of resistance networks

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

The construction of highly accurate attenuators is quite a difficult matter, especially when they are for use at radio frequencies. If precise attenuation is required, resistor values must also be precise, and careful screening is needed. It is rarely worthwhile to attempt to construct an accurate r.f. attenuator; it is wiser to buy one! However, cases arise where no great precision is required and the attenuator is needed mainly to control the magnitude of an r.f. voltage in steps with only a rough sort of calibration of the magnitude of the steps.

Provided that the total attenuation is restricted to about 30 dB, little or nothing in the way of screening is needed and an ordinary wafer switch can be used. It is then often very worthwhile to "roll your own".

The basis of an attenuator is the section (which is connected between adjacent switch contacts). There are two types, the π and the T. Here only the π type will be treated, but any π section can be converted into a T section by applying the star-delta theorem to its elements. One π -type section is shown in Fig. 1. It is designed so that the resistance measured between the input (output) terminals is R_0 when the output (input) terminals are connected to a resistance R_0 . When this is done

$$R_0 = R_2 \parallel [R_1 + (R_2 \parallel R_0)] \quad (1)$$

where \parallel means "in parallel with"; i.e., $R_1 \parallel R_2 = R_1 R_2 / (R_1 + R_2)$.

The section is shown in Fig. 2 connected to a generator v_s of internal resistance R_0 and at the other end the attenuator section is terminated by R_0 . It is obvious that if the

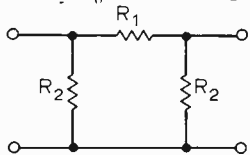


Fig. 1. Single section of attenuator.

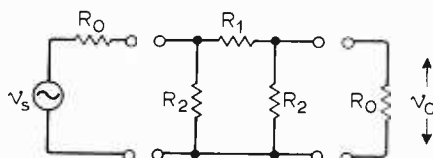


Fig. 2. Attenuator section with terminations R_0 .

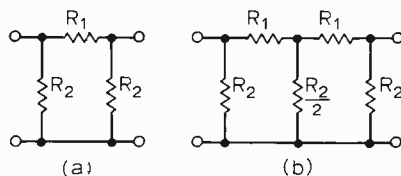


Fig. 3. A given attenuation can be obtained from one section (a) or from two in cascade (b). R_1 and R_2 do not have the same values in (a) and (b).

section were absent, so that the termination were connected directly to the signal source, there would be an impedance match between the signal source and the termination and v_0 would be $v_s/2$.

With the attenuator section in place there is still an impedance match at each end and so the voltage across the input of the section is $v_s/2$. The output voltage is then

$$v_0 = \frac{v_s}{2} \cdot \frac{R_2 \parallel R_0}{R_1 + R_2 \parallel R_0} = \frac{v_s}{2x} \quad (2)$$

where the insertion loss of the attenuator section is $20 \log_{10} x$ decibels.

From (1) and (2) a little algebra gives

$$R_1 = R_0 \frac{x^2 - 1}{2x} \quad \text{and} \quad R_2 = R_0 \frac{x + 1}{x - 1}$$

The proper values for R_1 and R_2 are thus readily calculated for any given resistance R_0 and any required attenuation $20 \log_{10} x$. Suppose $R_0 = 75 \Omega$ and 6-dB attenuation is wanted; then $x = 2$. We have

$$R_2 = 75 \times 3 = 225 \Omega$$

$$\text{and} \quad R_1 = 75 \times \frac{3}{4} = 56.25 \Omega$$

If 12 dB is needed, then $x = 4$ and

$$R_2 = 75 \times \frac{5}{3} = 125 \Omega$$

$$\text{and} \quad R_1 = 75 \times \frac{15}{8} = 140.625 \Omega$$

An attenuation of 12 dB can be obtained in two different ways as shown in Fig. 3. At (a) there is a single section with three resistors, at (b) there are two 6-dB sections in cascade with five resistors. In the middle the two resistors R_2 for the output of one section and the input of the other come in parallel and so can be replaced by a single resistor of $R_2/2 = 112.5 \Omega$. Of course, three sections of 4 dB each could be used, or four sections of 3 dB and so on. At high frequencies, it is often desirable to use several sections in cascade because stray capacitance has less effect.

If Thevenin's theorem is applied to a signal source v_s , R_0 , connected to an attenuator providing attenuation $20 \log_{10} x$, the open-circuit output voltage is v_s/x and the impedance measured at the output terminals with $v_s = 0$ is R_0 . Therefore, the equivalent circuit of the signal source and attenuator is as shown in Fig. 4.

In a signal generator a ladder attenuator is commonly used and one form is shown in Fig. 5. Each section is built to have a characteristic impedance R_0 using the preceding formulae and on the right the termination (shown dotted) is R_0 . At any switch contact the resistance is $R_0/2$ because the resistance of the attenuator is R_0 looking to the left and also R_0 looking to the right. With $R_0/2$ in series with the switch arm the total impedance viewed from the left is R_0 .

Looking into the right-hand terminals, however, the resistance varies with the switch position. At minimum attenuation and a source of impedance R_0 connected on the left, the resistance is $R_0 \parallel \frac{3}{2} R_0 = 0.6 R_0$. As the switch arm is moved to increase the attenuation the output resistance rises and tends towards R_0 as the attenuation increases.

The attenuator can, of course, be used the other way round, with the source connected

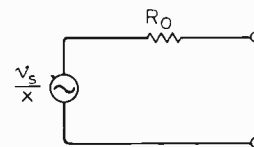


Fig. 4. A voltage source as in Fig. 2 connected to an attenuator has the equivalent circuit shown here.

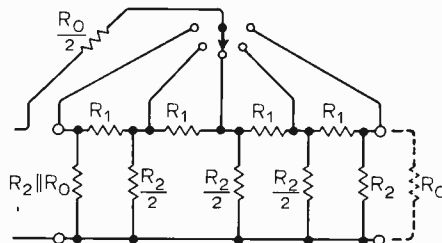


Fig. 5. A ladder attenuator with switch has a constant impedance R_0 viewed from the left, but an impedance which varies with the switch position when viewed from the right. This impedance lies between R_0 and $0.6 R_0$.

on the right and the output taken from the left. This is commonly done in a signal generator, for the attenuator then matches the output coaxial cable at all settings of the switch, and by using a voltmeter across the input and an additional input voltage control, the effect of the input resistance variations can be taken up on this control. When such a control is not fitted, there is always a mis-match at one end at low attenuation and the way round of connecting the attenuator depends on which end a mismatch has the less effect.

When an attenuator section such as that of Fig. 1 has values calculated for a desired attenuation $20 \log_{10} x$ with a termination R_0 , but is used instead with a termination Z_0 , which is not necessarily equal to R_0 , then the impedance at the other end is, from (1),

$$Z_0' = R_2 \parallel [R_1 + (R_2 \parallel Z_0)]$$

Inserting the values of R_1 and R_2 in terms of x gives, after a little algebra

$$Z_0' = R_0 \frac{Z_0(x^2 + 1) + R_0(x^2 - 1)}{Z_0(x^2 - 1) + R_0(x^2 + 1)} \quad (3)$$

Clearly, $Z_0' = R_0$ if $Z_0 = R_0$, and for any value of x . However, if $x^2 \gg 1$, $Z_0' \rightarrow R_0$ for any value of Z_0 . Therefore, if the extra attenuation can be tolerated, a mismatch can be overcome by introducing this extra attenuation.

The relation (3) can be used to determine the output impedance of the attenuator of Fig. 5 at settings of the switch other than full output. It was shown earlier that at minimum attenuation the output resistance is $0.6R$. With the switch connected to the previous contact the resistance looking to the left is still $0.6R_0$. Between this and full output there is the final section. If $x = 2$ (6 dB), $Z_0 = 0.6R_0$ and $Z_0' = 0.88R_0$. For a switch position one more step earlier $x = 4$ and $Z_0' = 0.975R_0$.

It is sometimes required to effect a change of impedance level with a resistance network instead of a transformer. This can be done at the expense of introducing attenuation. It is useful for matching, say, 75-Ω cable to a 50-Ω.

Such a matching pad is shown in Fig. 6. It has the property that the resistance looking into the left-hand terminals is R_{01} when the right-hand are terminated by R_{02} and conversely, where $R_{01} > R_{02}$. The basic equations are

$$\begin{aligned} R_{01} &= R_3 + R_4 \parallel R_{02} \\ R_{02} &= R_4 \parallel (R_3 + R_{01}) \end{aligned} \quad (4)$$

whence

$$\begin{aligned} R_3 &= \sqrt{[R_{01}(R_{01} - R_{02})]} \\ R_4 &= R_{02} \left[\frac{R_{01}}{R_{01} - R_{02}} \right] \end{aligned}$$

As an example, if

$$\begin{aligned} R_{01} &= 75 \Omega, \\ R_{02} &= 50 \Omega, \\ R_3 &= 25\sqrt{3} = 43.3 \Omega \\ \text{and} \\ R_4 &= 50\sqrt{3} = 86.6 \Omega \end{aligned}$$

The attenuation is not negligible. If the input and output voltages of the pad are

v_{in} and v_{out} respectively, the input power is v_{in}^2/R_{01} and the output power is v_{out}^2/R_{02} . Now

$$v_{out} = v_{in} \frac{\frac{R_4 R_{02}}{R_4 + R_{02}}}{R_3 + \frac{R_4 R_{02}}{R_4 + R_{02}}}$$

and a little algebra gives the attenuation as

$$10 \log_{10} \frac{R_{01}}{R_{02}} \left[1 + \sqrt{1 - \frac{R_{02}}{R_{01}}} \right]^2$$

If $R_{01}/R_{02} = \frac{3}{2}$, the attenuation is 5.7 dB.

With purely resistive networks impedance matching as such is not of any special importance. Nothing catastrophic happens if the input resistance of a part of a network does not match the output resistance of a preceding part. It is desirable, however, because it enormously simplifies calculations and because attenuators are commonly used in conjunction with coaxial cables, and with these impedance matching is important.

In testing a television i.f. amplifier it is sometimes necessary to feed a series-tuned circuit from a signal generator via a coaxial cable. It is required, for instance, to have the nearest equivalent to the arrangement of Fig. 7; this is a generator of 50 Ω internal resistance feeding the resonant circuit of the amplifier. This circuit might have $r = 10 \Omega$ and the reactance X varying from zero at mid-band to $\pm 150 \Omega$ at the band edges. The signal generator must normally be connected by a coaxial cable and it is obvious that the tuned circuit forms a very bad termination for it. A cable length of only some 4 ft forms a quarter-wave line at around 36 MHz with the result that the true circuit is very different indeed from the desired one. In practice, too, the signal generator will probably be of 75-Ω impedance and be used with a 75-Ω cable.

This is a case where the matching pad of Fig. 6 comes in, as was pointed out earlier, for matching 75 Ω to 50 Ω, R_3 is 43.3 Ω with R_4 86.6 Ω. It is now necessary to find out the input impedance Z_{in} of this section when it is terminated by $Z = r + jX = 10 + j150$ instead of by $R_{02} = 50 \Omega$. Equation (4) applies if we write Z_{in} for R_{01} and Z for R_{02} to get

$$\begin{aligned} Z_{in} &= R_3 + \frac{R_4 Z}{R_4 + Z} = R_3 + \frac{R_4(r + jX)}{R_4 + r + jX} \\ &= R_3 + \frac{R_4[r(R_4 + r) + X^2 + jXR_4]}{(R_4 + r)^2 + X^2} \end{aligned}$$

Inserting values we find that a mid-band where $X = 0$, $Z_{in} = 52.3 \Omega$ and at the band-edges, where

$$Z = 10 \pm j150, \quad Z_{in} = 90.2 \pm j35.5 \Omega$$

The match to a 75-Ω cable is still not good,

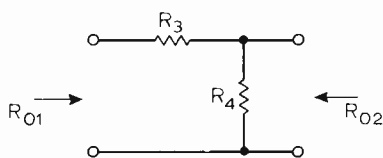


Fig. 6. Matching pad for joining two different impedance levels.

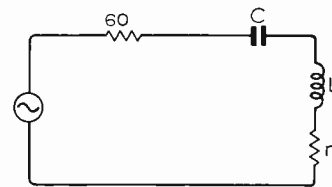


Fig. 7. Tuned circuit connected to a signal generator.

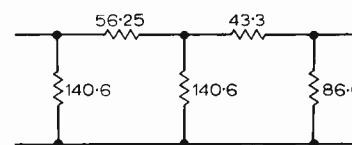


Fig. 8. Pad for matching 75 Ω to 50 Ω (right) with 6 dB attenuator at 75 Ω (left).

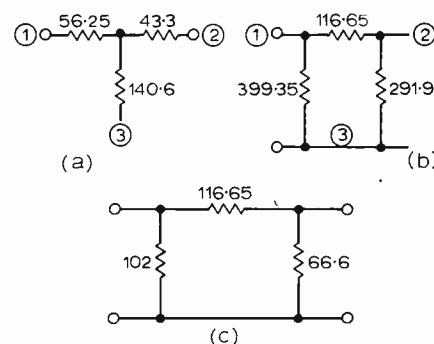


Fig. 9. Development of an equivalent network (c) for Fig. 8.

but it is very much better than the tuned circuit alone would provide. However, let us now insert a 6-dB attenuator immediately in front of the matching pad. Equation (3) applies if we replace Z_0 by $Z_{in} = 52.3 \Omega$ (mid-band) = $91.2 \pm j35.5$ (band-edge). With $X = 2$, the equation then becomes

$$Z_0' = 75 \frac{225 + 5Z_{in}}{375 + 3Z_{in}}$$

At mid-band

$$Z_0' = 75 \frac{225 + 261.5}{375 + 156.9} = 68.5 \Omega$$

At the band edges

$$\begin{aligned} Z_0' &= 75 \frac{225 + 456 \pm j177.5}{375 + 273.6 + j106.5} \\ &= 79.6 \pm j7.4 \Omega \end{aligned}$$

The resistive component now varies from 68.5 Ω to 79.6 Ω and forms a reasonable match to 75 Ω while the reactive component at its worst is under 10% of the resistance and is almost certainly negligible. Fig. 8 shows the arrangement of a 6-dB attenuator section and the matching pad. The five resistors used can be reduced to three! To do this, apply the star-delta theorem to the star of 56.25 Ω, 43.3 Ω and 140.6 Ω as in Fig. 9(a) to get the delta equivalent (b). The input and output shunt arms of Fig. 8 can then be combined with the similar arms of Fig. 9(b) to get the final form (c) which is a combined attenuator and matching pad exactly equivalent to Fig. 8. In practice, Fig. 8 is usually more convenient.

Simple Class A Amplifier and Modular Pre-amp

A postscript to designs published last year

by J. L. Linsley Hood

The author has had the benefit of an extensive and frequently helpful correspondence with readers following the publication of the two circuit designs.* Attention has been drawn to some obscurities in the original articles and to certain possible improvements in the design. Details are given below.

Class-A amplifier

Power supply. Although much interest was aroused among constructors by the good technical (and audible) performance given by the amplifier, it is clear that the principal feature in the eyes of many users was the relative simplicity of the circuit. This being so it must have seemed a pity that the power-supply unit was somewhat complex. However, the purpose of this power-supply design was to avoid possible degradation of the amplifier performance by h.t. ripple. The sawtooth ripple voltage across the reservoir capacitor in a class-A amplifier drawing some 2-3 A will be many times greater than in a class-B system, particularly at the audibly important low-signal levels.

However, a number of measurements have been made since the publication of the original article on an amplifier of this type, operated from a simple power supply unit of the type shown in Fig. 1. There is little difference in the performance above 100Hz either in total harmonic distortion or in intermodulation distortion, although the shape of the output power/distortion curve at the onset of overload is modified, as would be expected, by the ripple on the h.t. line. Below 100 Hz the distortion curve rises more steeply to about 0.2% at 20 Hz.

*"Simple Class-A Amplifier", *Wireless World*, April 1969.
 "Modular Pre-amplifier Design", *Wireless World*, July 1969.

A thermistor is necessary, in this case, to slow down the rate of rise of the h.t. voltage. This will get hot in use.

It now appears that the mains transformers used in the development of the prototype of this amplifier were not as efficient in respect of apparent secondary circuit resistance or secondary leakage reactance as some of those which have been supplied for this purpose since the publication of the article. The reservoir voltage found with the 15Ω system may be above that given by the author. At switch-on this can cause a transient overloading of the transistors specified for the series regulator circuit in the original article. In view of this, it is suggested that these should be an MJ481 or 2N3055, used in conjunction with a 2N1613 or, better still, a 2N699. These amendments are shown in Fig. 2.

Adjustment of amplifier output current and centre-line potential.

The author had supposed, somewhat naively, that most constructors of the circuit would have somewhere in their workshops a collection of odd-value resistors needed for trimming circuit parameters, and it was mentioned in the original article that the desired quiescent levels could be set by adjustment to R_2 and R_5 or R_6 . This sort of comment is unhelpful if one is writing away for a kit of parts. In view of this it is suggested that R_2 should be replaced by a resistor in series with a potentiometer, as shown in Fig. 2. The necessary value of resistors R_5 and R_6 to give an entirely adequate accuracy in mid-point voltage setting can be predicted, and the suggested amended values are shown.

Some obscurity arose, inadvertently, in the original diagram concerning the reason for the different values of input decoupling resistor quoted for mono and stereo use. This was because it was intended that the one decoupling circuit should serve both

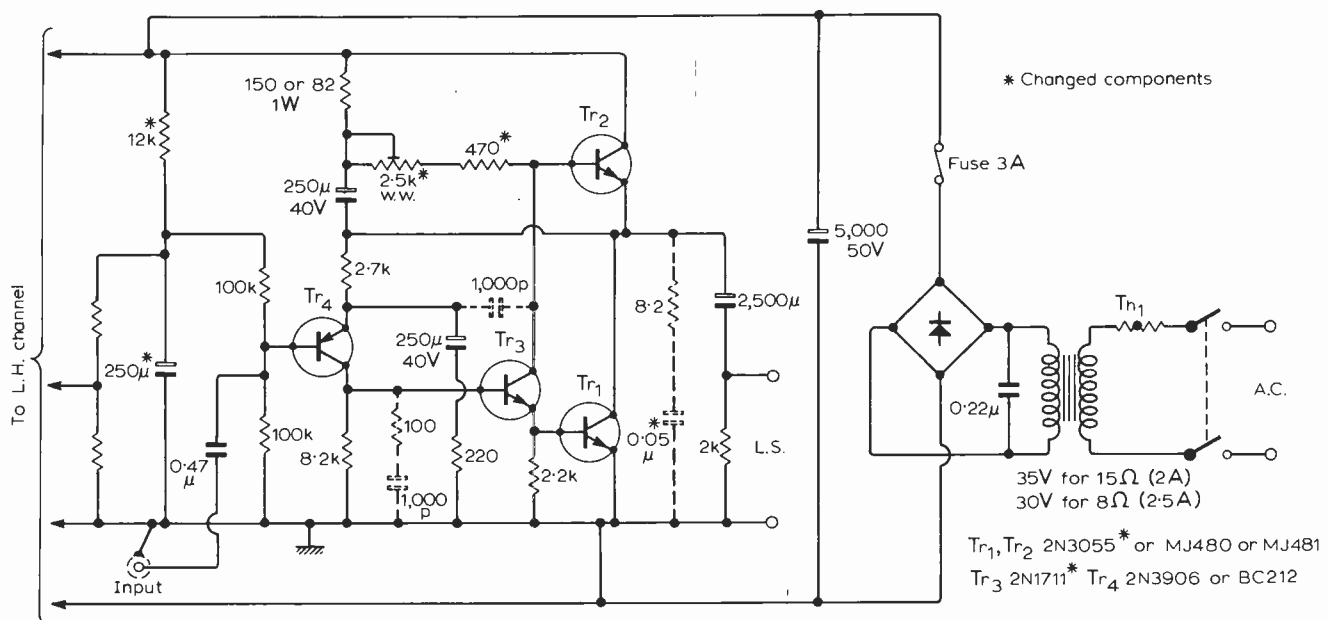


Fig. 1. Suggested amended circuit for 8 or 15Ω, use employing a simplified power supply. The dotted components reduce the h.f. response and should be used with capacitive loads.

channels. Where an unsmoothed h.t. supply is used it is recommended that the decoupling capacitor should be increased in value to $250\mu\text{F}$.

Stability of output current setting. Some criticism has been voiced because there is no specific control over the output current value in the simplest form of this circuit, other than that due to the stability of the current gain of Tr_2 , whose performance determines this parameter. In order to meet this point (in anticipation) a circuit was described in the original article which allowed precise control over the operating 'quiescent' current without detriment to the performance of the amplifier.

However, measurements made on an amplifier without this addition have shown no significant change in operating current in somewhat over two years use, and there is also little measurable difference in current from a minute or so after switch-on to the end of a six-hour period of continuous use. In practice therefore, in temperate climates at least, the simplest form of the circuit is adequate in this respect. If any user cares to experiment with an alternative and somewhat more elegant form of quiescent-current control another regulation circuit is shown in Fig. 3. The transistor used as Tr_5 requires to be somewhat more massive than that used for Tr_3 since the mean collector current is twice that of Tr_3 and the maximum voltage and current occur simultaneously. The 2N2905A is just about adequate with a good heatsink, but a larger power device such as the 2N4919 is preferable.

Alternative transistor types. The amplifier has been built successfully with a wide variety of transistors, including fully complementary versions to operate from an existing negative h.t. line, and in one case two identical amplifiers have been made for use with the inputs in paraphase, in order to double the available output voltage swing. One constructor has, indeed, made a stereo 30W system using two such pairs of amplifiers plus input phase splitter, as shown in Fig. 4.

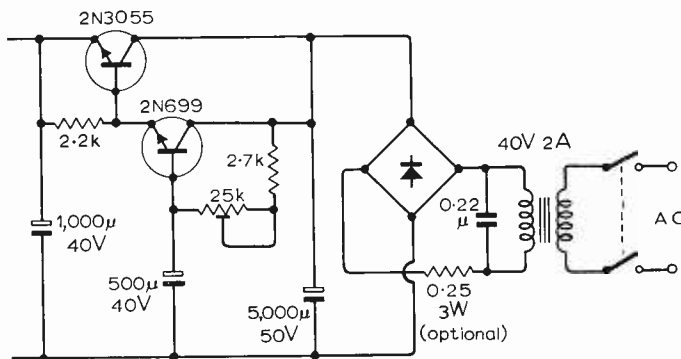


Fig. 2. Amended circuit of power supply for 15Ω systems.

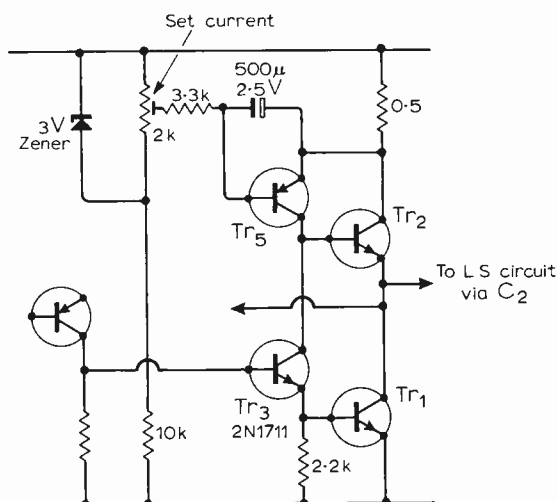


Fig. 3. Alternative method of quiescent-current control. R_1 , R_2 and C_1 in the original have been deleted. Tr_5 is 2N4919 on heatsink or alternative type.

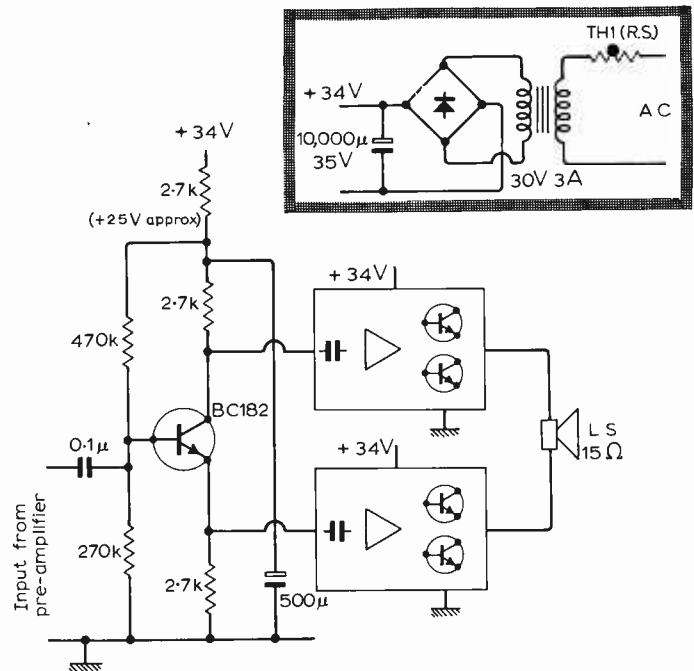


Fig. 4. Scheme for 30W class-A system. Two identical class-A amplifiers, each designed to give 15W into 8Ω (34V at 1.5A), are driven by a phase-splitter. The output capacitors have been removed.

However, one transistor change which is recommended is the use of a 2N1711 as Tr_3 . This has a high voltage capability equal to that of the 2N1613, and a current gain which is double that of either the 2N1613 or the 2N697. The use of the 2N1711 instead of the former types suggested for Tr_3 increases the feedback factor and approximately halves the typical distortion factor of the system (0.025% at 9W or 0.05% at full power) without detriment in other respects.

Also, a 2N1711 as Tr_3 allows the use of 2N3055 devices as Tr_1 and Tr_2 , with a final performance which is equal to that of the original specification below 100kHz. (The typical current gain of the 2N3055s is only half that normally found with the MJ480/1 output transistors, and their use was not originally recommended for this reason.)

Gain/frequency and power/frequency characteristics. These are, in fact, better than the curves published in April 1969. As mentioned in a letter to the editor published in October 1969, the h.f. fall-off shown was mainly due to an error in the measurement instrument. Although the performance at h.f. depends to some extent on the layout employed, the small signal voltage gain, with the component arrangement shown, is flat (within 1dB) to beyond 2MHz. This may be a snag in some cases because even a small feedback capacitance between output and input (as may happen, for example, if the output heatsinks are not earthed) may cause the amplifier to oscillate. A suitable circuit change to reduce the amplifier h.f. response to more normal levels was described in the letter above. This is not an essential modification—the author's own units are still exactly as described in April 1969.

The output power response of the unmodified amplifier is flat within 1dB to 200kHz.

The l.f. response shown in the original gain/frequency and power/frequency graphs was that determined for an earlier prototype of the amplifier. During the development of the circuit the values of some of the capacitors were increased to improve the l.f. performance, and by an oversight the graphs accompanying the article were not amended. In fact the gain and power graphs can be shown as 'flat' from 10Hz–200kHz. In this respect, and that of transient response, the class-A design is probably better than any circuit so far published. The i.m. distortion, at 10W output, (70Hz and 7kHz, 4:1) is less than 0.1%.

Miscellaneous. Surprise—and even alarm—has been caused to some constructors by the fact that the output transistors get hot.

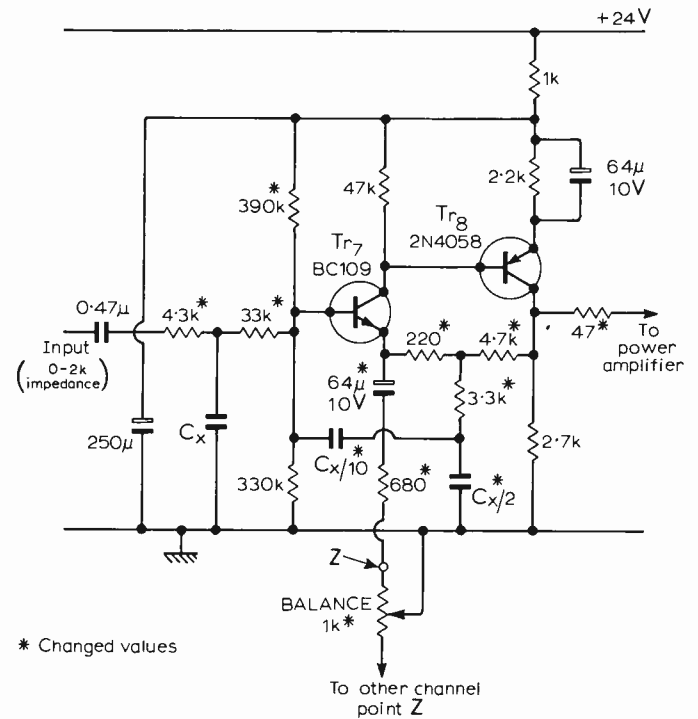
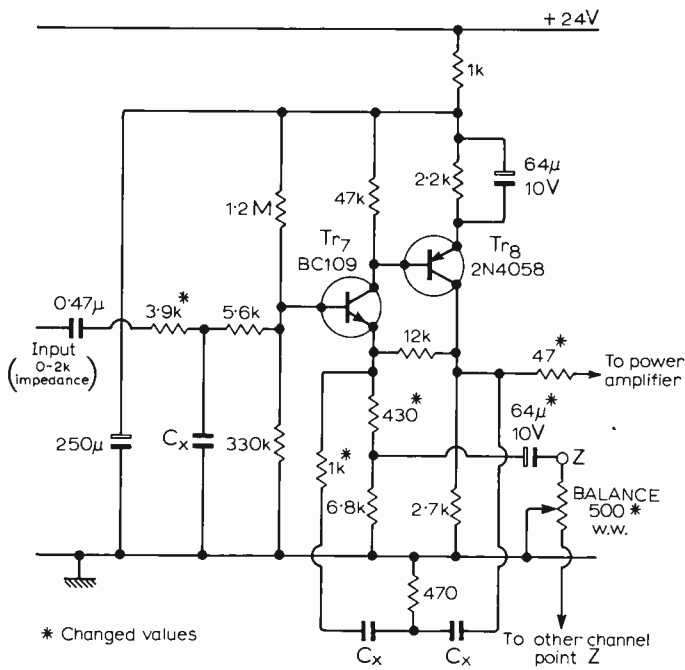


Fig. 5. Low-pass filter and driver stage (a) with gain of 20, (b) with a gain of 5. The cut-off frequencies, dependent on C_x , are the same as in the original article.

However, with adequate heatsinks, which should be black painted, the dissipations in the transistors are only a small fraction of the maker's permitted level, and provided that some care is taken in the layout to make sure that sensitive components, such as electrolytic capacitors, remain cool, no reduction in the working life of such a system, in comparison with an equivalent class-B unit for example, is to be expected.

Some difficulty has apparently been encountered by some constructors because the power supply regulation system is inoperative when the supply is operated without a load. If an equivalent resistive dummy load is connected for bench-testing, all should be found to be well.

Finally, it is prudent to wire a small resistor of about 2kΩ across the loudspeaker terminals to make sure that the output capacitor charges even with the speaker disconnected. Charging of the capacitor by an accidental short-circuit could cause damage. This addition is shown in Fig. 1. No damage is caused by operating the amplifier on an o/c output.

Modular pre-amplifier

The intention in the original article was not to offer a complete pre-amplifier design, but rather to describe a series of versatile 'building blocks' from which the potential user could assemble a 'custom built' pre-amplifier to suit his own needs or preferences. To increase the scope of this some additional circuit modules are described below.

Steep cut low-pass filter. It is certainly prudent to include a low-pass filter somewhere fairly close to the input of the main amplifier whenever a wide-bandwidth main amplifier is to be used with a good-quality loudspeaker system. Doing so will prevent unwanted high-frequency components, arising from component noise, record surface noise, and similar causes, from impairing the long-term listening comfort of the user, and from producing avoidable intermodulation effects due to non-linearities in the loudspeakers.

The combination of such a steep-cut low-pass filter with a low-distortion, low-output impedance driver stage, with a gain of 50 and an output capability of some 2V r.m.s. at 0.02% t.h.d., appeared to provide the most versatile system for use with a wide variety of power amplifiers.

However, many power amplifiers require an input voltage of only 0.25–0.8V r.m.s., and there are snags in respect of hum and component noise if the stages following the volume control are operated at levels below some 50mV. The preferred level to

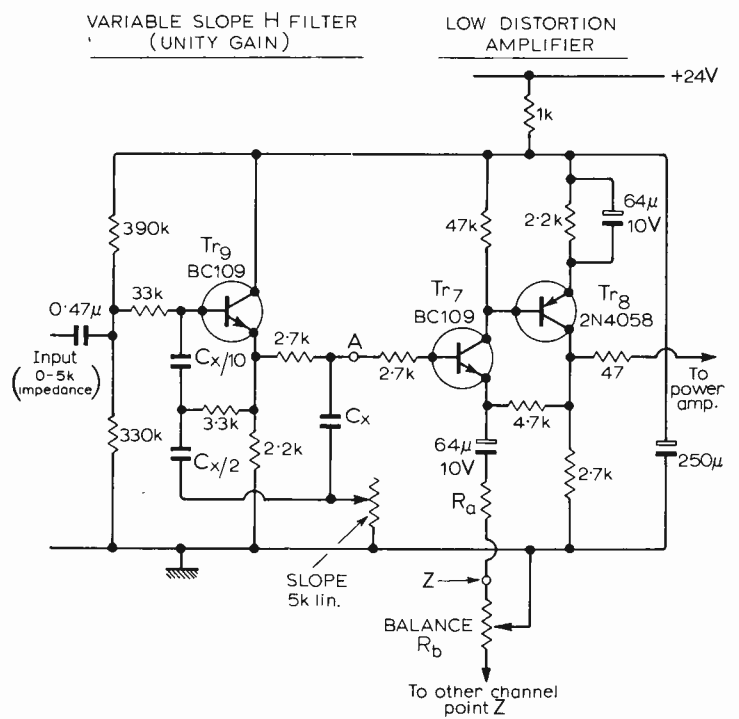


Fig. 6. Pre-amplifier driver stage incorporating a switched low-pass filter with slope variable from 6-18dB/octave. The gain of the filter is variable through the range 5-100 by choice of R_a and R_b .

$$Gain = 1 + \frac{4.7}{R_a + R_b/2}$$

Output can be taken from point A if only unity gain is required.

achieve an optimum balance of noise and distortion components is probably in the 100–200mV region. In these circumstances a driver-stage gain of 50 is excessive, and much of the available gain must be removed by an input attenuator, and if a potentiometer is used for this it can introduce noise.

To meet this need more conveniently, two further versions of the driver amplifier, incorporating steep-cut low-pass filter characteristics which are identical to that of the original circuit, and having gains of 20 and 5, are shown in Figs. 5(a) and 5(b). An

alternative, three-transistor arrangement whose cut-off slope is variable over the range -6 to -18 dB octave, at any chosen (switchable) frequency, is shown in Fig. 6. This consists of a single transistor version of the 'H' filter used in the two previous pre-amplifier designs (the nomenclature derives from the shape of the component layout in the 'op-amp' form), followed by a very low-distortion two-transistor amplifier whose gain can be chosen, as required, over the range 5 to 100, by adjustment of R_a and R_b . If a unity-gain stage is all that is required (actually the gain is about 0.9) the output can be taken from the point marked 'A' on the diagram, and Tr_7 and Tr_8 omitted.

The response curve of the filter circuit, at any chosen turnover frequency is shown in Fig. 7. The slope is smoothly variable by adjustment to the $5k\Omega$ pot. If the slope pot. is open circuit the response is flat to 20kHz and beyond, but in this case the load impedance should not be less than $50k\Omega$.

For completeness, an equivalent single-transistor high-pass filter, having a cut-off slope approaching 18dB/octave, and suitable for use as a 'rumble' filter or a pre-amplifier woofer/tweeter cross-over filter, is shown in Fig. 8. The frequency response characteristics of this filter are shown in Fig. 9. Both of these filter circuits should be driven from a source having a fairly low impedance—not higher than $6k\Omega$.

If single transistor 'H' filters are to be used at output signal levels exceeding 100mV a Darlington transistor, e.g. Motorola MP5A14, is to be preferred.

The apparent noise level, referred to the input, of the two-transistor driver amplifiers, using reasonably low noise transistors and an input impedance of the order provided in the normal circuit, is about $4-6\mu V$. The output noise voltage in the original circuit was $0.2-0.3mV$, which should be inoffensive. With a lower gain driver stage this noise will be reduced even further.

The use of a variable negative feedback type of balance control in these circuits is deliberate, in that it permits a low output impedance to be obtained from the driver stage. Measurements made with a wide range of published transistor-operated power amplifiers have shown that substantially lower distortion levels are often given by using a low-impedance drive circuit, and that there is frequently an advantage also in terms of hum, noise, and transient response.

Tone-control circuit. This stage has a worst case (bass and treble controls set to maximum 'lift') distortion level which is typically less than 0.1% at 1V r.m.s. output. It is perfectly capable of

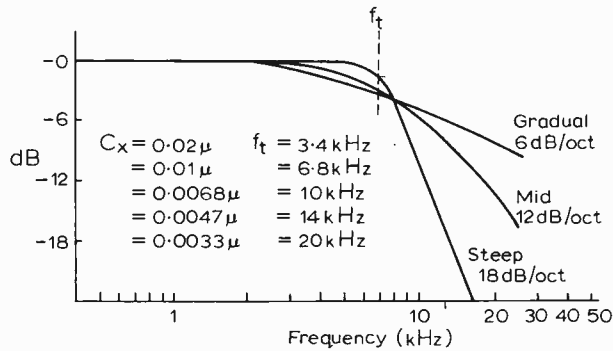


Fig. 7. Slope characteristics and turn-over frequencies of variable-slope 'H' filter.

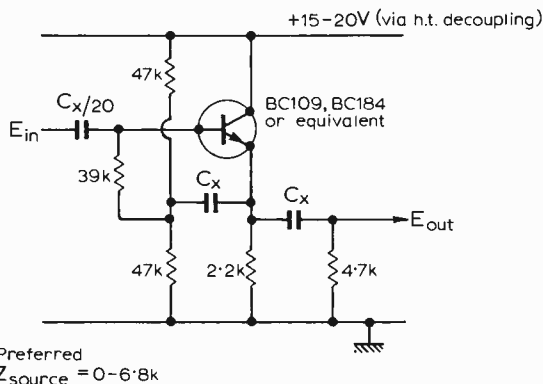


Fig. 8. Single transistor high-pass 'H' filter.

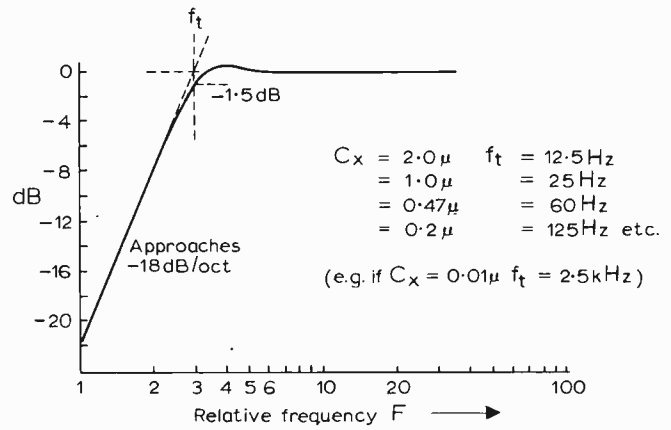


Fig. 9. Transmission characteristic of high-pass 'H' filter.

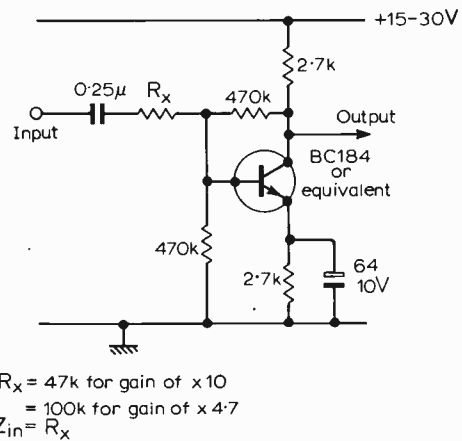


Fig. 10. Simple low-distortion single transistor amplifier.

driving a normal high-quality power amplifier without the interposition of other pre-amplifier stages. The required signal amplification could then be provided prior to the volume control. This is tending to be the normal practice in commercial 'hi-fi' amplifiers, in that it gives the highly-sought-after zero noise-level at minimum volume control settings, and makes for economies in the use of components.

Noise in the tone-control stage due to the f.e.t. has caused occasional troubles. This should not occur with the f.e.t. now recommended for this part of the circuit (the Amelco 2N4302), which appears to have a consistently low noise level. The necessary bias-adjustments were described in a letter to the editor published in April 1970.

The input impedance level suggested for the tone-control stage was $50k\Omega$, because it was thought that most of the other systems likely to be used with this unit would be transistor operated; and this would be of a suitable level for this purpose, while avoiding some of the hum pick-up problems likely to be encountered at higher impedance levels. However, if this impedance is too low, and if a high gain (beta greater than 400) transistor is selected for Tr_4 —in fact most BC109s will do—the base bias resistors can be increased to $1M\Omega$ and $560k\Omega$ (instead of $200k\Omega$ and $100k\Omega$) enabling the volume control and auxiliary control potentiometers to be increased to $25k\Omega$.

If an even higher input impedance is required, the f.e.t. impedance conversion shown in Fig. 4 in the original pre-amp. article can be substituted in its entirety for Tr_4 . To preserve the function of the rumble filter in this circuit, with the $0.47\mu F$ capacitor desired to feed the tone control network, a $4.7k\Omega$ resistor should be connected from the output side of this capacitor to the earth line. A low-noise f.e.t. is of course preferable.

If additional amplification is required on any signal source prior to the tone-control stage (if this is working at the 100mV level) a simple single-transistor feedback amplifier such as that shown in Fig. 10, can be used with confidence, in that its performance is stable, its noise level is low, it is almost impossible to damage by

an input overload, and its distortion is well below 0.1% at output voltages up to 0.25V r.m.s., and with gains up to 10.

Magnetic pickup equalization circuit. Some requests have been received for component values for the use of this circuit for tape-replay characteristic equalization. The author remains of the opinion that this type of provision is best left to the manufacturers of the tape recorder, in that the actual head characteristics can influence the replay frequency/voltage characteristics.

However, a fairly close approximation to the replay curve theoretically required for 7.5 i.p.s. is given if C_2 and R_2 in the original equalization network A are altered to 100pF and 27k Ω .

The noise level of this circuit is almost entirely determined by the performance of Tr_1 . The BC184C and 2N5089 transistor types may be of interest in this position.

The maximum output which can be obtained from this circuit at 0.02% t.h.d., is 2V r.m.s. If the normal input to the tone control circuit, or other following stages, is 100mV, this gives a 26dB overload capability. The gain of the equalization circuit can be increased by a factor of 3, (i.e. to 30 at 1kHz) without upsetting the rumble filter characteristics if R_3 is reduced to 68 Ω and C_4 increased to 100 Ω F.

Miscellaneous. An omission from the original article was the suggestion that high value resistors (2-5M Ω) should be

connected across the switch contacts, from slider to each C_x . This removes 'plops' on switching ranges.

A number of correspondents have queried the need for a separate h.t. power supply for the pre-amp. (The reservoir capacitors for the unit shown should have read 35V working, not 25V). It is always possible to run the pre-amp via a suitable voltage-dropper circuit from the main amplifier power supply, and if a zener diode is included in this line, this scheme may be satisfactory. However, measurements on channel separation, and harmonic and i.m. distortion, with identical amplifier systems invariably show some advantage, particularly at the low-frequency end of the audible spectrum, in the use of a separate power supply for the pre-amp (even when the electrolytic bypass capacitors are still new) and this arrangement is still recommended by the author as well worth the small additional cost.

One point which has not been published, to the best of the author's knowledge, concerns the particular advantage conferred by the feedback pair amplifier using complementary transistors, such as that used in the low-pass filter circuit, in comparison with the more usual n-p-n/n-p-n pair, where the bias for the first transistor is derived from the h.t. line. In the case of the n-p-n/p-n-p pair, any h.t. line feedback, due to inadequate h.t. line bypass, will be negative rather than positive, and this can assist in obtaining good t.h.d. figures down to low signal frequencies.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Class AB amplifiers again

Mr. Linsley Hood's reply in the October issue to my letter (August) does indeed clear up the difficulties I experienced in following his article and his reply to Mr. Gibbs (August issue), but I feel bound to justify my objections more fully. I understand the mutual conductance of a transistor or a pair of transistors to be $\frac{dI_c}{dV_{be}}$; V_{be} is

measured between the input base and output emitter, under precisely those near zero source impedance conditions to which he refers. With values of less than an ohm the shape of the basic mutual characteristic of the MJ481 is preserved. The curve obtained with 100- Ω source resistance looks much more like the current gain characteristic, except at low collector currents. If the effect of the 10- Ω resistor is removed from Mr. Linsley Hood's curve A, the slope does become steeper than that

of curve C. Consider an MJ481 with and without a 0.2- Ω emitter resistor and with and without a 40361 driver in the Darlington pair configuration, with zero source impedance (Fig. 1), with modifications where appropriate. It is easier to work in terms of mutual resistances than conductances, and representative values of these are shown in Table 1 (R is infinite here).

The mutual resistance of combinations of these three, including the MJ481, is the sum of these resistances seen at the output

emitter. The MJ481 is assumed to have a current gain of 100; this does not prejudice the argument as the characteristic of the 40361 is nearly exponential, so that the slope is approximately inversely proportional to I_c . The results for the four cases are shown in Table 2. The optimum quiescent current for a voltage driven stage is normally the collector current at which the resistance slope is twice its high current value.

It can be seen that the addition of an

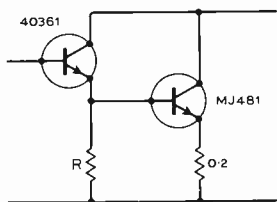


Fig. 1.

Table 1

$\frac{dV_{be}}{dI_c}$ of MJ481	I_c of MJ481	I_c of 40361	$\frac{dV_{be}}{dI_c}$ of 40361
			100
Ω	mA	mA	Ω
0.16	1,000	10	0.04
0.25	200	2	0.15
0.32	100	1	0.30
0.50		0.5	0.50

Table 2

Combination of components	Slope at 1 A output current	Optimum quiescent current	Slope at this current
	Ω	mA	Ω
MJ481	0.16	100	0.32
MJ481 + 0.2 Ω	0.36	50	0.70
MJ481 + 40361	0.20	200	0.40
MJ481 + 40361 + 0.2 Ω	0.40	100	0.82

emitter resistor reduces the optimum quiescent current and of a driver increases it, although either addition reduces the overall mutual conductance at all currents. The effect of finite values of R is to reduce the change introduced by the driver.

The p-n-p/n-p-n configuration is more complicated (c.f. Mr. Baxandall's letter in the September 1969 issue), but in general it has a higher mutual conductance (Fig. 2, $r=0$) than the simple output transistor. With common values of r the combination is linear down to much lower

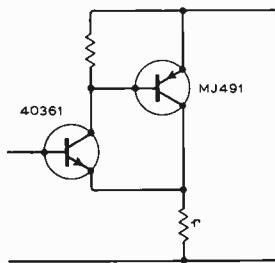


Fig. 2

collector currents in the output transistor, giving a lower half-slope current. With a high source impedance the optimum quiescent current for a complementary or quasi-complementary output stage is not so readily defined. It may well be Mr. Linsley Hood's experiences in these circumstances which leads him to the conclusion (August issue) that the optimum quiescent current varies inversely with the absolute magnitude of the current gain in half of the output stage.

The "circumstances which would exist in a practical design" are precisely those put there by the designer; source impedances of under 1Ω are perfectly feasible. It begs the question to insert resistors in the base lead before even measuring the basic properties of the transistors. The mutual characteristic so obtained is only relevant to a complete amplifier which has these impedances in series with each half of the output stage—resistors R_1 & R_2 in Figs. 3 & 4—excepting pure class B using transistors which cut off perfectly and do so with zero base-emitter voltage. If R_1 & R_2 are

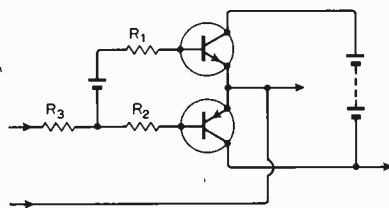


Fig. 3

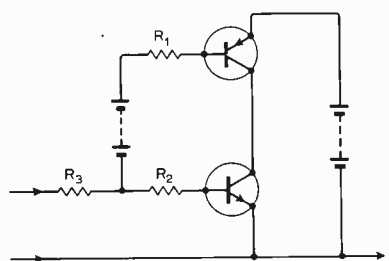


Fig. 4

zero, and R_3 is finite, the overall transfer characteristic of the complete output stage is best not looked at in terms of the mutual conductance measured when one transistor is omitted.

I apologize for making objections in terms of the article, since it does not convey the sense that the author intended, but I based my arguments on the design itself. The source impedance to the output stage is genuinely low. The minimum current gain of an MJ521 at 50 mA collector current is about 80, giving a drive impedance of 70Ω at the most (derived from the $6.8\text{ k}\Omega$ resistor). The input impedance of the output stage varies between 50 and 100 ohms in the 15Ω version with output transistors of current gain 100. It is the inappropriate ratio between these two quantities which is responsible for the effects to which I referred.

It would be convenient if the bootstrap capacitor could supply the extra current required to drive low gain MJ491s which need a base current in excess of the standing current in the driver stage. This could only occur if the bootstrap capacitor temporarily sustained a greater voltage than it does under static conditions. This situation arises during a short negative transient (MJ491 on) a short time after a long positive excursion (MJ481 on). Short and long are referred to the time constant of the bootstrap capacitor and R_4 in Fig.3 of the article. Quite how common these conditions are in music (with whatever d.c. components there might have been removed well before bootstrap capacitor has its say) I can't imagine.

The other points I should like to make are best left to a future date—we both appear to be drawing on material which should see the light of day in articles rather than in letters.

DUNCAN MITCHELL,
Postgraduate School of Electrical and
Electronic Engineering,
University of Bradford.

In defence of the amateur

I was glad to see the present situation regarding some of the v.h.f. amateur bands put so lucidly by Dr. Saxton in "Letters", October issue. It is obvious that for short-term gains, many commercial interests as well as national ones around the world would like to insert the thin end of the wedge into amateur frequency allocations. Clearly there is an increasing ignorance in commercial circles of the unique long-term value of amateur radio. When industry itself profits so much by the (admittedly immeasurable) boost to its own manpower by the early influences of amateur radio, then the movement can still ask for and I think expect support from this quarter. The

influence of such a movement on young people not only develops the interest of future technical employees but also broadens the experience of non-technical staff into the hardware field, to say nothing of the enrichment of social life for every one. I feel that industry and commerce that have an interface with amateur radio should offer technical assistance, market small quantities of materials even if inconvenient, and give general encouragement for the maintenance and extension where possible of the wavebands allocated to the amateur service.

The matter of early influences particularly interest me and when assisting in the work of setting up the R.S.G.B. Education Committee, I voiced the opinion that an Hon. Youth Officer, or some such worker, should be established. It would be excellent if, during his presidential year, Dr. Saxton could bring a little influence to bear for inaugurating a youth programme, as I know he is very interested in the educational aspects of amateur radio.

It is to be deplored that in the same issue we read in paragraph two of the report on the R.S.G.B. Exhibition that *Wireless World* itself is suggesting that the amateur frequency bands be re-allocated to a more deserving cause. This is a terrible generalization from insufficient data. (one stand in one exhibition!). Come now Mr. Editor, ask your reporter to call in at any club constructional evening, R.A.E. lecture, or R.A.E.N. practice session etc. etc. KEN SMITH, G3JIX, Canterbury, Kent.

"Toothpaste bass"

Your readers all over the world think of *Wireless World* as being above the controversy*, as being a legitimate source of unbiased information on the world of electronics, and music reproduction in particular. Therefore, it pains me to hear (in your report on Sonex 70) our IMF loudspeakers characterized as having "toothpaste bass"—a highly emotional characterization of the only loudspeakers in England which have any bass, to my knowledge!

In the cause of scientific truth, and also be it known, of British exports, I do believe your reporter should be called to order a bit. The role of criticism is not to destroy, or to create personal vendettas, but to inform the public and to instruct the public and the industry on current techniques for furthering the art—and not some individual's pique!

IRVING M. FRIED,
IMF Products,
Philadelphia, Pa.

*The publication of this letter shows that we are not; we do allow both sides to have their say. ED.

Active Filters

16. Some uses of amplifiers with differential input

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

A method of deriving a family of two-amplifier circuits from the standard three-amplifier two-integrator loop is described and their properties discussed. Only some members of the family appear to be of practical use, but these offer high Q factor at reduced cost. The discussion is also of interest for the light it throws on the relationship between bootstrapping and the use of amplifiers with differential input.

In Part 15 it was shown how a two-amplifier circuit, using amplifiers with differential input, can be derived from the standard three-amplifier two-integrator loop (the basic high- Q active CR circuit). The two-amplifier circuit, Fig. 1(c), may be thought of as a bootstrapped version of the standard circuit; and the essential steps in the derivation are (1) the moving of the

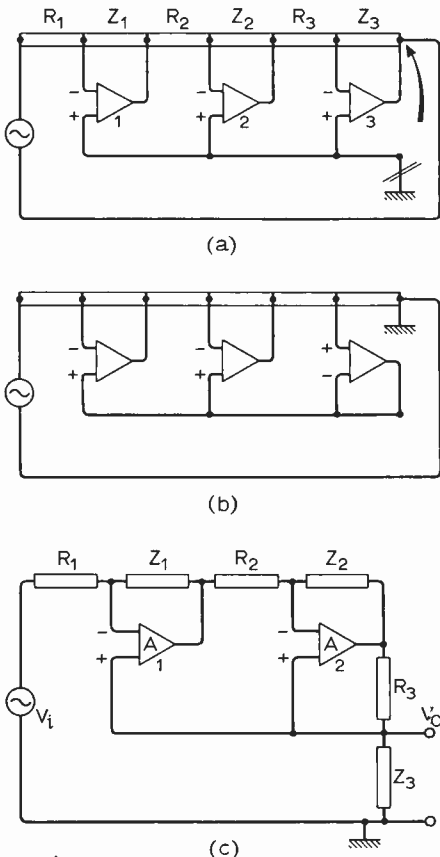


Fig. 1. Summary of derivation given in Part 15.

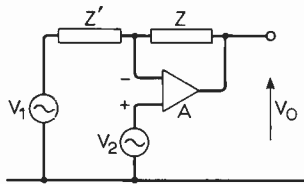


Fig. 2. Application of input voltages and feedback to an amplifier with differential input.

earth point as shown in Fig. 1(a) and the consequent conversion of amplifier 3 to a voltage follower, Fig. 1(b), (2) the removal of the voltage follower and the direct connection of the + terminals of amplifiers 1 and 2 to the junction of R_3 and Z_3 .

A constant feature of the standard two-integrator loop, and of the circuits derived from it as described, is the string of six passive elements which in Fig. 1(c) are drawn conventionally as rectangular boxes; and it is helpful, both to emphasize this feature and for economy in space and effort, to draw the string as a bar, Figs 1(a) and (b). It is to be understood that two of Z_1, Z_2, Z_3 (any two) are capacitances and the remaining one a resistance, giving three distinct circuits (see Fig. 3, diagram 1). The two having Z_3 a capacitance can, by the removal of that capacitance, serve as earthed inductances: the third with Z_1, Z_2 capacitances is useful as a low-pass filter section, since in this case R_3 and Z_3 are both resistances, and therefore form an aperiodic potential divider, and a low-impedance low-pass output can be taken from the output terminal of amplifier 2. But three circuits from one is only a beginning.

Commutating the amplifier input terminals

In a feedback arrangement as shown in Fig. 2, since the amplifier needs only a very small (ideally infinitesimal) differential input voltage to give any output voltage within the linear range, the negative (inverting) input terminal of the amplifier virtually follows the positive input terminal, i.e. provided the amplifier is not saturated the negative input terminal of the amplifier carries a voltage very nearly equal to V_2 . It seems reasonable to suppose, therefore, that

the performance of the two-amplifier system, Fig. 1(c), will be essentially unchanged if the positive input terminal of amplifier 2 is connected to the negative input terminal of amplifier 1. The same conclusion can also be reached by reference to the original three-amplifier loop: the correct voltage to pass on to the second stage is virtually the voltage developed across Z_1 .† Thus diagram 2 of Fig. 3 is obtained. For the time being no notice should be taken of the numbers

†R. C. Bowes derived this circuit independently as an equivalent of the standard two-integrator loop. (Private communication.)

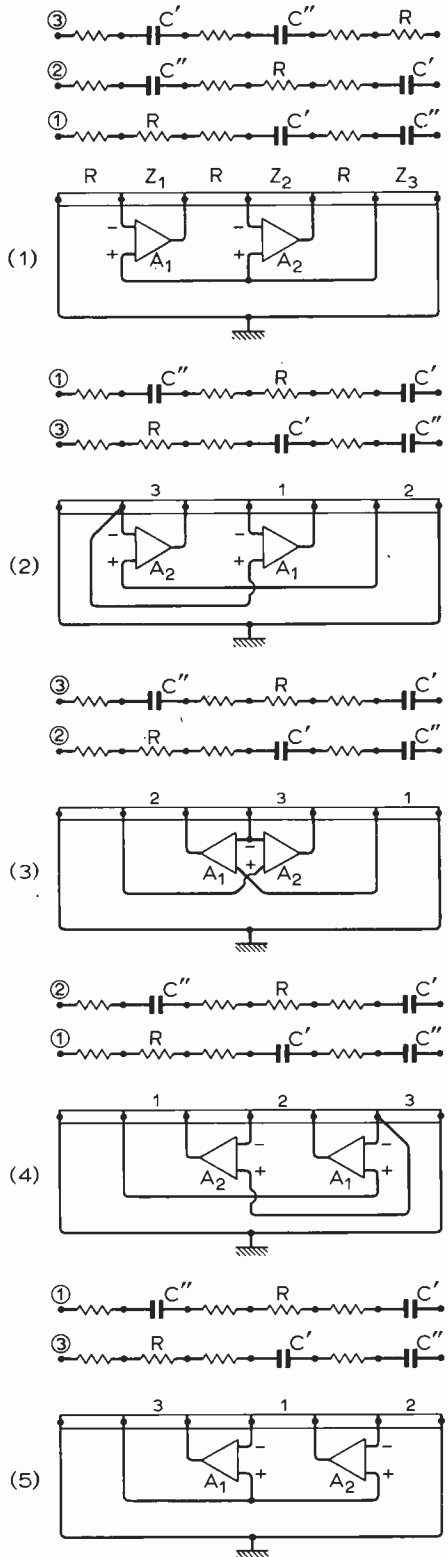


Fig. 3. Five of the twelve connection diagrams.

*Royal Radar Establishment.

written beside the amplifiers and element strings; they are for cross-reference with Fig. 5.

No input or output terminals are shown in Fig. 3. A key to their normal positions is given in Fig. 4, where the analogy is made with an equivalent passive circuit. With ideal amplifiers, the voltages V_0' and V_0'' in the active circuit (b) will be equal, and will correspond to the voltage V_0 in the passive circuit (a). Thus l-p response is obtained at both V_0' and V_0'' with input at V_1 (V_2 shorted), and h-p response with input at V_2 (V_1 shorted). In practice the voltages V_0' and V_0'' are not themselves very suitable as outputs, because they come from high-impedance points. But, as described in Part 15, by arranging the element string so that the Z adjacent to the desired output point is a resistance, a similar output may be taken from the output terminal of one of the amplifiers.

Fig. 4(c) shows the two arrangements of the passive elements which give an inductance, and the two corresponding complete element strings (six elements) are shown against diagrams 2 to 5, Fig. 3, and in Fig. 7; but the third string may of course be used, as in diagram 1, for the synthesis of a quadratic factor.

Connection diagram 2 of Fig. 3, with the lower of the two strings shown (3), corresponds to the circuit given as a simulated inductance in Fig. 6(c) of Part 15, and which analysis and experiment show to be satisfactory for high- Q circuits, though, when finite gain in the amplifiers is allowed for, of somewhat lower intrinsic Q than those connected as diagram 1.

Another circuit already given, Fig. 6(b) of Part 15, which was brought to the present authors' attention by L. T. Bruton, is seen to be also a member of the family, and to

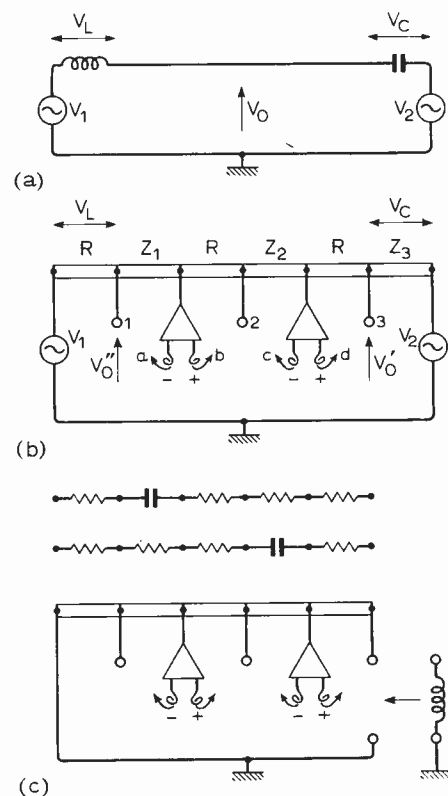
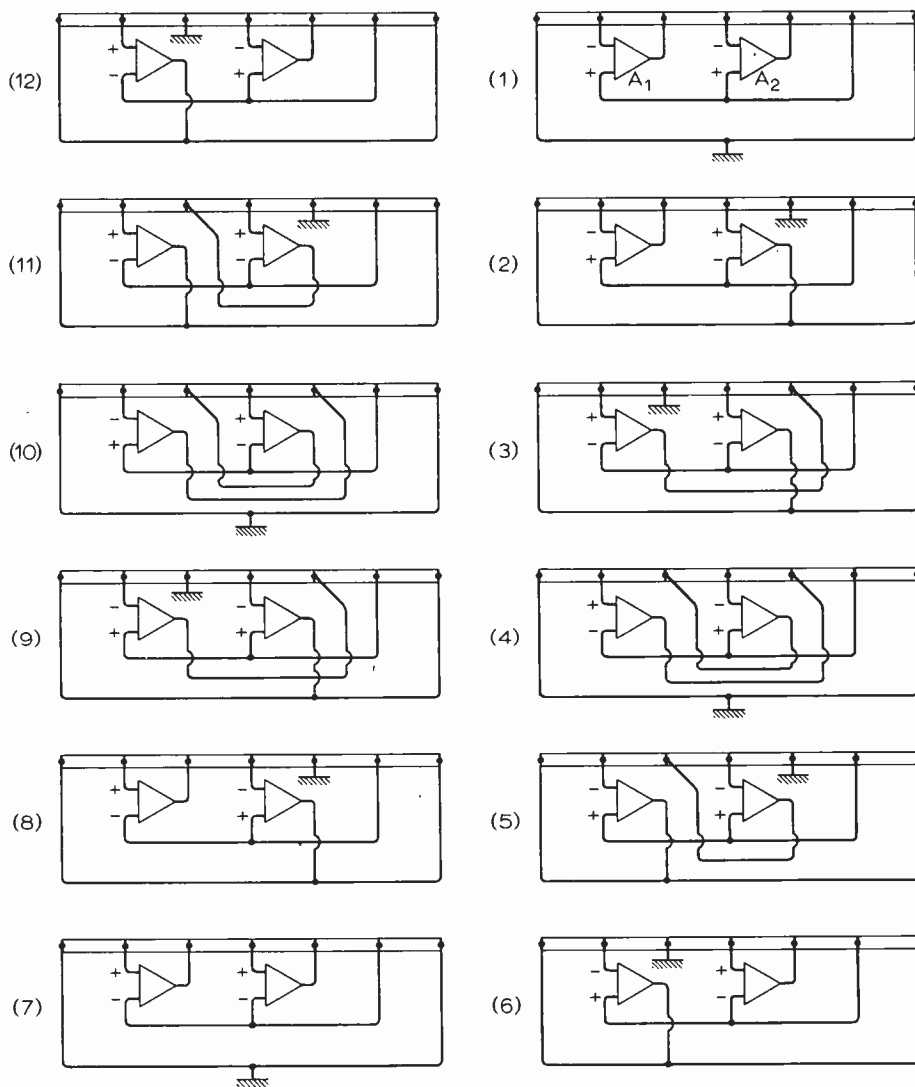


Fig. 4. Some essential features.

Fig. 5. The twelve connection diagrams as derived by moving the earth point.



conform to connection diagram 3 of Fig. 3, used with the upper string (3).

The change from diagram 1 to diagram 2, and from diagram 2 to diagram 3, are each a change in connection to one amplifier input terminal, and it is interesting to enquire if the progression can be continued and other diagrams obtained.

The argument already given, that ideally (i.e. when $A \rightarrow \infty$) the negative input terminal of each amplifier follows (i.e. has the same potential as) the positive terminal, shows that ideally in the circuits so far derived all four amplifier input terminals are at the same potential; and it may be supposed that provided certain rules are observed there could be a large number of ways in which leads a, b, c, d, and terminals 1, 2, 3 may be interconnected, Fig. 4(b). Connection diagrams 1, 2, 3 are specified in the first three entries of Table 1, and it can be seen that the progression consists in the moving on to the next network terminal that amplifier input lead which has just been joined by another. So diagram 4 is obtained from 3 by moving lead c from terminal 2 to terminal 3, diagram 5 from 4 by moving lead

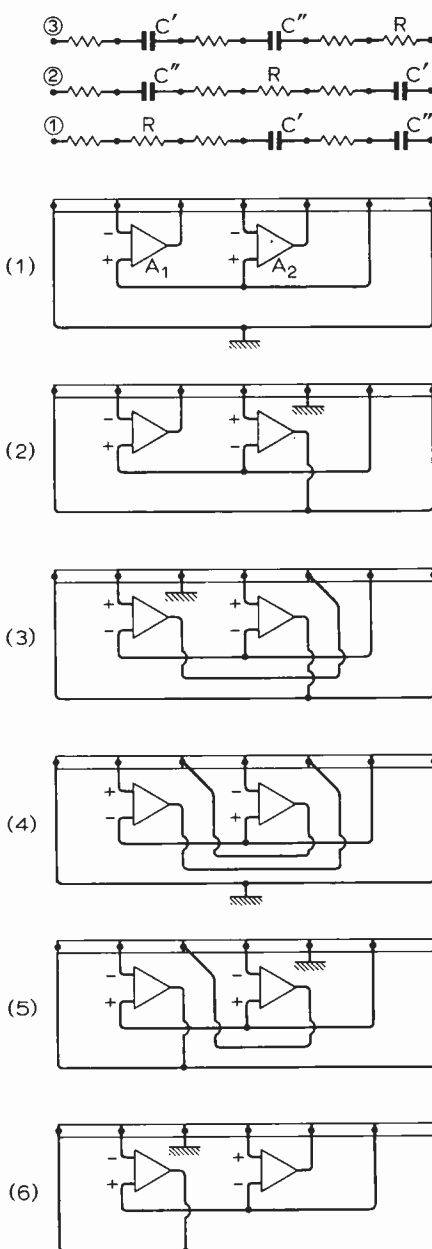


TABLE 1

Permutation of the connections between the three terminals of the passive-element string, 1, 2, 3, and the four input terminals of the amplifiers, a, b, c, d, giving twelve connection diagrams and hence thirty-six circuits.

	1	2	3		1	2	3		1	2	3
1	a	c	bd	5	bd	a	c	9	c	bd	a
2	da	c	b	6	b	da	c	10	c	b	da
3	d	ac	b	7	b	d	ac	11	ac	b	d
4	d	a	cb	8	cb	d	a	12	a	cb	d

b from terminal 3 to terminal 1, and so on till diagram 1 is again reached, making a total of twelve diagrams.

Although plausible, Table 1 is actually a back-construction, and the twelve diagrams were first obtained by further logical movements of the earth point. In Fig. 1(b), for example, the terminals of the chain fall into two interlaced sets. The first set are low-impedance points, being connected to earth through the low (ideally zero) output im-

pedance of amplifiers, through a voltage source, or directly. The second set are joined only to high, ideally infinite, impedances, the input terminals of amplifiers. This is an aspect of the fundamental nature of these circuits. Current may, indeed must, enter or leave the chain at the first set of nodes; current should not enter or leave at the second set.

Circulating the earth point

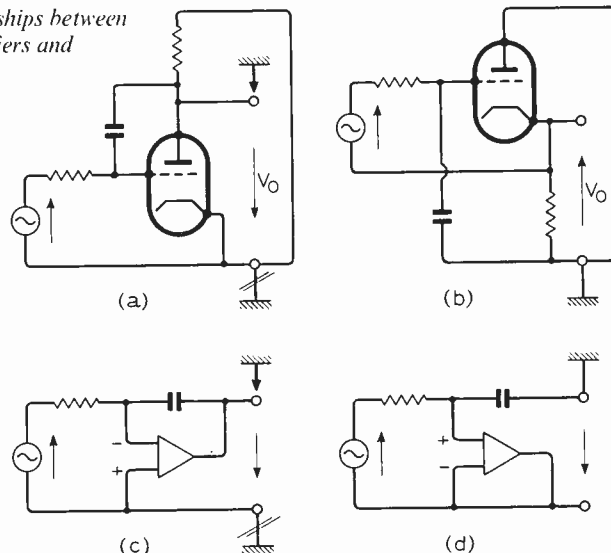
The first step then is as shown in Fig. 5, diagram 2. The earth connection is moved to the next low-impedance node, and the output terminal of the amplifier that was connected to that node is connected to that node just vacated by the earth connection. By using the argument already given for feedback arrangements of the type shown in Fig. 2, it can be seen that to perform its new function correctly this amplifier must also have its input connections reversed so that the node one removed from the right-hand end of the chain may follow (i.e. be caused to have virtually the same potential as) the node connected to the other input terminal. (The reader may like to relate these steps to the familiar example shown in Fig. 6, where the bootstrapped integrator (b) is simply the Blumlein integrator (a) with only a change of earth, and (d) and (c) respectively are the corresponding circuits using operational amplifiers.) This process may then be repeated, until, when the earth connection has made four circuits of the outer loop, diagram 1 is again obtained—twelve connection diagrams.

To arrive at the more intelligible layouts presented in Fig. 3, the diagrams of Fig. 5 have to be re-ordered so that the earth point on each appears in the same position as it does in diagram 1. When this is done the amplifiers and passive elements in the diagrams of Fig. 5 actually take up the positions in Fig. 3 as shown by the numbering. Later it will be found that some economy in analysis and tabulation is possible if the circuit elements are consistently identified according to the positions they occupy in Fig. 5.

Some of the circuits have been described in published papers (Refs. 1, 2), but they seem to have been discovered individually rather than as members of a family. Diagram 5 gives the two-differentiator circuit, Part 15, Fig. 22, which is therefore also a member of the family.

Moving the earth connection is not altogether the unimportant change it may appear. In the diagrams only the live output terminals of the amplifiers are shown. The other or "neutral" terminals are connected to the earth. Consequently, whereas in diagram 1 the output voltage of amplifier 1 is developed across the first two elements of the chain, in diagram 2 it is developed across the second two elements. Therefore, except when $A_1 \rightarrow \infty$, the input voltages of the amplifier in the two connections will be different and, as departures from ideal performance ($Q = \infty$) are caused by such input voltages, the damping caused by finite A_1 in the two connections may be expected to be different. Amplifier 2, on the other hand, still develops its output voltage across the

Fig. 6. Illustrating the relationships between high-gain sign-inverting amplifiers and voltage followers.



same pair of elements. Both the magnitude and sign of the damping terms caused by finite A_1 and A_2 differ for different circuits, and an analysis is given in a later section.

Ideally the natural motion of all the circuits is described by an equation of the form

$$(p^2 T^2 + 1)V = 0, \tag{1}$$

which defines simple harmonic motion. So it might be supposed that with high-gain amplifiers and a little added damping, say from lossy capacitors, all will be satisfactory high-Q circuits. Inspection of the circuit diagrams, however, suggests that not more than half the thirty-six circuits are likely to be satisfactory in practice.

First zero-frequency stability and the possibility of locking-over after a transient or overload. In two of the circuits to diagram 1 amplifier number 2 has a resistor connected from output terminal to positive input terminal, but in normal operation more than counterbalancing negative feedback is applied to the negative terminal through the other amplifier. If, however, this negative feedback should fail temporarily, it seems possible that amplifier number 2 could lock-over and become inoperative. Even more impractical looking (Fig. 7) is diagram 12 with passive-element string 1. Here the only zero-frequency feedback round each amplifier is positive. So the biasing is unstable, and each amplifier must in practice, sooner or later, slip into saturation.

Practical difficulty is also likely to be found with circuits in which an amplifier has a capacitor connected between output terminal and positive input terminal, or where an amplifier is working as a differentiator. For practical applications circuits chosen from those connected according to diagrams 2, 3, and 4 can be recommended as the most likely to be free from serious instability problems, and the present authors have had satisfactory performance from circuits connected according to diagram 1. Circuits connected according to diagrams 3 and 4 are favoured by A. Antoniou (Ref. 2).

The amplifiers must have open-loop frequency-response shaping virtually to the point that gives stability with 100% feedback, and in a very high-gain amplifier this

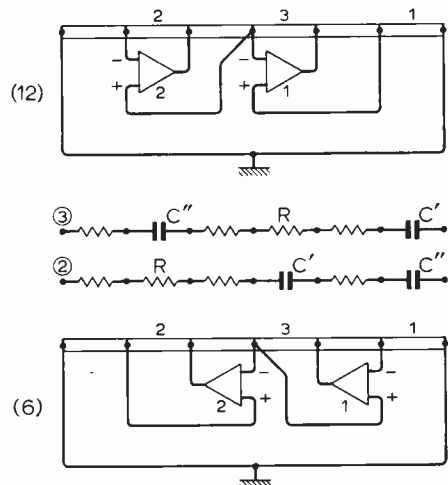


Fig. 7. Two more of the twelve diagrams laid out to conform with Fig. 3.

can mean beginning to cut at a frequency as low as 10 Hz. Thus at low audio frequencies the type 741 operational amplifier is very satisfactory; but above a few thousand Hz the cutting network becomes almost a short circuit, and an early stage overloads before the amplitude at the output has reached a useful level. For use at higher frequencies an amplifier of greater bandwidth and (probably) lower gain must be used.

An amplifier with a mosfet input stage can, because of its electrometer-like input impedance, be used with very high values of resistance, and hence can easily be used in circuits tuned to very low frequencies. A development model of such an amplifier in integrated form has been made by G.E.C., Wembley, and a similar circuit could be made from individual components. Ref. 3.

Some amplifiers are not suitable for the circuits described in this article, as they cannot accept more than a very small amplitude of common-mode signal in the positive direction.

As an example of the ideal working of the circuits consider Fig. 1(c), and call the output voltages of A_1 and A_2 V_1 and V_2 respectively. Then when $V_1 = 0$,

$$V_2 = \left(1 + \frac{Z_1}{R_1}\right) V_0 \tag{2}$$

$$V_2 = \left(1 + \frac{Z_2}{R_2}\right) V_0 - \frac{Z_2}{R_2} V_1 \quad (3)$$

$$= \left(1 - \frac{Z_2 Z_1}{R_2 R_1}\right) V_0 \quad (4)$$

So

$$V_0 = \frac{Z_3}{Z_3 + R_3} V_2 = \frac{Z_3}{Z_3 + R_3} \left(1 - \frac{Z_2 Z_1}{R_2 R_1}\right) V_0 \quad (5)$$

whence $V_0 = -\frac{Z_3 Z_2 Z_1}{R_3 R_2 R_1} V_0$. (6)

Thus, for example, when $Z_1 = 1/pC_1$, $Z_2 = 1/pC_2$, and $Z_3 = R_3$

$$V_0 = -\frac{1}{p^2 C_1 R_1 C_2 R_2} V_0 \quad (7)$$

i.e. $(1 + p^2 C_1 R_1 C_2 R_2) V_0 = 0$. (8)

The expression for V_2 when either $Z_1 = R_1$ or $Z_2 = R_2$ is $(1 - 1/pC_2 R_2) V_0$ or $(1 - 1/pC_1 R_1) V_0$, which confirms that when Z_3 is removed the impedance that the circuit presents is an inductance of value $C_2 R_2 R_3$ or $C_1 R_1 R_3$. (See Part 15.)

When both Z_1 and Z_2 are capacitances, the impedance is $p^2 C_1 R_1 C_2 R_2 R_3$, or in steady-state (sine-wave) terms

$$Z = -\omega^2 C_1 R_1 C_2 R_2 R_3, \quad (9)$$

a negative resistance increasing in magnitude as the square of the frequency.

To introduce intentional damping a resistance may be placed across one of the capacitances. Thus if $1/Z_2$ becomes $pC_2 + 1/R_d$, eqn. (8) becomes

$$(1 + pC_1 R_1 R_2 / R_d + p^2 C_1 R_1 C_2 R_2) V_0 = 0. \quad (10)$$

So, by comparison with the standard form $1 + pT/Q + p^2 T^2 = 0$, (11)

$$Q = \frac{R_d}{R_2} \sqrt{\frac{C_2 R_2}{C_1 R_1}} \quad (12)$$

$$= R_d / R_2 \text{ (when } C_2 R_2 = C_1 R_1) \quad (13)$$

Finite A_1 and A_2

When A_1 and/or A_2 are finite and real (i.e. constant for all significant frequencies) damping terms appear in the equation of motion; and it can be shown that these can be separated into those which would have been present had the same amplifiers been used in the same positions in the standard three-amplifier two-integrator loop (with an infinite gain amplifier in the third position) and others that correspond to incomplete bootstrapping. These can be either positive or negative, according to the circuit, and so may either add to or subtract from those of the first kind. A simple example of the effect of this was given in Part 15. Another aspect of the presence of damping terms is their effect on the impedance presented when Z_3 is removed. Because they produce voltages in quadrature with the ideal voltages, simulated inductance appears with parasitic resistance, and negative resistance with parasitic reactance.

In Fig. 2 the input voltage of the amplifier is V_0/A . i.e. this is the difference between the voltage on the positive terminal and the voltage on the negative terminal. In a true

bootstrap arrangement, a three-terminal amplifier complete with power supply raised up on a voltage V_2' , the true amplifier output voltage is $V_0 - V_2'$, and hence the true amplifier input voltage (reckoned from the common terminal to the live input terminal) is $(V_0 - V_2')/A$. So, if the voltages at all terminals joined to the passive-element chain are to be unchanged, V_2' must be less than V_2 by V_2'/A . Thus

$$V_2 - V_2' = V_2'/A \quad (14)$$

$$V_2 = V_2'(1 + 1/A) \quad (15)$$

$$V_2' = V_2 A / (A + 1). \quad (16)$$

So, to replace the differential-input amplifier of Fig. 2, the three-terminal amplifier must have a bootstrapping voltage KV_2 , K being the familiar factor $A/(A + 1)$.

Now subtract this voltage from all voltages around the amplifier, so that it may be earthed in the normal manner, by introducing two differential boxes as shown in Fig. 8. This too must be equivalent to Fig. 2; and if the factor K is thought of as

$$K = 1 - \frac{1}{A + 1} \quad (17)$$

it is seen that the incomplete bootstrapping obtained with a finite-gain differential amplifier is equivalent to perfect bootstrapping (with a voltage V_2) less bootstrapping with a voltage $V_2/(A + 1)$. The latter may then be treated as a feedback voltage $-V_2/(A + 1)$ superimposed on a fully bootstrapped circuit and entering the circuit as shown in Fig. 8. Thus the effect of finite gain in one differential amplifier may be calculated; and it is assumed that this may be done for each amplifier individually, and the effects added.

Perhaps the easiest way of sorting out the sign of a feedback giving a damping effect is to represent it as an added resistance in the original three-amplifier circuit, where the existence of three low-impedance nodes and three virtual earths makes this easy. If the added resistance is across one of the capacitors, the effect given is of positive damping. If it is placed so that there is an integrator and an inverting stage in the superimposed feedback path, the extra feedback is of the sign used to turn the circuit into an oscillator and the effect is that of negative damping.

Algebraic analysis

Fig. 2 may be analysed as follows, where I is the current through Z' and Z . The voltage at the negative input terminal of the amplifier is $V_2 - V_0/A$. Hence

$$I = \frac{1}{Z} \left\{ V_2 - \left(1 + \frac{1}{A}\right) V_0 \right\} \quad (18)$$

$$V_1 = \frac{Z'}{Z} \left\{ V_2 - \left(1 + \frac{1}{A}\right) V_0 \right\} + V_2 - \frac{V_0}{A}$$

$$= (1 + f) V_2 + \left\{ f + \frac{1 + f}{A} \right\} V_0 \quad (19)$$

where $f = Z'/Z$; i.e. where $f = pT$ when $Z' = R$ and $Z = 1/pC$, and is equal to 1 when $Z' = Z = R$.

Then, by using this result, an equation for a complete circuit containing one finite-gain amplifier and one of infinite gain can be obtained in the form

TABLE 2
Residue

12	$\frac{1}{A_2} (1 + f_2) (f_1 f_3 - 1)$	1
	$\frac{1}{A_1} (1 + f_1) (f_2 f_3 - 1)$	
2	$\frac{1}{A_2} f_1 (1 + f_2) (1 + f_3)$	3
	$\frac{1}{A_1} f_3 (1 + f_1) (1 + f_2)$	
4	$\frac{1}{A_2} (1 + f_1) (1 + f_2)$	5
	$\frac{1}{A_1} (1 + f_1) (1 + f_3)$	
6	$\frac{1}{A_2} (1 + f_2) (1 - f_1 f_3)$	
	$\frac{1}{A_1} (1 + f_1) (1 - f_2 f_3)$	

$$f_1 = \frac{R}{Z_1} \quad f_2 = \frac{R}{Z_2} \quad f_3 = \frac{R}{Z_3}$$

$$\frac{V_{in}}{V_{out}} = 1 + f_1 f_2 f_3 + (\text{residue}) \quad (20)$$

where $f_1 = R_1/Z_1$, $f_2 = R_2/Z_2$, $f_3 = R_3/Z_3$ (see Fig. 1c), and the assumption is made that the circuit is arranged for 1-p response.

The first two terms represent the ideal response; so the residue, containing terms in $1/A_1$ or $1/A_2$ (as the case may be) contains the terms which give divergence from ideal performance. Of these the constant terms and those in p^2 change the resonant frequency: the terms in p change the damping. These residues calculated in terms of f_1, f_2, f_3 , are characteristic of connection diagrams, and in Table 2 are listed for connection diagrams 12 to 6. The $1/A_1$ residue for diagram 12 is also correct for diagram 1, the $1/A_2$ residue for diagram 1 for diagram 2, and so on. Then by substituting $f_1 = R/Z_1$, etc. appropriate to the chosen string of passive elements, and taking the sum of the coefficients of the terms in p , the two component $1/Q$ s may be obtained for each circuit, and some results are given in Table 3 under $1/q_r$. These are for circuits with different T s, but with equal R s in the inverting stage: i.e. in effect for circuits with four equal R s, $C'R = T/x$, $C''R = xT$. The numbering of the strings is as in Figs 5 and 3. It is assumed as in Part 7 (Feb. 1970) that individually damping terms can be added:

$$\frac{1}{Q} = \frac{1}{Q_1} + \frac{1}{Q_2} \quad (21)$$

Table 2 shows that residues for diagrams 12 and 6 are of the same magnitude but of opposite sign. This is also true as between

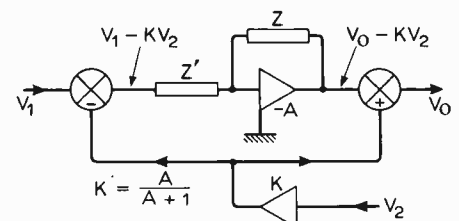


Fig. 8. Functional diagram equivalent to Fig. 2.

diagrams 1 and 7, 2 and 8, and so on. From Table 3 it can be seen that diagrams 7 through 11 will all have negative Q with finite gain.

When $x = 1$, and $A_1 = A_2 = A$, all the terms are $2/A$ or zero. The $2/A$ terms occur when the incomplete bootstrapping adds to the damping present in the standard circuit; the zeroes occur when the two effects for a particular amplifier cancel, or when that amplifier is in the inverting position. It is interesting to notice that for diagram 1 some entries tend to zero, others to $2/A$. For its neighbour, diagram 2, all the terms tend to $2/A$; for its neighbour in the other direction, diagram 12, all the terms are zero, or tend to zero and can become negative. Thus diagram 12 and its opposite number, 6, apparently offer infinite Q with finite gain; but they are likely to be difficult to use for reasons already given.

Parasitic phase shifts

The effects of phase shift within an amplifier cannot be calculated accurately without specifying its response in detail. Nevertheless some assessment of the comparative behaviour of the various circuit configurations is desirable; and the simplest model to take for this purpose is an amplifier whose frequency response is shaped by one dominant lag, time constant τ , i.e. with a characteristic defined by $A/(1+p\tau)$, the broken-line curve 3 in Fig. 9. In general this lag within the amplifier will be equivalent to a lag outside having a time constant equal to τ divided by the effective loop gain, i.e. of the order τ/A .

In the case of a loop based on two integrators, i.e. such as those in Fig. 3 with the

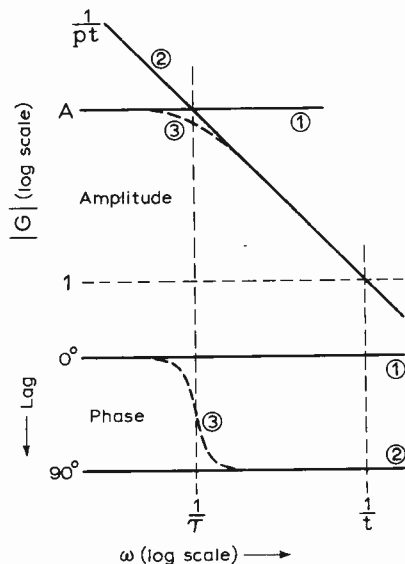


Fig. 9. Simplified amplifier characteristic assumed in the analysis.

amplifiers pointing to the right, an additional lag will add to the lagging phase shift given by the integrators and so contribute a negative increment of damping. Conversely in a loop based on two differentiators, amplifiers pointing to the left in Fig. 3, an additional lag will subtract from the leading phase shift given by the differentiators and so contribute a positive increment of damping. Thus in diagram 3, which is half-and-half, the effect may be expected to be minimised by opposing influences.

As in the previous section the departures from ideal performance can be separated into those present in the standard three-amplifier circuit, and those caused by using

differential amplifiers to give an approximation to bootstrapping; and the latter may again be thought of as additional feedbacks. If these feedbacks are positive the magnitude of the phase shifts caused by lags in the amplifiers is increased, i.e. if the amplifiers are acting as integrators there is further negative damping, and if as differentiators further positive damping: if the additional feedbacks are negative, damping terms in opposition to those of the standard loop appear. This behaviour is in contrast to the effects that the additional feedbacks have on the damping caused by finite aperiodic gain, for which, other things being equal, the effects are the same for both integrator and differentiator loops.

Analysis is most easily made if the effects of phase lag are considered independently of gain (assuming the effects from either cause to be small). Thus for phase effects, infinite zero-frequency gain is assumed and the amplifier response is taken to be represented by asymptote 2, Fig. 9, i.e. the internal response is given by the transfer function $1/pt$, where $t = \tau/A$. The sign reversal is already taken care of in the loop equations, so the residue may be evaluated using the expressions from Table 2 with $1/pt$ in place of A . The residues are of the form

$$apt + bp^2tT + cp^3tT^2.$$

As before, the term in p^2 is neglected. By comparison with the standard form the term apt is interpreted as contributing an increment of damping given by

$$\frac{1}{q} = \frac{at}{T} \tag{22}$$

Damping is positive if a is positive, and negative if a negative. Since at resonance $p^2T^2 \approx -1$, the effect of the term cp^3tT^2 will be essentially the same as $-cpt$, giving

$$\frac{1}{q} = -\frac{ct}{T} \tag{23}$$

This damping is negative if c is positive, if c is negative. The net effect, $(a-c)t/T$ is put in Table 3 under $1/q_s$.

A look at the combined effects of the results in Table 3 shows a prima facie case for preferring string 1 if diagram 1 is used, string 3 if diagram 3, string 2 if diagram 4, and either string 1 or string 3 if diagram 2.

Acknowledgements. From the many colleagues to whom we owe thanks, we should like to mention in particular P. J. Baxandall, R. C. Bowes, R. L. Ford, and S. W. Noble. On the subjects discussed in this and earlier Parts, their interest, comments, and encouragement, have over many years been of great assistance. (Series concluded.)

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2. A. Antoniou: "Realisation of gyrators using operational amplifiers, and their use in RC-active-network synthesis", *Proc. I.E.E.*, Nov. 1969, Vol. 116, No. 11, pp. 1838-1850.
3. D. J. Burt: "Marrying mos and bipolar transistors", *Electronics*, 3 Feb. 1969, Vol. 42, No. 3, p. 97.

TABLE 3 String

Circuit	String						Circuit
	1		2		3		
	$\frac{1}{q_r}$	$\frac{1}{q_s}$	$\frac{1}{q_r}$	$\frac{1}{q_s}$	$\frac{1}{q_r}$	$\frac{1}{q_s}$	
12	$\frac{1}{A_2} \left(x - \frac{1}{x} \right)$	$-\frac{2t_2}{T}$	0	$-\frac{4t_2}{T}$	$\frac{1}{A_2} \left(\frac{1}{x} - x \right)$	$-\frac{2t_2}{T}$	1
	0	$-\frac{4t_1}{T}$	$\frac{1}{A_1} \left(\frac{1}{x} - x \right)$	$-\frac{2t_1}{T}$	$\frac{1}{A_1} \left(x - \frac{1}{x} \right)$	$-\frac{2t_1}{T}$	
2	$\frac{1}{A_2} \left(x + \frac{1}{x} \right)$	0	$\frac{1}{A_2} (2x)$	$-\frac{2t_2}{T}$	$\frac{1}{A_2} \left(\frac{2}{x} \right)$	$-\frac{2t_2}{T}$	3
	$\frac{1}{A_1} (2x)$	$-\frac{2t_1}{T}$	$\frac{1}{A_1} \left(\frac{2}{x} \right)$	$-\frac{2t_1}{T}$	$\frac{1}{A_1} \left(x + \frac{1}{x} \right)$	0	
4	$\frac{1}{A_2} \left(\frac{2}{x} \right)$	$\frac{2t_2}{T}$	$\frac{1}{A_2} (2x)$	$\frac{2t_2}{T}$	$\frac{1}{A_2} \left(x + \frac{1}{x} \right)$	0	5
	$\frac{1}{A_1} (2x)$	$\frac{2t_1}{T}$	$\frac{1}{A_1} \left(x + \frac{1}{x} \right)$	0	$\frac{1}{A_1} \left(\frac{2}{x} \right)$	$\frac{2t_1}{T}$	
6	$\frac{1}{A_2} \left(\frac{1}{x} - x \right)$	$\frac{2t_2}{T}$	0	$\frac{4t_2}{T}$	$\frac{1}{A_2} \left(x - \frac{1}{x} \right)$	$\frac{2t_2}{T}$	
	0	$\frac{4t_1}{T}$	$\frac{1}{A_1} \left(x - \frac{1}{x} \right)$	$\frac{2t_1}{T}$	$\frac{1}{A_1} \left(\frac{1}{x} - x \right)$	$\frac{2t_1}{T}$	

World of Amateur Radio

Home construction

Many of the events which bring amateur radio to the notice of the public tend inevitably to give an impression that the vast majority of amateur equipment is commercially manufactured, and that the days of 'do-it-yourself' stations built on the proverbial kitchen table belong to the past as was suggested in the report on the R.S.G.B. show in the October issue. While unquestionably a lot of current s.s.b. transmitters and transceivers, and the bulk of h.f. communications receivers, are now factory built, it should not be forgotten that the extent of home construction and development of equipment is still considerable—and, in some respects, appears to be increasing rather than decreasing. A significant proportion of all v.h.f. converters, v.h.f. transmitters, almost all u.h.f. gear, many of the ancillary units and measuring instruments for all bands, and simple direct-conversion receivers and transceivers for 1.8 and 3.5 MHz are either 'home built' or represent heavily modified 'surplus' equipment. Then there is a minority of amateurs prepared to spend many months on building advanced receivers and transmitters—though only a small proportion of these are ever exhibited publicly.

Top-band activities

Although long-distance contacts in the 'top band' (1.8 MHz) are no longer confined to the winter months—the past summer for instance was notable for several U.K.-to-Brazil contacts—it is at this time of the year that the major interest in DX on this band occurs. "Most DX comes 'the hard way' and you have to really work for it—that is what makes it so intriguing" to quote that long-time top band enthusiast, Stew Perry, W1BB.

This winter there will be several new as well as established events to encourage amateurs to seek long-distance working on this often very surprising and rewarding band. The 1970-71 series of transatlantic tests (Sunday mornings from 05.00 to 07.30 G.M.T.) will be held on November 29th, December 27th, January 10th and 24th, and February 14th. Participants should observe carefully the time sequence (North

American stations call for the first five minutes, then Europeans for five minutes and so on) unless in contact. Europeans are expected to use the "DX window" (1823 to 1830 kHz) with American and Canadians between 1800 and 1820 kHz with west coast participants between 1975 and 2000 kHz. European to Japanese tests will take place following a series of transpacific tests on November 21st, December 5th and 19th, January 2nd and 16th, February 6th and 20th between 20.30 and 22.00 G.M.T. (The transpacific tests are between 13.30 and 16.00 G.M.T.—unlikely times for European DX operation). An A.R.R.L. 1.8-MHz contest is being held from 00.01 on December 12th to 16.00 G.M.T. on December 13th. Stew Perry, W1BB, suggests many new countries will be operating on 1.8 MHz this winter.

V.H.F. locations

For more than a decade, an approved system for indicating station locations has consisted of a single brief code group. For example, ZK23g would indicate The Needles, Isle of Wight, and ZL20g is an area near Digswell, Hertfordshire. This system can be used to transmit any location in Western Europe to an accuracy of about ± 2.5 km, but has never been particularly popular in the U.K. where the national grid reference has often been used for contest operation. The system, however, has been widely adopted on the Continent, as many operators have been discovering during recent tropospheric openings.

In this v.h.f. locator system, the first two letters represent a series of so-called "basic squares" each 2° longitude by 1° latitude, with the first letters U to Z covering areas to the west of the Greenwich meridian and the letters A to R east of Greenwich. Each basic square can be broken down into 80 smaller squares, indicated by the two figures, and there is a final sub-division into 9 units. It is possible to follow a series of basic rules to calculate the latitude and longitude represented by the code group, but in practice this is facilitated by using a special "QRA Locator" map. The I.A.R.U. Region I Bureau has recently published a new

low-cost card-printed map together with an explanation of how the system works. This map, almost essential nowadays to v.h.f. stations working to the Continent, is available for 1s 7d (including postage) from R.S.G.B., 35 Doughty Street, London W.C.1.

Poor operating discipline?

Over the years, at least on some amateur bands, there has been a sad deterioration in the orderly application of operating discipline and procedures. The competitive struggle to work ever more 'rare' long-distance stations is unquestionably a driving force behind many technical innovations. But it has unfortunately led also to a situation in which the DX station operator is almost obliged, whether he likes it or not, to hand out long strings of brief contacts under conditions of continuous interference from stations calling on the same frequency regardless of whether the station called is already in contact.

This can be accidental, in the 'pile-up' conditions that often occur, but it must be admitted, with regret, that a few amateurs, notably in certain European countries, now seem to pay little heed to accepted procedures. Recently, during a contact with an expatriate British amateur in Singapore, he complained bitterly of his attempts to work to the U.K. by means of directional CQ calls, only to have these calls answered "by all and sundry". He was even considering giving up the hobby because "of the bad and deteriorating manners that prevail on some bands".

In brief

The Irish Radio Transmitters Society has announced a change of address of its QSL bureau to: I.R.T.S. QSL Bureau, P.O. Box 462, Stella Avenue, Dublin 9, Ireland. . . . V.H.F. amateurs are initiating a series of special activity nights on 70 MHz (Wednesdays, 19.00 G.M.T. onwards). . . . The first two-way amateur television contact between the Isle of Man and Wales has been made by a group headed by Derek Whitehead, GD6FDZ/T, and John Lawrence, GW6JGA/T, at Prestatyn. . . . Membership of the R.S.G.B. rose by 1142 to a record total of 16,534 (including 1707 overseas members) in the year to June 30th, 1970. Over 200 local societies are in affiliation. . . . The Newbury and district society, after lying dormant for two years, has recently been revived and now holds meetings on the first Monday of each month at 19.30 at South Berks College, Oxford Road. (Details H. Newland, 42 Woodlands Road, Baughurst, Basingstoke). . . . Louis Varney, G5RV, whose callsign has for long been associated with a simple form of multi-band dipole and who became well-known for his early work on the reduction of television interference, is hoping to be active soon as VK9RV in Papua. . . . Alan Hitchcock of the Derby club won the 1970 national d.f. final; the 18 contestants had all previously qualified in a series of heats.

PAT HAWKER, G3VA

New Products

High-output Light-emitting Diodes

Diffused planar gallium phosphide/arsenide made by Monsanto Chemicals can be used in photographic darkrooms for the first time. Monsanto supply diodes made from this material to device and equipment makers for indicators and warning lights operating at high currents with a low duty cycle. Peak currents of up to 25A are allowed for 1 s in these new devices at rates of 300 pulses per second. Devices form a series, type MV4, with output levels of 17,000cd/m² at 2V, 1A. Wavelength of radiation is at 670 ± 20nm, well above the range of most photographic film. Mounted in a stud-type TO-5 header, the devices can withstand a continuous current of 1A. Price between £5 and £7 depending on quantity. Monsanto Chemicals Ltd, 10 Victoria St., London S.W.1.

WW 316 for further details

U.H.F. Field-effect Transistors

Intended for u.h.f. amplifiers and mixers the epoxy-clad Siliconix E300 f.e.t. is made from the same chip as the established 2N5397. It is a low-cost version, costing 15s 6d for a single device, reducing to 7s for large quantities. In the common-gate configuration, it has a power gain of 12dB at 450MHz with a noise figure of 4dB. Circuit adjustment can reduce this figure at the expense of 2 or 3dB of power gain. Siliconix Ltd, Saunders Way, Sketty, Swansea, Glam.

WW 302 for further details

Dual Monostable

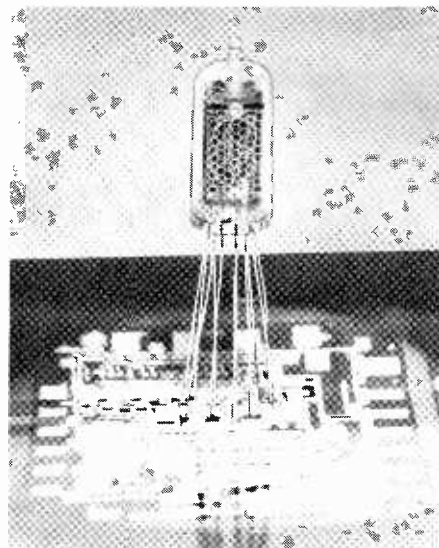
The Motorola MC675, from the high-threshold t.t.l. family, consists of two monostable circuits for use where high noise immunity is required. Pulse length is set using external components; additional flexibility is obtained by the use of a two-input NOR gate at the output, giving pulse stretching in addition to monostable action. Operating on threshold levels the circuits are independent of input pulse rise and fall times, and are not affected by high-level

noise of shorter duration than the pre-set pulse width. Power dissipation is 180mW and worst case propagation delay is typically 150ns. Jermyn Industries, Vestry Estate, Sevenoaks, Kent.

WW 303 for further details

Display Tube for I.C. Decoder/Drivers

Latest addition to the Hivac Numicator range is the XN24, a display tube specifically designed for easy interconnection with i.c. decoder/drivers. This side-viewing, gas-filled 0-9 digital display tube with 16mm high numerals and a viewing angle greater than 120°, is available in six types.



These include clear glass bulb, and red or amber filter versions, all available with 38mm leads for soldering direct into circuits, or 6mm leads for use with TS1 sockets. The height is 48mm and the diameter 19mm. Hivac Ltd, Stonefield Way, South Ruislip, Middx.

WW 313 for further details

Converter for 115V, 60Hz

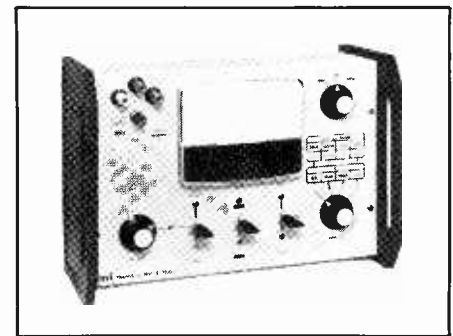
Saunders Electronics have introduced a 1.5kVA frequency converter providing a sinusoidal source of 60Hz at 115V derived

from a standard 50Hz/240V mains supply. The unit is designed to meet the needs of users of computers and peripheral equipment, recorders, instruments, teleprinters vibrators, card or tape readers and punches, data loggers, control systems motors and similar equipment designed for 115V 60Hz operation but for which the only available supply is 200/250V at 50Hz. The overall conversion efficiency of this solid-state design at full resistive load is better than 65%. Control is by push buttons and an alarm system operates on the failure of the 60Hz supply. Complete protection for the equipment is provided. The equipment is contained in a floor-standing cubicle measuring 1730mm high by 610mm wide and 610mm deep. Saunders Electronics Ltd, Hinckley, Leicestershire.

WW 312 for further details

Transistor Test Set

The ability to measure high gains (up to 2000) at very low collector currents (down to 1μA) is a feature of TF 2703, a transistor test set introduced by Marconi Instru-

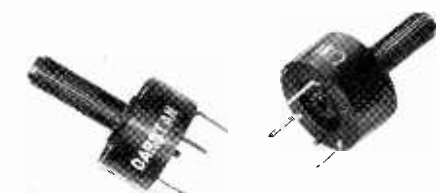


ments. The instrument is primarily intended for the rapid measurement of bipolar transistor d.c. parameters. It is simple to operate and is suitable for both p-n-p and n-p-n transistors and diodes other than zener diodes. It is powered by two 9V batteries, measures 216 × 318 × 127mm and weighs 3kg. The price is £67 f.o.b. U.K. Marconi Instruments Ltd, St. Albans, Herts.

WW 307 for further details

Wirewound Trimmer Pot.

A trimmer potentiometer, the P121, from Darstan is 20mm diameter and 11.65mm high and can be provided with a wide range of terminal spacings on 0.1in module, or otherwise. The case, lid and spindle are moulded in glass-filled nylon. Screwdriver adjustment can be made at the end of the spindle or from the underside of the potentiometer.



meter. The standard spindle length is 19mm, but lengths of 1, 13 and 25mm can be provided. The spindle is knurled. The potentiometer is wire wound and available in values from 5Ω to $50k\Omega$ with a standard tolerance of $\pm 10\%$. The power rating is 2W at 40°C derated to zero at 85°C . Prices range from 1s 4d to 3s each, depending on quantity and resistance value. Darstan Ltd, Church Road, Thundersley, Essex SS7 4QR.

WW 306 for further details

R.F. Signal Generator

High power r.f. source, model 470B-200 from Philco-Ford, is suitable for any application where a stable source of controllable r.f. power is required from milliwatts to greater than 50W. Characteristics include the following:

frequency range	50MHz to 200MHz
power output	50W (max)
stability	0.02% per hour
(after warm up)	
power stability	0.2dB/hour
(after warm up)	
frequency stability	0.003%
(short term)	

Power output is monitored on a front panel meter, a monitor output jack is provided for waveform analysis and

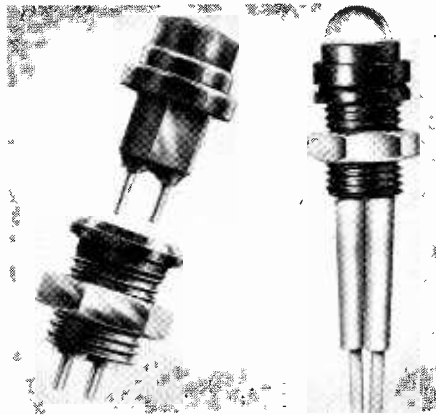


frequency calibration and protection circuits are provided. Wessex Electronics Ltd, Stover Trading Estate, Yate, Bristol BS17 5QP.

WW 305 for further details

Plug-in Indicator Lamps

A series of plug-in sub-miniature indicator lamps with glass or plastic lenses in a range of colours, is now available from Engel & Gibbs. They are claimed to be less expensive on initial outlay than conventional lampholders with separate lamps, and do not require access to the rear of the panel for replacement. The plug-in lampholders are made from glass-filled nylon and will fit a 6.35mm hole and a panel thickness of up to 5mm. They can have



either standard terminations or 6-inch-long insulated p.v.c. flying leads. The lamps are made for the range 5, 6, 12, 24 and 28V, and measure 8mm overall diameter with a frontal protrusion of 8mm for plastic lenses and 7mm for glass lenses. Lens colours available in both plastic and glass are: red, yellow, blue, green, orange. A plastic opal lens is also made. Engel & Gibbs Ltd, Elstree Way, Boreham Wood, Herts.

WW 314 for further details

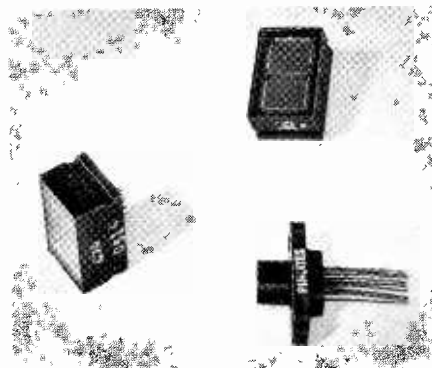
Metal-glaze Resistors

Insulated metal-glaze resistors, which have better load life stability than metal-oxide or carbon resistors, are available for both commercial and military use from Dubilier. Triple-rated, the resistors are well within the DEF 5115 load life stability limits of 3, 2 and 1% for $\frac{1}{4}$, $\frac{1}{8}$ and $\frac{1}{16}$ W ratings. The commercial version has a typical stability of $\pm 1\%$ for a $\frac{1}{2}$ W rating at 70°C . Stability is $\pm 0.7\%$ for $\frac{1}{4}$ W rating. This type of resistor is compared with other types in an article in the June issue (p.295/6). Available in 1, 2 and 5% tolerances from Dubilier Ltd, Victoria Rd, London W.3.

WW 315 for further details

Digital Readout Indicators

A range of hermetically sealed miniature Pinlite indicators is available from F.R. Electronics. The indicators are seven-segment digital readouts, each segment being a directly viewed incandescent filament giving a surface brightness of up to 5000 cd/m^2 . The segments are mounted in a single plane providing a viewing angle of about 120° . Character heights range from



$\frac{3}{16}$ in to $\frac{1}{2}$ in with operating voltages between 1.5 and 5 d.c. Units displaying a decimal point, or + and -, are also available. F.R. Electronics, Wimborne, Dorset BH21 2BJ.

WW 304 for further details

Conductive Inks, Solder Alloys, Solvents

Specially designed soldering materials for hybrid microcircuits include solders, inks, fluxes, and cleaners. Solders are intended for attaching leads and chips to microcircuits. Fluxes can be used on gold and silver thick and thin films. Conductive inks include silver-palladium and gold-palladium for thick film circuits. Solvents remove ionic materials and dry chips and substrates. Alpha Metals Inc, 56 Water Street, Jersey City, New Jersey 07304, U.S.A.

WW331 for further details

Low-cost Reed Switch

Claimed to be the lowest-cost precision proximity switch, a new reed switch and magnet can replace many existing microswitches where actuation by a ferrous material is possible. The iron-nickel contacts are normally held closed by a magnet and when ferrous material is held close, part of the flux is diverted, opening the contacts. Required force is 0.2gm. Various grades are available, the cheapest grade costing \$1.9 for quantities of 1000.



McClintock Matrixes Inc, Washington Road, Woodbury, Connecticut 06798, U.S.A.

WW330 for further details

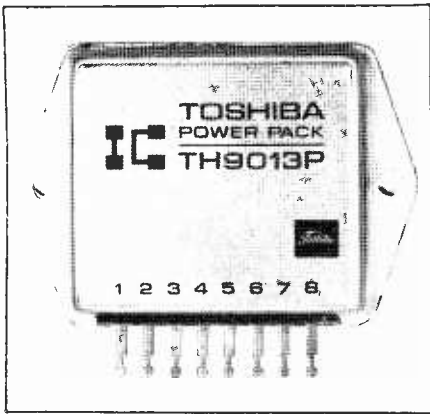
Metal Oxide Resistors

The S20 series of $\frac{1}{4}$ W metal-oxide resistors from WEL have a range of values from 10Ω to $1M\Omega$. Price is 4d each (100). WEL Components Ltd, 5 Loverock Road, Reading, Berks.

WW324 for further details

F.e.t.s and 20-watt hybrid i.c. amplifier

Integrated circuit, made by Toshiba, will give 20 watts output with a typical distortion of 0.2% (1kHz). Circuit operates from a 45-volt line or two 22.5-volt lines. Output transistors are Toshiba types 2SB 234 and 2SB 434. Pack which measures about $5 \times 5\text{ cm}$ includes seven



monolithic chips. Price is £2 14s for large quantities and £4 9s 6d for single units.

Also made by Toshiba are two f.e.ts, 2SK30 for audio use and 3SK22 for v.h.f. tuners. Designed as a low-noise input transistor, 2SK30 has a voltage rating of 50V and a noise figure of 0.5dB. Cost is 2s 5d—2s 11d for 1000 with a 60% mark-up for small quantities. The 3SK22 is a dual-gate device, for use as a mixer or as a gain-controlled amplifier. Price is 5s for 1000 up. Erie Electronics Ltd, South Denes, Great Yarmouth, Norfolk. WW328 (20-W i.c.), WW329 (f.e.ts) for further details

Printing Digital Voltmeter for under £300

A new digital voltmeter has a built-in printer for permanent display, as well as a normal panel display. It has two modes of operation—manual and automatic. In the manual mode printing is commanded and completed at the end of the analogue-digital conversion. In the automatic mode a print-out is made at the end of every conversion. Voltmeter is accurate to 0.1% ± 1 digit and has a common-mode rejection of 80dB with a 1000-ohm unbalanced input. Cost is less than £300. Pye TMC Ltd, Dunthorpe House, St Dunstons Street, Canterbury, Kent. WW325 for further details

H.F. S.S.B. Transceiver

Designed for long-range communication in marine and ground-to-air installations, the Hatfield s.s.b. transceiver covers 2-8MHz. Output is 25 watts p.e.p. from a single silicon transistor. Receiver gives 2 watts



output for 1 μ V input and has adjustable squelch threshold. Single-sideband transmitters are more efficient than d.s.b. types and also have an improved s/n ratio. Because d.s.b. transmission is still frequently used, this transceiver can also operate in this mode. Hatfield Instruments Ltd, Burrington Way, Plymouth PL5 3LZ, Devon.

WW332 for further details

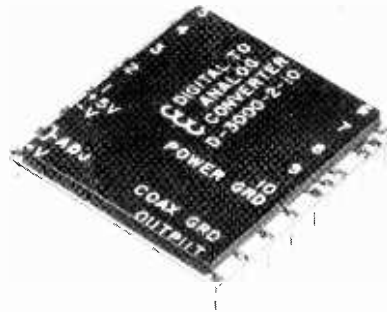
Packaged Oscillators

Range of oscillator modules covering 250kHz to 20MHz are available with either sine or square-wave outputs. This 5900 series includes a medium-stability type (7.5 ppm from 0 to 60°C) and a temperature-controlled type (0.1 ppm for the same range). The range will be extended to 120MHz and may be developed to use microcircuits. Cathodeon Crystals Ltd, Linton, Cambs.

WW335 for further details

Thick-film A to D Converters

Made by Columbia Components Corpn (U.S.A.) and available in the U.K. through Computer Controls Ltd, is a hybrid thick-film analogue-to-digital converter. Designated D-3000, the unit has a settling time of 50ns, and an output of 100mA at 5V driving a 50-ohm cable. It incorporates a

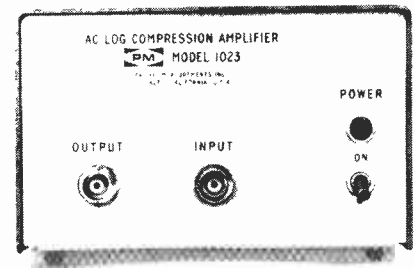


high slew-rate amplifier, ladder network, voltage reference source and high-speed switching circuits. Also available is a D-4000 digital-to-analogue converter, with a settling time of 100ns. Both measure about 3.8cm square and are t.l. compatible. Computer Controls Ltd, 19 Buckingham St., London W.C.2.

WW301 for further details

A.C. Log Compression Amplifier

The Pacific Measurements Inc. model 1023 log compression amplifier provides an output whose peak-to-peak amplitude is proportional to the logarithm of the peak-to-peak input amplitude. The frequency and phase of the output waveform will be the same as that of the input signal amplitude. The waveform will however be distorted because it is amplified logarithmically. Operating over a 60dB



dynamic range of input signal the unit compresses signal magnitude so that signals varying over a wide range can be observed on an oscilloscope. It provides an easy method of compressing the dynamic range required to monitor the signal at its output while preserving frequency and phase information. The peak-to-peak output amplitude changes by 100mV for each decibel of input amplitude change. The input impedance is similar to that of an oscilloscope and thus is compatible with accessory voltage divider probes manufactured for use with oscilloscopes. Output impedance is low so that signal loss can be avoided when driving low input impedance circuits. Wessex Electronics Ltd, Stover Trading Estate, Yate, Bristol BS17 5QP. WW308 for further details

Low-drift Op. Amps

Differential operational amplifiers with a drift of 3 μ V per month are supplied by Analog Devices Ltd. Designated 184 Series, the new amplifiers have a 100 μ V initial offset. By using a highly matched transistor input pair V_{BE} temperature drift is reduced to 0.25 μ V/degC. Common mode input impedance is 10⁹ ohms and rejection 10⁵. New series is designed for use as null detectors for bridge measurement, high impedance buffer amplifiers, comparators, instrument pre-amplifiers, and in other similar circuits. Direct current gain is 300,000. Analog Devices Ltd, 59 Eden St., Kingston-upon-Thames, Surrey. WW311 for further details

Integrated Light Sensor and Amplifier

Using thin-film techniques a silicon photo-detector, wideband amplifier and threshold logic are combined in a single 'pen'. Intended for use with c.r.t. displays, the photopen has spectral response from 400 to 700nm, with maximum sensitivity at 450nm, and can be used with P-4, P-7 and P-31 phosphors. Sanders Associates Inc, P.O. Box 907, Nashua, New Hampshire 03060, U.S.A.

WW340 for further details

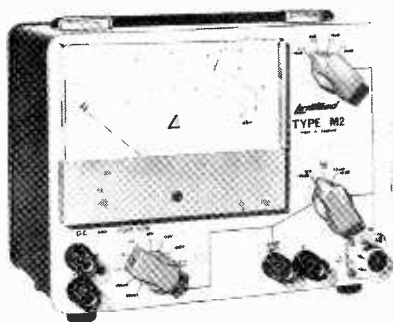
Low-power Op. Amp.

Monolithic operational amplifier with low power consumption is available from MCP Electronics. With minimum guaranteed operating voltage of ± 2.5 V, the

amplifier takes 30mA. Total current drain can be set with external components. Input offset voltage is 5mV and large signal voltage gain is 90dB at $\pm 15V$ and 80dB at $\pm 5V$ for a 1mA drain. Operating temperature range of the MP102 is -55 to $+100^{\circ}C$. MCP Electronics Ltd, Alperton, Wembley, HA0 4PE, Middx.
WW 310 for further details

Low-cost A.C./D.C. Millivoltmeter

Battery-operated millivoltmeter type M2 is specially suitable for teaching establishments. Priced at £30, it has d.c. ranges from

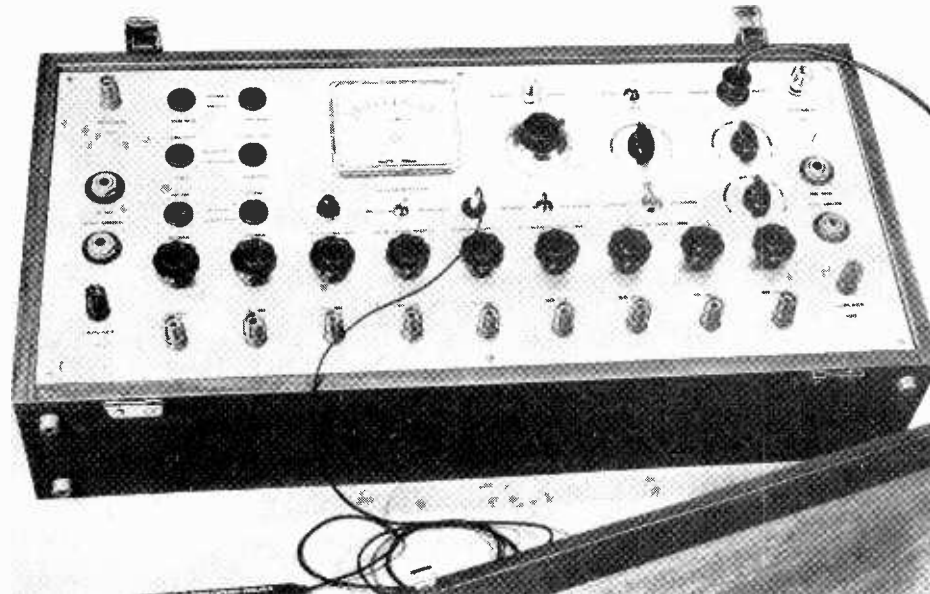


120mV to 400V and a.c. ranges from 1.2mV to 360V. Meter scale includes decibel markings. Linstead Electronics Ltd, Roslyn Works, Roslyn Road, London, N.15.

WW 333 for further details

Automatic Field Plotter

Automatic electric field plotting is provided by a new field plotter by Sensitised Coatings Ltd. Suitable for both electrostatic and electromagnetic fields, it is a form of analogue computer for solving flow and stress problems. It can be used for plotting electron paths in magnetron and klystron models, investigating dielectric stresses, aircraft static charges, electron lenses, high-voltage transformers and may also be useful for plotting aerial field strengths.



It is also useful for solving field distribution problems in aerodynamics, hydrodynamics and thermodynamics. The instrument solves two- and three-dimensional Laplace and Poisson equations using the resistive paper analogue method. Unique feature is automatic detection and marking of a selected equipotential line by the pen on resistive paper to an accuracy of 0.1%. Sensitised Coatings Ltd, 108 Church Street, Croydon CR0 1RE.

WW 317 for further details

High-Performance Op. Amps

Monolithic operational amplifiers are supplied in three grades by Silicon General Inc, of California. Specifications are given in the table. With a temperature range of -55 to $+125^{\circ}C$ they are available in TO-99 cans. A commercial version, temperature range 0 to $70^{\circ}C$, is also made and housed in dual in-line plastic packages. Devices, designated SG741, 747 and 748 (all electrically similar), can be used in place of normal 709 amplifiers when superior performance is needed.

	SG 748 A version AA version		
offset voltage	5mV	2mV	1mV
offset current	200nA	5nA	2nA
bias current	500nA	50nA	20nA
gain	50,000	100,000	100,000
common-mode rej.	70dB	80dB	86dB
price (\$)	3.80	12	24

Made by Silicon General Inc, of 7382 Bolsa Avenue, Westminster, Calif. 92683, U.S.A., they are available in the U.K. from Rastra Electronics Ltd, 275 King Street, London W.6.

WW 322 for further details

Plastic COS/MOS Devices

Complementary-symmetry m.o.s. devices made by R.C.A. are now available in plastic dual in-line packages. The CD4000E series is a broad range of gates and m.s.i. circuits featuring low power dissipation and high noise immunity. Quiescent power

dissipation of the gates is 50nW per package and 10mW per package for the m.s.i. circuits, both figures relating to a V_{DD} of 10V. They require a single unregulated supply of 5-15V and operate over the temperature range of -40 to $+85^{\circ}C$. R.C.A. Electronics Components Division, Sunbury-on-Thames, Middlesex.

WW 321 for further details

Cermet Potentiometers

Printed-circuit mounting is used in a new range of cermet-track potentiometers. Rated at 1 watt at $40^{\circ}C$ (de-rating to zero



at $125^{\circ}C$) they are 20-turn types. Resistance range is 100 ohms to 1 megohm. Reliance Controls Ltd, Drakes Way, Swindon, Wilts.

WW 334 for further details

Thin-film Video Amplifier

Circuit for video amplifier drives 75-ohm load with 1V pk-pk, normally beyond the capabilities of monolithic circuits STC claim. Output impedance is less than 0.1 ohm. External feedback connections determine gain. Designed by BBC Research Dept the amplifier is available from STC Ltd, Edinburgh Way, Harlow, Essex.

WW 309 for further details

Hybrid I.F. Amplifier

Thick-film i.f. amplifier with a 45-MHz bandwidth has been announced by Newmarket Transistors. Intended for use in television and radar amplifiers, it has a maximum noise output of 50mV and consumes 14mA at 14V. Gain can be controlled and the device, type NMC809A, is made in a dual in-line package. Newmarket Transistors Ltd, Exning Road, Newmarket, Suffolk.

WW 318 for further details

Photo-Darlington Transistors

Light sensors with a minimum hFE of either 2,500 or 5,000 are available in TO-98 packages. With controllable gain these can handle currents up to 250mA with V_{CEO} ratings of 25 and 40V. Designated 2N5777 to 2N5780 they are available from Jermyn Industries, Vestry Estate, Sevenoaks, Kent.

WW 319 for further details

Transistors for Television Receivers

New transistors for television and oscilloscope use have recently been introduced by AEG-Telefunken. Designated types BF 177-9 and BF 257-9 in TO-5 cases they feature controlled collector-base time constant and low feedback capacitance. BF 177 is intended for video output stages with small c.r.t.s and also for reference oscillator circuits in colour receivers. The higher voltage device, BF 178, can be used in monochrome receiver output stages. The BF 179, designed for PAL chrominance output stages, is available in three versions (A, B and C) with a V_{cer} of 160, 220 and 250V respectively. These devices are also made in TO-39 packages (BF 257-9). Made by AEG-Telefunken they can be obtained in the U.K. from MCP Electronics Ltd, Alperton, Wembley, HA0 4PE.

WW326 for further details

Log Diodes for Function Generators

Diodes with logarithmic $V-I$ characteristic are claimed by the manufacturers, CODI International, to be accurate to $\pm 2mV$. They are designed to operate over a three-decade range of current, from $10\mu A$ to $10mA$. The performance of every diode is said to be checked at over 800 testing points in the $V-I$ curve by computer. Available from CODI International, 30 Church Street, New York, N.Y. 10007, U.S.A.

WW323 for further details

10.7MHz Filters

Quartz filters made by Tyco (U.S.A.) are distributed in the U.K. by Steatite Insulations. With a centre frequency of 10.7MHz these filters have 6-dB bandwidths from 5 to 36MHz and are shock and vibration tested to MIL specifications.

60/6dB b-w ratio	1.8
100/6dB b-w ratio	2.7
insertion loss	3dB
in-band ripple	1dB
temperature range	-55 to +90°C

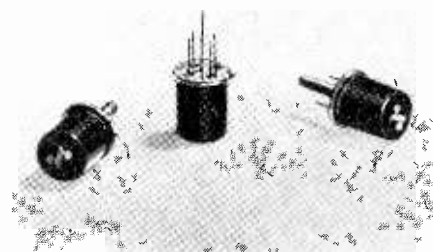
Steatite Insulations Ltd., Hagley House, Hagley Road, Birmingham 16.

WW320 for further details

Balanced Mixer

A double balanced mixer made by Anzac (U.S.A.) can also be used as a modulator, phase detector, attenuator and r.f. switch. The mixer has three mutually interchangeable ports for large signal input, small signal input and sum and difference outputs. Input frequencies can be from 2 to 500MHz. Isolation between local oscillator and r.f. ports is 25dB minimum and 35dB maximum at 50MHz. Type MAC-51 is packaged in a TO-5 case. Available in the U.K. through Wessex Electronics Ltd, Stover Trading Estate, Yate, Bristol BS17 5QP.

WW329 for further details



Darlington Power Transistors

Monolithic Darlington-pair transistors are now available with a dissipation of five watts from Pirgo Electronics, an affiliate of Sprague Electric Co. Intended for driving transducers from i.c.s, thermal printers and relays, they have high gain coupled with a good secondary breakdown characteristic. Typical parameters of type 1162 are given in the table.

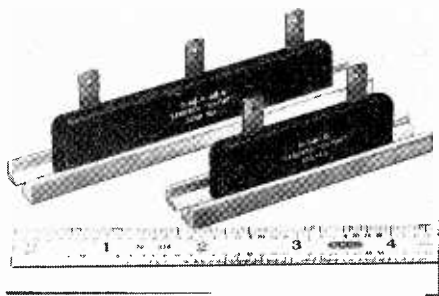
I_c	7A
$I_{c(pk)}$	10A
gain	2500
BV_{CBO}	60V
BV_{EBO}	12V
$BV_{CEO(sus)}$	60V
price	\$22.50 (1-99)
	\$15.90 (100-999)

Pirgo Electronics Inc, 130 Central Avenue, Farmingdale, Long Island, N.Y. 11735, U.S.A.

WW337 for further details

High-voltage Silicon Rectifiers

Rectifier diodes with peak reverse voltage ratings up to 15kV are made by Semtech (U.S.A.). They are available in two corona-free packages, one with a centre-tap for use in voltage doublers. Average rectified current is 0.4A, single-cycle surge current 50A and reverse current $10\mu A$ at the rated



reverse voltage (all at 25°C). U.K. agents are Bourns (Trimpot) Ltd, 17 High Street, Hounslow, Middlesex.

WW336 for further details

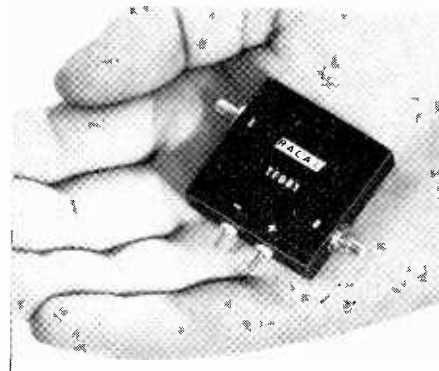
Balanced Modulator/Demodulator

For use in suppressed carrier modulation, synchronous detection, and frequency and phase detection, a planar epitaxial balanced modulator/demodulator is announced by Silicon General (U.S.A.). Output voltage is

the product of input signal voltage and a switching carrier function. Features include up to 65dB carrier suppression, 85dB common-mode rejection and adjustable gain. Silicon General Inc, 7382 Bolsa Avenue, Westminster, Calif. 92683, U.S.A. WW338 for further details

Wideband Thick-film Amplifier

A 450-MHz thick-film amplifier is available from Racal Research Ltd. Designed using the REDAC computer-aided design service it has a gain of 20 ± 1 dB and an output level of 5dBm. It is fitted with miniature



50-ohm coaxial connectors and is specially suitable for use as an input amplifier for oscilloscopes and digital voltmeters. It can also be used as a general-purpose 'buffer' amplifier. Racal Research Ltd, Newton, Tewkesbury, Gloucestershire GL20 8HE.

WW327 for further details

40-W U.H.F. Power Transistors

Made by TRW, the J-O series of transistors utilize tuning of the base lead inductance to enable high-power (40 watts) operation in the 225-400MHz band. Maximum bandwidth (equal to $r_b^2/2\pi L_{in}$) is achieved by making the base lead effectively a transmission line nearly $\lambda/4$ -long rather than a lumped inductance. The base lead is broken into sections loaded by appropriate capacitors, forming a lumped-constant equivalent of a transmission line. Using a number of such sections (the Wilkinson power divider technique) the transistors provide a purely resistive input impedance (8 ohms for the JO2001), an attenuation of the input signal at the low end of band giving a uniform output from 225-400MHz, and an input impedance at the base terminal compatible with 3dB couplers. A further point of interest in these transistors is a self-regulating power distribution mechanism, which protects the device against increased dissipation caused by a local decrease in base resistance. This works because the input impedance of the transmission correspondingly increases, as $Z_{in} = Z^2/Z_{load}$. Continuous collector current of type JO2001 is 9A, V_{CBO} is 60V and the total dissipation 115W. Available in the U.K. from MCP Electronics Ltd, Alperton, Wembley, Middlesex. HA0 4PE.

WW 339 for further details

Literature Received

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

ACTIVE DEVICES

A well illustrated booklet published by the Central Advertising Dept, Joseph Lucas Ltd, Great King St, Birmingham B19 2XF, describes, in a simple way, the operation and manufacture of semiconductors. The book is intended for older school children and is free of charge.

Plessey Microelectronics, Cheyney Manor, Swindon, Wiltshire, have recently started to produce an off-the-shelf range of medium scale integrated circuits using metal oxide silicon transistors. Data sheets which are currently available are:

MP 1402/3/4, 1024-bit dynamic shift register	WW 401
MP 1101/11, 256-bit, fully decoded, random access memory	WW 402
MP 1406/7, 1506/7, dual 100-bit dynamic shift register	WW 403
MP 1405, 512-bit, dynamic recirculating shift register	WW 404

Integrated active filters is the subject of a booklet received from Genisco Technology Corporation, 18435 Susana Rd, Compton, California 90221. The booklet includes data, application notes and ordering information WW 405

A price list of Zaerux valves as well as transistors, diodes and integrated circuits is obtainable from Z & I Aero Services Ltd, 44a Westbourne Grove, London W.2 WW 406

ITT Semiconductors, Footscray, Sidcup, Kent, have published a new semiconductor price list which includes a list of their franchised distributors WW 407

The following two data sheets are available from Damon/Electronics Division, 115 Fourth Avenue, Needham Heights, Massachusetts 02194, U.S.A.

Crystal filters, 10.7 and 21.4MHz	WW 408
6424WYA, 100kHz crystal oscillator	WW 409
6424WYA, 100kHz crystal oscillator	WW 409

An indicator lamp (type MTLED) consisting of a light-emitting diode, a transistor and associated resistors is the subject of a leaflet from Litton Precision Products, 95 High St, Slough, Bucks, the lamp is about 13 by 55mm and is compatible with d.t.l. and r.t.l. logic WW 410

Ferranti Ltd, Gem Mill, Chadderton, Oldham, Lancs, have sent us the following batch of literature:

equivalents list, Ferranti to CN	WW 412
t.t.l. general notes	WW 413
t.t.l. complex networks data sheets, counters, registers etc	WW 414
Micronor-5 (54 series, -55 to +125°C) t.t.l. standard range data sheets	WW 415
ZN 1002E, m.s.i. store matrix driver	WW 416
ZN 1010E, t.t.l. monostable	WW 417
ZLD709CE, 709 in d.i.l. package	WW 418

Erie Electronics Ltd, South Denes, Gt. Yarmouth, Norfolk, have a thick film manufacturing service which is described in a brochure WW 419

PASSIVE COMPONENTS

Reed relays manufactured by B & R Relays Ltd, Temple Fields, Harlow, Essex, are the subject of a leaflet WW 420

Mullard capacitors type C280 can now be supplied with crimped and cropped leads. Mullard Ltd, Torrington Place, London WC1E 7HD WW 421

'Terminal Catalog and Soldering Handbook' is the self-explanatory title of a booklet received from United Products Company, 55 South Penton Ave, New Hyde Park, New York 11040, U.S.A. WW 422

International Rectifier, Hurst Green, Oxted, Surrey, have announced that they are to market a range of paper-oil (not foil) capacitors. The capacitors are described in a leaflet WW 423

Erie Electronics, South Denes, Great Yarmouth, Norfolk, have produced a handsome loose leaf price list as a companion to their current catalogue WW 424

A 20-page booklet describes line matching and microphone transformers manufactured by Gardners Transformers Ltd, Christchurch, Hampshire BH23 3PN WW 425

HARDWARE

We have received a designer's handbook from MB Metals Ltd, Victoria Rd, Portslade, Sussex, which is called 'Flexible Filmwire Printed Circuits', and gives advice on the design and use of flexible printed circuits; a technical data section is included WW 426
Self-amalgamating tape is described in a leaflet published by Rotunda Ltd, Denton, Manchester. The tape, called Klingfast p.i.b. (polyisobutylene), is used to make water-tight joints on plastic and rubber insulated cables up to 46kV WW 427

Jermyn Industries, Vestry Estate, Sevenoaks, Kent, have available a card which contains a host of samples of plastic integrated circuit and transistor mounting pads. The samples are held by an adhesive which makes it possible to remove and replace the samples at will WW 428

EQUIPMENT

You can build your own 26-inch colour television set (£210) or colour monitor (£265 10s) from printed circuits, with components already mounted, supplied as a kit. Cabinet, tube, etc, are included. Use the reader reply service or write direct to CTC Electronics Ltd, 110 Leagrave Rd, Luton LU4 8HX, for descriptive literature and detailed price list WW 429

We have received the following data sheets from Sivers Lab, Box 42018, S-126 12 Stockholm 42, Sweden:

PM7101 vane attenuator, 0 to 60 dB, 5.8 to 18GHz	WW 430
PM7550 miniature coaxial switches, d.c. to 18GHz	WW 431
PM751/2/3 coaxial video attenuators, 2 to 18GHz	WW 432
PM7015X/P tuneable Gunn oscillators, 8.5 to 17GHz	WW 433
Price list for above	WW 434

Data on the following instruments have been received from SE (Engineering) Laboratories Ltd, North

Feltham Trading Estate, Feltham, Middlesex. SM213 d.v.m. using dual-ramp system.	WW 435
SM201 counter-timer, 100MHz, six digits	WW 436
SM112 'scope, d.c. to 100MHz, 10mV/cm	WW 437
SM215 d.v.m., accuracy 0.001% of reading	WW 438
EM102 oscilloscope series, general pamphlet	WW 439
Short-form catalogue, instruments	WW 440
Short-form catalogue, transducers, amplifiers, connectors etc	WW 441

Dana Electronics Ltd, Bilton Way, Dallow Rd, Luton, Beds, have produced two booklets which describe their range of digital instruments.

5800 series digital multimeters	WW 442
8000 series electronic counters	WW 443

Catalogue of the range of instruments produced by Dymar Electronics Ltd, Colonial Way, Radlett Rd, Watford, Herts WD2 4LA WW 444

A wide range of electronic test equipment is described in a leaflet from Wandel & Goltermann (U.K.) Ltd, 40-48 Acton High St, London W.3 WW 445

Details of a wide range of transducers and associated equipment are given in a leaflet from Kistler Instruments Ltd, The Ridges, 2 Clockhouse Rd, Farnborough, Hampshire WW 447

Marconi Instruments, St. Albans, Herts, have a leaflet which describes transistor test set type TF 2703 WW 448

Two brochures are available from the Solartron Electronic Group Ltd, Farnborough, Hampshire:

d.v.m. selector	WW 449
Frequency synthesizer, generators and frequency meters	WW 450

The 'product directory' of electronic equipment manufactured by Fluke International Corp., Garnett Close, Watford WD2 4TT, lists a wide variety of instruments WW 451

Test equipment from Hatfield Instruments Ltd, Burrington Way, Plymouth PL5 3LZ, Devon, is the subject of a short-form catalogue WW 452

GENERAL INFORMATION

Supplement No. 2 to BS3939: 'Graphical symbols for electrical power telecommunications and electronics diagrams' may be obtained from the British Standards Institution, 2 Park St, London W1A 2BS, price 14s.

'Basic principles of television distribution systems' is the title of a booklet published by Teleng Ltd, Arisdale Ave., South Ockendon, Essex RM15 5TR WW 453

We have received three more 'ten-bob' books (price includes postage) in the measurement and circuit concept series from: Tektronix U.K. Ltd, Beaverton House, Harpenden, Herts. They are called 'Sweep Generator Circuits' (second edition), 'Transducer Measurements' and 'Time-domain Reflectometry Measurements'.

Information sheet 1919 (14) is a map of v.h.f. transmitting stations for radios-2, -3, and -4. Engineering Information Department, B.B.C. Broadcasting House, London W1A 1AA. Also information sheet 1034 (14) lists these stations with frequencies.

'SI Engineering' is the latest publication on metrology to be published by the Council of Technical Examining Bodies. It may be obtained, price 5s, from The City & Guilds of London Institute, 76 Portland Place, London W1N 4AA.

'Ultrasonic cleaning techniques' is the title of a booklet from Dawe Instruments Ltd, Concord Rd, Western Ave., London W.3 WW 454

The Mullard Educational Service, Torrington Place, London WC1E 7HD, have produced a booklet called 'Educational Electronic Experiments' which costs 10s post free.

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- F.E.T. as a class-A audio amplifier, P. L. Matthews, 454 Sept.
- F.M. tuner survey, 465 Sept.
- 15-20W class-AB audio amplifier, J. L. Linsley Hood, 321 July, *Letters*, 387 Aug., 442 Sept., 489 Oct., 611 Dec.
- Four-quadrant division, 188 Apr.
- Frequency divider, 100MHz, D. R. Bowman, 389 Aug.
- doubler chain, m.o.s.t., *Circuit Idea*, J. A. Roberts, 12 Jan.
- standard and crystal oven, L. Nelson-Jones, 269 June
- Future of linear i.c.s, R. Hirst, 6 Jan., *Letters*, 75 Feb.
- Generating fast complementary pulses, *Circuit Idea*, A. Ivanov, 292 June
- Generator, constant current, *Circuit Idea*, M. Cadwallader, 12 Jan.
- current, B. L. Hart, 511 Oct.
- pulse, using integrated circuits, C. Djokic, 130 Mar., *Correction*, 164 Apr.
- Sawtooth, linear, *Circuit Idea*, S. F. Weber, 580 Dec.
- square pulse, unijunction transistor, *Circuit Ideas*, G. M. Paul, 123 Mar., P. J. Granger, 402 Aug.
- square-wave, 16 Jan.
- triangular waveform, *Circuit Idea*, S. Nagarajan, 292 June
- Goonhilly station uses microstrip circuits, 561 Nov.
- Graphical symbols, S. W. Amos, 63 Feb., *Letters*, 112 Mar., 213 May
- H.F. PREDICTIONS: 46 Jan., 94 Feb., 108 Mar., 154 Apr., 248 May, 262 June, 345 July, 376 Aug., 429 Sept., 514 Oct., 544 Nov., 603 Dec.
- Holograms on metal film, 332 July
- High-power amplifier, I. Hardcastle & B. Lane, 477 Oct.
- quality tape recorder, J. R. Stuart, 524 Nov., 587 Dec., *correction to part-1*, 595 Dec.
- Home video again, 529 Nov.
- Horn loudspeaker system, low-cost, 'Toneburst', 202 May, review, Ralph West, 263 June, *Letters*, 329 July
- I.C. manufacturing technique, advance, 72 Feb.
- I.E.A. show, exhibitors, 220 May
- Ignition system, capacitor discharge, R. M. Marston, 2 Jan., *Letters*, 109 Mar., 173 Apr., 213 May, 330 July
- Indicator, immersion heater, *Circuit Idea*, B. S. Crank, 316 July
- Improving the 13A oscilloscope, N. W. Vale, 451 Sept.
- Industrial telemetry, R. E. Young, 23 Jan.
- Instrumentation amplifier, A. E. Crump, 70 Feb.
- Integrated circuit stereo pre-amplifier, L. Nelson-Jones, 312 July, *Letters*, 442 Sept., *Correction*, 376 Aug.
- Letter from America, 408 Aug.
- LETTERS TO THE EDITOR:
- Aerial noise, H. F. Lewis, 388 Aug.
- Amateurs and television interference, H. S. Wood, 213 May
- Amplifier, class AB, D. S. Gibbs, 388 Aug.
- class AB—some questions, M. Walker & H. P. Smith, D. L. D. Mitchell, 442 Sept., 611 Dec.
- Dinsdale, mods, J. L. Linsley Hood, 74 Feb.
- Capacitor-discharge ignition, I. M. Shaw, J. F. Henderson, J. M. Bryant, M. J. Meadows, D. E. Bolton, 109 Mar., J. Vanderkooy, H. Harper, D. J. White, 173 Apr., D. Burn, 213 May, D. E. Bolton, M. A. Spence, 330 July
- Class AB amplifiers again, D. Mitchell, 611 Dec.
- Dice, electronic, S. E. Oliver, 268 June
- Digital Christmas tree, N. M. Morris, 111 Mar.
- Doctors in industry, D. A. Bell, 112 Mar.
- Dynamic range versus ambient noise, G. I. O'Veering, 331 July
- Electronics in medicine, R. E. George, A. S. Velate, 441 Sept.
- F.E.T. modulators, T. Mollinga, 214 May
- In defence of the amateur, K. Smith, 612 Dec.
- Industry and research in universities, B. G. Marchent, 268 June
- Intolerable tolerance code, A. Sproxtion, 549 Nov.
- Jupiter probe, J. M. Bryant, 442 Sept.
- 1 linear integrated circuits, J. Bryant, 75 Feb.
- Local radio, Basil Jackson, 74 Feb.
- Logic symbols, new, E. A. Foulkes, 111 Mar.
- Low-cost horn, W. Groomer, 329 July
- Measuring crossover distortion, J. F. Golding, 112 Mar., G. J. King, 172 Apr.

- Microelectronics, crisis in, D. A. Bell, 489 Oct.
 Millivoltmeter, op.amp. a.c., M. V. Dromgoole, 75 Feb.
 Mobile radio and amateur bands, R. A. Villiers, 441 Sept., J. A. Saxton, 489 Oct., I. M. Waters, 549 Nov.
 Modular pre-amplifier design, J. L. Linsley Hood, 174 Apr.
 Oscillator, sine-wave power, T. Roddam, 490 Oct., H. L. Armer, 549 Nov.
 Pre-amplifiers, i.c.s in stereo, L. R. Avery, 442 Sept.
 Rationalizing trade associations, H. E. F. Taylor, 441 Sept.
 Relay contact symbols, J. M. Little, 112 Mar.
 S.I. units, in defence of, B. J. C. Burrows and reply from Stanley Kelly, 74 Feb.
 Self-starting with ring-of-two, P. Williams, 331 July
 Some improvements in class B, M. J. Hamer, 267 June
 Sonex '70 report criticized, R. E. Cooke, 386 Aug.
 Stereo decoder adaptor, A. Royston, 11 Jan.
 — gramophone pickups, H. J. N. Riddle and reply from Stanley Kelly, 11 Jan.
 Symmetry in class B, A. H. King, 330 July, P. J. Baxandall, 386 Aug.
 The engineer in state and private enterprise, R. L. Clarke, 75 Feb., A. J. Sargent, 113 Mar.
 The i.c. industry, R. V. Killick, 549 Nov.
 The problem of dynamic range, Ivor Nedake, 214 May and reply from G. I. O'Veering, 331 July
 Theoretical and measured response, P. M. Quilter and reply from A. R. Bailey, 172 Apr., G. J. Bignold, 331 July
 'Toothpaste bass', I. M. Fried, 612 Dec.
 Transistor distortion characteristics, G. W. Short and reply from J. L. Linsley Hood, 10 Jan.
 V.H.F. services, T. Roddam, 172 Apr.
 Voltmeter, simple linear a.c., F. P. Mason, 113 Mar., D. L. Clay, 331 July, replies from G. W. Short, 113 Mar., 331 July, *Correction*, 164 Apr.
 Words, pictures and customs, W. W. Martin, 213 May
- Linear i.c.s, future of, R. Hirst, 6 Jan., *Letters*, 75 Feb.
 Linear scale millivoltmeter, A. J. Ewins, 592 Dec.
 LITERATURE RECEIVED: 46 Jan., 96 Feb., 140 Mar., 198 Apr., 248 May, 304 June, 362 July, 416 Aug., 458 Sept., 522 Oct., 572 Nov., 624 Dec.
 Loudspeaker performance, Paul W. Klipsch, 50 Feb.
 — units, moving-coil, the design and use of, E. J. Jordan, 533 Nov.
 Low-angle radiation, L. A. Moxon, 155 Apr.
 — cost horn loudspeaker system, 'Toneburst', 202 May, reviewed by Ralph West, 263 June, *Letters*, 329 July
 — distortion bias and erase oscillator, D. Griffiths, 20 Jan., *Correction*, 80 Feb.
 — pass and high-pass filter, combined, *Circuit Idea*, J. L. Linsley Hood, 123 Mar.
- Magnetic recording techniques for colour television, Aubrey Harris, 85 Feb.
 Magnetoresistance and its applications, B. E. Jones, 17 Jan.
 Matching, R. W. Cotterhill, 66 Feb.
 Measuring zero drain-current coefficient in f.e.t.s, *Circuit Idea*, P. R. Thrift, 206 May
 Mechanical filters for TV receivers, 320 July
 Mediator cleared for take-off in 1971, 120 Mar.
 MEETINGS: 44 Jan., 94 Feb., 139 Mar., 200 Apr., 230 May, 516 Oct., 540 Nov., 599 Dec.
 Metal glaze resistors, K. L. Dove, 295 June
 Metronome, electronic, D. T. Smith, 35 Jan., *Correction*, 80 Feb.
 Microammeter, d.c., 16 Jan.
 Microcircuit?, which type of, 276 June
 Microcircuits, elements of linear, T. D. Towers, 472 Oct., 545 Nov., 600 Dec.
 Microelectronics at the Paris components show, 215 May
 Millivoltmeter, linear scale, A. J. Ewins, 592 Dec.
 Mobile radio communication, 494 Oct.
 Modern direct voltage calibration system, H. Stern, 288 June
 Morse keyer, electronic, C. I. B. Trusson & M. R. Gleason, 379 Aug., *Correction*, 429 Sept.
 Multivibrator, astable, from logic gates, *Circuit Idea*, B. L. Hart, 182 Apr.
 — timing control, *Circuit Idea*, L. V. Gibbs, 206 May
- Negative resistance of transistor junction, *Circuit Idea*, J. A. H. Edwards, 12 Jan.
- NEW PRODUCTS: 39 Jan., 88 Feb., 142 Mar., 193 Apr., 249 May, 299 June, 355 July, 411 Aug., 459 Sept., 517 Oct., 567 Nov., 619 Dec.
- NEWS OF THE MONTH: 8 Jan., 61 Feb., 120 Mar., 167 Apr., 217 May, 264 June, 326 July, 382 Aug., 427 Sept., 492 Oct., 530 Nov., 581 Dec.
- Noise in transistors, F. N. H. Robinson, 339 July
 100MHz frequency divider, D. R. Bowman, 389 Aug.
- Oscillator, audio, digitally set, *Circuit Idea*, D. Taylor, 73 Feb.
 — bias and erase, low distortion, D. Griffiths, 20 Jan., *Correction*, 80 Feb.
 — circuit for measuring beta and f_T , *Circuit Idea*, J. P. Holland, 580 Dec.
 — long-tailed pair LC, *Circuit Idea*, D. T. Smith, 123 Mar.
 — low-distortion 30Hz-20kHz, *Circuit Idea*, C. A. Pye, 12 Jan.
- , push-pull, f.e.t., *Circuit Idea*, J. A. Roberts, 182 Apr.
 —, sine-wave power, *Circuit Idea*, H. L. Armer, 402 Aug., *Letters*, 490 Oct., 549 Nov.
 —, sinusoidal, for high temperatures, P. Williams, 332 July, *Correction*, 376 Aug.
 Oscilloscope, improving the 13A, N. W. Vale, 451 Sept.
- PERSONALITIES: 32 Jan., 87 Feb., 133 Mar., 192 Apr., 247 May, 298 June, 363 July, 410 Aug., 456 Sept., 496 Oct., 573 Nov., 604 Dec.
 Phase-locked stereo decoder, R. T. Portus & A. J. Hayward, 418 Sept.
 Physics exhibition, London, 241 May
 Plotting semiconductor characteristics, W. G. Allen, 211 May
 Postscript to simple class A amplifier and modular pre-amp., J. L. Linsley Hood, 607 Dec.
 Pre-amplifier, audio, simple, J. L. Linsley Hood, 207 May, *Correction*, 294 June
 —, integrated circuit stereo, L. Nelson-Jones, 312 July, *Letters*, 442 Sept., *Correction*, 376 Aug.
 Programmable unijunction transistor, O. Greiter, 430 Sept.
 Progress in tape recording techniques, S. Feldman, 33 Jan.
 Pulse generator using integrated circuits, C. Djokic, 130 Mar., *Correction*, 164 Apr.
- R.F. resistance and electroplating, 565 Nov.
 R.S.G.B. exhibition, 491 Oct., *Letter*, 612 Dec.
 Radio fire alarm, 158 Apr.
 Radiometer and temperature controller, YIG, I. J. Kampel, 501 Oct., *Correction*, 547 Nov.
 REAL & IMAGINARY: 'Vector', 48 Jan., 148 Mar., 254 May, 364 July, 464 Sept., 574 Nov., *Letters*, 172 Apr., *Correction*, 254 May
 Receiver, 80-metre s.s.b., W. B. de Ruyter, 106 Mar., *Correction*, 164 Apr.
 Receivers, domestic, 485 Oct.
 Regulator, battery, low-drain, *Circuit Idea*, P. Lacey, 548 Nov.
 —, d.c., variable voltage reference source for, Peter Williams, 83 Feb.
 Resistors, metal glaze, K. L. Dove, 295 June
 Resolver with gain control, *Circuit Idea*, C. N. Gordon, 73 Feb.
 Root Hog or Die, Thomas Roddam, 281 June
 Roots and responses, Thomas Roddam, 349 July
- Schmitt triggers, *Circuit Idea*, P. Gascoyne, 316 July
 Semiconductor characteristics, plotting, W. G. Allen, 211 May
 —, amorphous, J. E. Carroll, 13 Jan.
 Signal monitoring networks, A. E. Crump, 353 July
 Simple active filters, M. Bronzite, 117 Mar.
 — audio pre-amplifier, J. L. Linsley Hood, 207 May, *Correction*, 294 June
 — op. amp., D. Griffiths, 337 July
 Sinusoidal oscillator for high temperatures, P. Williams, 332 July, *Correction*, 376 Aug.
 Slant polarization for local radio, 492 Oct.
 SONEX '70. 171 Apr. report 'Cecilia—saint or temple prostitute?', 274 June, *Letters*, 386 Aug., 612 Dec.
 Sound '70, 222 May
 Speakers in corners, H. D. Harwood, 162 Apr.
 Spring song, Thomas Roddam, 226 May
 Stabilized power supply, A. J. Ewins, 150 Apr., *Letters*, 331 July, *Correction*, 240 May
 Stereo decoder, phase-locked, R. T. Portus & A. J. Hayward, 418 Sept.
 — test tone transmissions, 466 Sept.
 Switch, audio, *Circuit Idea*, J. du P. de Beer, 73 Feb.
 —, level-sensitive battery, *Circuit Idea*, N. L. Bolland, 236 May
 —, loudspeaker transmit/receive, *Circuit Idea*, D. A. Tong, 476 Oct.
- Swings and roundabouts, Thomas Roddam, 114 Mar.
 Symbols, graphical, S. W. Amos, 63 Feb., *Letters*, 112 Mar., 213 May
- TV centres, London's new colour, 104 Mar.
 Tape recorder, high-quality, J. R. Stuart, 524 Nov., 587 Dec., *correction to pt-1*, 595 Dec.
 — recorder pre-amplifier, digitally controlled, P. C. Grossi & C. Marcus, 127 Mar., *Correction*, 164 Apr.
 — recording techniques, progress in, S. Feldman, 33 Jan.
 Telecine film corrector, 84 Feb.
 Telemetry, industrial, R. E. Young, 23 Jan.
 Television wobulator, W. T. Cocking, 372 Aug., 423 Sept., 497 Oct.
 Temperature control, 'proportional' output stage for, *Circuit Idea*, M. Gluyas & B. W. James, 292 June
 Test record for audio systems, 116 Mar.
 Test your knowledge, L. Ibbotson, 45 Jan., 93 Feb., 141 Mar., *Correction*, 164 Apr.
 Thermostat, sensitive, *Circuit Idea*, A. Sewell, 316 July
 30-MHz video amplifier with variable power source, A. M. Pardoe, 597 Dec.
 Time, Thomas Roddam, 53 Feb.
 — delays, H. D. Harwood, 317 July, 398 Aug.
 Timebase, linear, with a u.j.t., *Circuit Idea*, M. Salem, 580 Dec.
 Timer, process, 188 Apr.
 Tone-balance control, R. Ambler, 124 Mar., *Correction*, 240 May
 — control circuit, P. B. Hutchinson, 538 Nov.
 Track-while-scan radar system, J. L. Sendles, 243 May
 Transient trinity, Thomas Roddam, 394 Aug.
 Transients, Thomas Roddam, 178 Apr.
 Transistor breakdown-voltage meter, Jens Langvad, 443 Sept.
 — tester, D. E. O'N. Waddington, 261 June
 Tuner design, u.h.f., 563 Nov.
- Ultra-low distortion class-A amplifier, L. Nelson-Jones, 98 Mar.
 Unijunction transistor, O. Greiter, 333 July, 403 Aug., 430 Sept.
- Variable voltage reference source for d.c. regulators, Peter Williams, 83 Feb.
- 'Vector' articles:
 'Yellow, and black, and pale and hectic red', 48 Jan.
 'Pipes and whistles in his sound', 148 Mar., *Letters*, 172 Apr., *Correction*, 254 May
 Electronic totem, 254 May
 Off the record, 364 July
 'Have we eaten the insane root that takes the reason prisoner?', 464 Sept.
 Good will towards dealers, 574 Nov.
 Vehicle location systems, R. A. Tyler, 435 Sept.
 Video disc, J. C. G. Gilbert, 377 Aug.
 — recording, colour electronic, Peter C. Goldmark, 366 Aug.
 — records, domestic, 340 July
 Voltage reference, cheap, *Circuit Idea*, E. R. Rumbo, 476 Oct.
- WESCON show 1970, Aubrey Harris, 541 Nov.
 Wobulator, television, W. T. Cocking, 372 Aug., 423 Sept., 497 Oct.
- WORLD OF AMATEUR RADIO: 34 Jan., 92 Feb., 132 Mar., 191 Apr., 223 May, 297 June, 361 July, 409 Aug., 457 Sept., 515 Oct., 566 Nov., 618 Dec. *Letters*, 213 May
- YIG radiometer and temperature controller, I. J. Kampel, 501 Oct., *Correction*, 547 Nov.

NEW PRODUCTS

- Active filters, RC (Kemo), 90 Feb.
 Adaptor, f.a.m. colour (Amplex), 356 July
 Aerial, broadband 'discone' (Microwave), 299 June
 — feed, microwave (Microwave), 300 June
 — mobile communication (Antenna Specialists) 301 June
 — v.h.f. communications (Panorama Radio), 414 Aug.
 Aircraft 'Homer' (Burndept), 143 Mar.
 Amplifier, A.C. log compression (Pacific Measurements), 621 Dec.
 —, audio, general purpose (Elcom), 39 Jan.
 —, audio, i.c. (U.E.C.L.), 252 May
 —, audio 3W, uses i.c. (Newmarket), 570 Nov.
 —, broad-band travelling-wave (M-O Valve), 43 Jan.
 —, data (Data Device Corp.), 146 Mar.
 —, high c.m.r. differential (Computing Techniques) 414 Aug.
 —, high-frequency video (Silicon General), 250 May
 —, hybrid i.c., 20-W (Toshiba), 620 Dec.
 —, hybrid i.f. (Newmarket), 622 Dec.
 —, instrument (Fylde), 145 Mar.
 —, lock-in, reference unit for (Brookdeal), 358 July
 —, logarithmic (AIM), 253 May
 —, low-cost i.c. (Motorola), 301 June
 —, low-noise S-band transistor (Watkins-Johnson), 411 Aug.
 —, measuring filter and frequency analyser (B & K), 355 July
 —, microphone (Mullard), 414 Aug.
 —, microphone, integrated (Polar), 89 Feb.
 —, operational (MCP), 621 Dec.
 —, operational differential (Burr-Brown), 302 June
 —, operational, f.e.t. (Advance), 43 Jan., (Computing

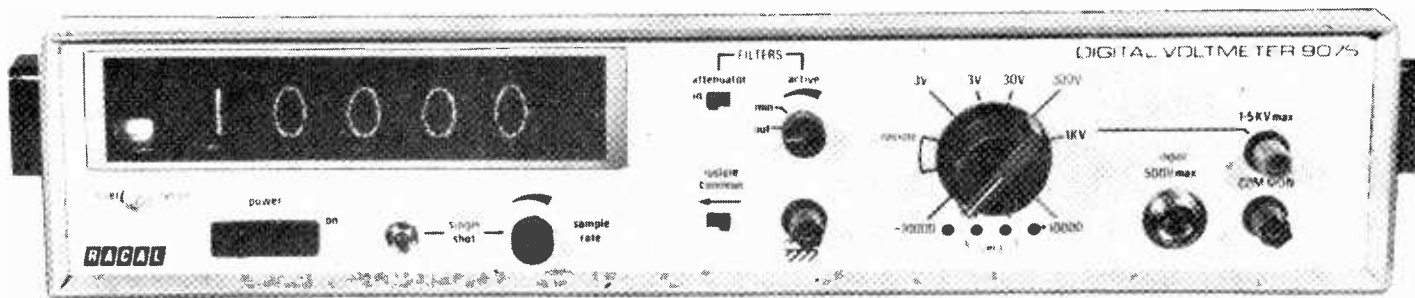
- Techniques, 88 Feb., (G.D.S.), 415 Aug.
- , operational, high-performance (Silicon General), 622 Dec.
- , operational, high power (Ancom), 252 May
- , operational, low drift (Advance), 197 Apr., (Analog Devices), 621 Dec.
- , operational, low input current (Comtec), 411 Aug.
- , power (Ancom), 303 June
- , power, TO-3 packaged (TRW), 249 May
- , S-band low-noise transistor (Watkins-Johnson), 194 Apr.
- , signal (Bournlea Dynamic), 461 Sept.
- , thin-film video (STC), 622 Dec.
- , 3.5-W (General Electric), 460 Sept.
- , TV boost (Teleng), 300 June
- , transistor, for 1-2GHz (Electro/Data) 43 Jan.
- , transmitter, add-on (M.E.L.), 42 Jan.
- , u.h.f. wired TV (Thorn Bendix), 300 June
- , wideband thick-film (Racal), 623 Dec.
- Analogue computer teaching system (Feedback), 40 Jan.
- Analysers, v.l.f., third-octave (AIM), 43 Jan.
- Attenuator, direct reading (Flann), 145 Mar.
- , 1-MHz, p-i-n diode (Hewlett-Packard), 567 Nov.
- Beam tetrode (M-O Valve), 143 Mar.
- Board, i.c. mounting (Shirehall), 250 May
- , p.c., instant (Oswald E. Boll), 301 June
- Bridge, auto, wide-range (Wayne Kerr), 249 May
- , universal (Wayne Kerr), 142 Mar.
- C.R.T., instrument (M-O Valve), 569 Nov.
- Cable identity tester (Information Computer Systems), 303 June
- Calibration sound source (B & K), 358 July
- Calibrator, colour monitor (Grafikon), 42 Jan.
- , d.c. voltage (Time Electronics), 568 Nov.
- Capacitor, ceramic (Steatite), 571 Nov.
- , electrolytic (Steatite), 569 Nov.
- , miniature electrolytic (ITT), 302 June
- , miniature tantalum (Sprague), 300 June
- , monolithic mica (WEL), 519 Oct.
- , polystyrene foil (ITT), 89 Feb.
- , single-film silvered mica (Sprague), 303 June
- , solid tantalum (Emihus), 91 Feb.
- , tantalum, small (Union Carbide), 521 Oct.
- , trimmer (Polar), 302 June, (JFD Electronics), 411 Aug.
- , trimmer, precision (Voltronics), 89 Feb.
- Cartridge, magnetic (K.E.F.), 142 Mar.
- Case, equipment (Case Systems), 197 Apr.
- Cassette tape editing kit (Multicore), 143 Mar.
- Cell, reference voltage (Muirhead), 413 Aug.
- Choke, sub-miniature (Cambion), 359 July
- Coaxial plug, solderless (Belling-Lee), 569 Nov.
- Coils for p.c. boards (Cambion), 250 May
- Conductive inks/solder alloys/solvents (Alpha Metals), 620 Dec.
- Connector, coaxial, multi micro (Radiall), 91 Feb.
- , 64-pole, p.c.b. (ITT), 414 Aug.
- Converter, d.c./d.c., (Plessey), 40 Jan.
- , digital-to-synchro (Moore Reed), 356 July
- , frequency (R. Gillfillan), 299 June
- , frequency, 50 to 60Hz (Saunders), 619 Dec.
- , thick-film A to D (Columbia Components), 621 Dec.
- Counter, automatic (Dana), 40 Jan.
- , computing (Racal), 359 July
- , 50MHz (Dawe), 568 Nov.
- , presettable (Industrial Numerical Controls), 251 May
- , reversible decade (Quarndon), 412 Aug.
- , 32-MHz (General Radio), 40 Jan.
- Counter/timer, 40-MHz (Marconi), 42 Jan.
- , 50-MHz (Marconi), 197 Apr.
- Counting, storing and display, modular system (Contraves), 359 July
- Coupler, signal, contactless (MCP), 251 May
- Crystal units, quartz (McKnight), 460 Sept.
- Current source (Keithley), 302 June
- D.V.M., six-digit system (Dana), 567 Nov.
- Delay line, PAL (Impetron), 299 June
- , line, video (Matthey Printed Products), 251 May
- Demodulator i.c., f.m./a.m. (Signetics), 299 June
- Desoldering wick (Southern Watch & Clock), 91 Feb.
- Dial, turns-counting (R. C. Knight), 39 Jan.
- Digital probe (Aircraft Supplies), 413 Aug.
- Diode assemblies, silicon, centre tap (Westinghouse), 462 Sept.
- , 18-GHz detector (AEI Semiconductors), 568 Nov.
- , electroluminescent (Plessey), 411 Aug.
- , 5W, 7-12.5GHz pulsed Gunn-effect (Plessey), 412 Aug.
- , GaP light-emitter, plastic (Jermyn), 570 Nov.
- , light-emitting (Monsanto), 39 Jan., 619 Dec., (Litton), 570 Nov.
- , log, for function generators (Codi International), 623 Dec.
- , 1-12GHz mixer (Mullard), 570 Nov.
- , photoelectric, silicon (EMI), 91 Feb.
- , rectifier, 1 amp (WEL), 569 Nov.
- , switching (Mullard), 463 Sept.
- , 300-W zener (Mullard), 518 Oct.
- , varactor, high-Q, for X-band (Mullard), 197 Apr.
- , zener, miniature (Mullard), 413 Aug.
- 'Diode, zener', programmable (Jermyn), 570 Nov.
- Display tube for i.c. decoder/drivers (Hivac), 619 Dec.
- Duplexer, microwave (Motorola), 462 Sept.
- Electromyograph (Isleworth Electronics), 39 Jan.
- Faders and potentiometric controls, audio (Penny & Giles), 89 Feb.
- Field plotter, automatic (Sensitized Coating), 622 Dec.
- Filter, active, low-pass (Lionmount), 43 Jan.
- , coherent (Brookdeal), 193 Apr.
- , crystal, for marine communications (Plessey), 519 Oct.
- , crystal, monolithic (ITT), 357 July
- , for marine communications (Cathodeon), 249 May
- , 10.7MHz (Tyco), 623 Dec.
- , tunable (KEMO), 252 May
- , variable (Barr & Stroud), 359 July
- Flying spot scanner tube (Mullard), 459 Sept.
- Galvanometer for education (Educational Measurements), 570 Nov.
- Generator, colour TV grey-scale (Marconi), 250 May
- , colour TV pattern (Pye Unicam), 195 Apr.
- , f.m. signal (Measurements), 249 May
- , function (Wavetek), 460 Sept.
- , noise (ADM), 303 June
- , noise, modular (ADM), 143 Mar.
- , pulse (Farnell), 145 Mar.
- , pulse, modular (Bradley), 461 Sept.
- , r.f. (Philco-Ford), 620 Dec.
- , servo carrier (Prosser Scientific), 360 July
- , time scale (Comark), 460 Sept.
- , voltage-coupled waveform (Environmental Equipments), 197 Apr.
- Gunn-effect devices (Mullard), 571 Nov.
- Heat sink adaptors (Jermyn), 146 Mar.
- , sink, flat-based (Jermyn), 145 Mar.
- , sink for i.c. module (Redpoint), 300 June
- I.C. for TV sound systems (RCA), 299 June
- I.C.s for data communication interfaces (Motorola), 300 June
- Indicator, digital readout (F. R. Electronics), 620 Dec.
- , lamps, plug-in (Engel & Gibbs), 620 Dec.
- , logic-state (Newmarket), 301 June, (Hewlett-Packard), 517 Oct.
- , low-voltage neon (Solitronics), 567 Nov.
- , solid-state plus/minus sign (Hewlett-Packard), 463 Sept.
- , transducer read-out (Southern Instruments), 359 July
- , tube driver, (K.G.M.), 88 Feb.
- , tube, low-voltage (Counting Instruments), 144 Mar.
- , tuning, monolithic (Motorola), 520 Oct.
- , units symbol nomenclature tubes (Hivac), 521 Oct.
- Inductor cores for p.c. boards (Mullard), 568 Nov.
- Insulation tester, transportable (Miles Hivolt), 42 Jan.
- Inverter assemblies (Gardners), 299 June
- Iron, quick-heating (Home Radio), 459 Sept.
- Isolator/coupler, electronic (Cole), 303 June
- Jack receptacle, 3mm (Sealectro), 40 Jan.
- Klystron, reflex (English Electric Valve), 414 Aug.
- Ladder network, binary (Morganite), 40 Jan.
- Light detector, silicon infra-red (MCP), 519 Oct.
- , sensor and amplifier, integrated (Sanders), 621 Dec.
- Logic: dual monostable (Motorola), 619 Dec.
- ; dynamic shift register (National Semiconductor), 518 Oct.
- ; elements, high-level (SGS), 569 Nov.
- ; function analyser (ITT), 567 Nov.
- ; high noise-immunity i.c.s. (Mullard), 415 Aug.
- ; m.o.s. adaptive, for pattern recognition (Integrated Photomatrix), 518 Oct.
- ; m.o.s. multiplex switches (Marconi-Elliott), 517 Oct.
- ; m.o.s. random access memory (Motorola), 415 Aug.
- ; m.o.s. shift register (Plessey), 413 Aug.
- ; m.o.s. shift register for delay lines and memories (National Semiconductor), 571 Nov.
- ; plastic COS/MOS devices (RCA), 622 Dec.
- ; semiconductor random-access memory (Motorola), 358 July
- , t.t.l. integrated circuits (Mullard), 146 Mar.
- Magnetic shield, flexible (Telcon), 571 Nov.
- Meter, a.f. power portable (Dymar), 360 July
- , digital, panel (Bach-Simpson), 91 Feb., (Computer Techniques), 571 Nov.
- , distortion factor (Sign Electronics), 358 July, (Marconi), 520 Oct.
- , field strength (Microwave), 252 May
- , frequency, 50-MHz digital (Orbit), 461 Sept.
- , frequency-period (Racal), 146 Mar.
- , megohms, super (British Physical Labs.), 356 July
- , millivolts, r.f./u.h.f. (Millivac), 41 Jan.
- , modulation (Dymar), 43 Jan.
- Microcircuit, frequency-selective (Consumer Microcircuits), 145 Mar.
- , power (WEL), 415 Aug.
- Microphone, variable directivity (Jagor Interelectric), 91 Feb.
- , with interchangeable capsules (Bang & Olufsen), 569 Nov.
- Millivoltmeter, low-cost a.c./d.c. (Linstead), 622 Dec.
- Milliwatt test set (Hatfield), 193 Apr.
- Mixer, balanced (Anzac), 623 Dec.
- Mixer, video signal (Aston Micro-Electronics), 521 Oct.
- , waveguide balanced (Micro Metalsmiths), 301 June
- Mobile radio test set (Marconi), 412 Aug.
- Modulator/demodulator, balanced (Silicon General), 623 Dec.
- Monitor, a.m. (Gates Radio), 144 Mar.
- , C.C.T.V. (Cotron), 301 June
- , current (G. & M. Electronics), 249 May
- , frequency (Orbit), 194 Apr.
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- , digital (Dana), 253 May, (Solartron), 360 July
- , electronic (Bach-Simpson), 142 Mar.
- Multiplier, quarter squares (Computing Techniques), 517 Oct.
- , sources, solid-state (Microwave & Electronic Systems), 88 Feb.
- Organ, electronic, m.o.s. arrays for (WEL), 463 Sept.
- Oscillator, constant amplitude wideband (Hewlett-Packard), 90 Feb.
- , klystron (English Electric Valve), 463 Sept.
- , package (Cathodeon), 621 Dec.
- , sinewave 10Hz to 10MHz (Britec), 88 Feb.
- , test, wideband (Omega), 521 Oct.
- , tunable Gunn-effect (Microwave & Electronic), 357 July
- , wide-range (Levell), 360 July
- , X-band Gunn-diode (Sivers), 90 Feb.
- Oscilloscope, educational (Mitre), 42 Jan.
- , 50-MHz (Pye Unicam), 143 Mar.
- , for TV servicing (Pye Unicam), 194 Apr.
- , simple (Meteronic), 568 Nov.
- , 10-MHz, dual trace (Telequipment), 41 Jan.
- , X-Y display (Marconi), 415 Aug.
- Over-voltage protection unit (ITT), 42 Jan.
- Photoconductive cells (Guest), 571 Nov.
- Pickup cartridge (Bang & Olufsen), 519 Oct.
- Potentiometer, cermet (Bourns), 571 Nov., (Reliance), 622 Dec.
- , clutch/brake, precision (Fairchild), 249 May
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- , micro-miniature (Painton), 359 July
- , miniature wire-wound (Guest), 193 Apr.
- , motorized (Rayleigh), 90 Feb.
- , 10-turn (GDS), 252 May, (Bourns), 519 Oct.
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- R, C and L boxes (Lionmount), 413 Aug.
- Radiotelephone, marine v.h.f. (Cossor), 250 May
- , v.h.f. a.m. (S.T.C.), 143 Mar.
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- , dual-in-line, without contact bounce (Fifth Dimension), 570 Nov.
- , in TO-5 can (ITT), 570 Nov.
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Sh-h-h don't make a noise!

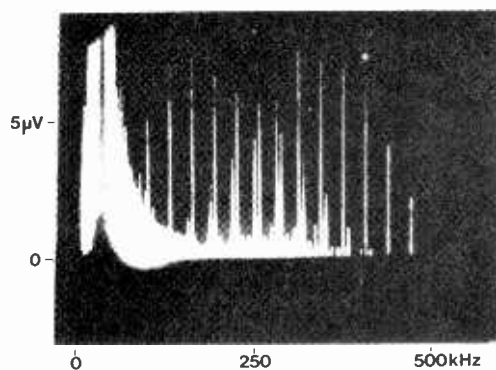


Well how much noise can a digital voltmeter make?

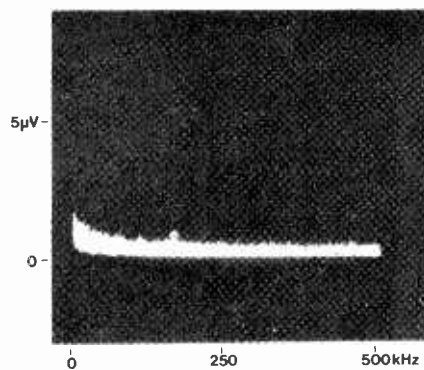
Some of them quite a lot.

Spillback from the input terminals of a digital voltmeter can result in errors – errors often attributed to external noise – particularly in systems measuring low level inputs. So down with spillback and welcome to the quiet one – the Racal 9075.

Still need convincing? Take a look.



"Spillback" from input terminals of a chopper stabilised digital voltmeter.



"Spillback" from Racal 9075 digital voltmeter.

And that's not all !

Full programmability. Accuracy 0.01%. Resolution 1 part in 30,000. 10 microvolt sensitivity. Scale length 29999 (over-range to 39999). Totally passive input. Auto-range versions available.

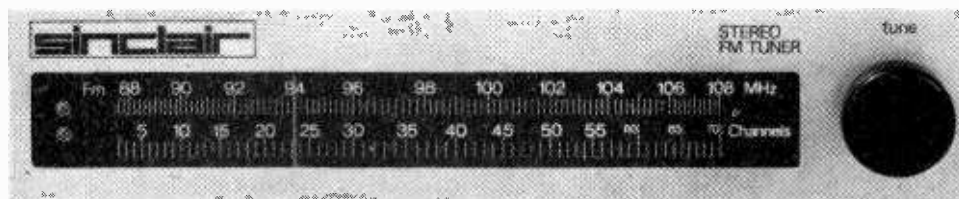
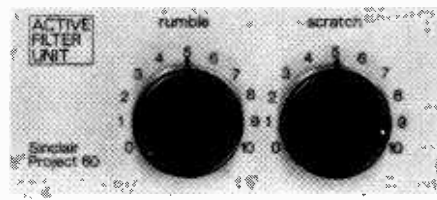
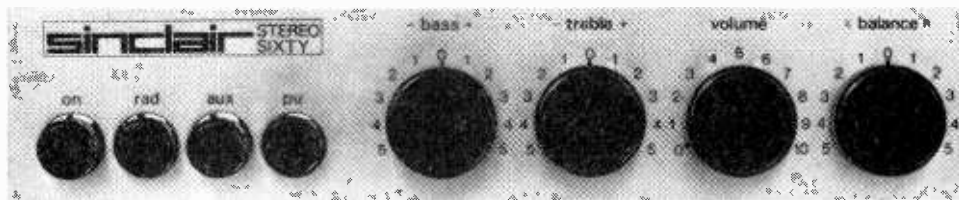
p.s. If this isn't the DVM for you, Racal have six others you can choose from.

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Project 60



the world's most advanced high fidelity modules

With the introduction of an entirely new and original high fidelity stereo F.M. tuner, the Project 60 range can be said at this stage to be complete. It offers the constructor a most attractive choice of modular arrangements whereby a high fidelity system can be selected to suit the user's personal requirements. Equally, it is possible to use any Project 60 modules separately or partially grouped and so benefit greatly from the flexibility in use these modules afford. The chart below shows some of the most popular applications for constructors to assemble. The Project 60 manual (free with the modules) suggests others as well and its 48 pages are packed with valuable information. The new tuner, for example can be used with any good high fidelity system as well as Project 60.

Project 60 now falls into four interdependent groups . - 1. The Z.30 and Z.50 amplifiers which have only 0.02% distortion at all output levels and are useful in a wide variety of other applications. 2. The control units comprising the Stereo 60 preamp and control unit and the Active Filter Unit (A.F.U.) with which both high pass and low pass filtering can be introduced between control unit and power amplifiers. 3. The Stereo F.M. tuner as described opposite ; and 4. The power supply units PZ.5,

PZ.6 and PZ.8. For most requirements when using Z.30 power amplifiers, the PZ.5 will be perfectly adequate ; if low efficiency (high quality) loud speakers are used, the PZ.6 stabilised power supply unit will be used. The PZ.8 will be needed with Z.50s which can be used for any Project 60 system.

Project 60 modules incorporate some of the most advanced circuitry in the world to achieve unsurpassed standards of high fidelity and modern manufacturing techniques enable these modules to be sold at exceptionally attractive prices. Assembling the modules requires no skill or previous experience since the manual supplied with the modules explains clearly how everything can be done with nothing more than the simplest of domestic tools.

Project 60 manuals

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.

System	The Units to use	In conjunction with	Cost of Units	+ Project 60 tuner
A Car Radio	Z.30	Existing car radio, Sinclair Micromatic	89/6	
B Simple battery powered record player	Z.30	Crystal pick-up, 12V or more battery supply and volume control	89/6	
C Mains powered record player	Z.30 and PZ.5	Crystal or ceramic P.U. Volume control etc.	£9.9.0	£34.9.0
D 20+ 20 watts R.M.S. stereo amplifier for most needs	Two Z.30s, Stereo 60 and PZ.5	Crystal, ceramic or magnetic P.U., most dynamic speakers, F.M. tuner etc.	£23.18.0	£48.18.0
E 20+ 20 watts R.M.S. stereo amplifier for use with low efficiency (high performance) speakers	Two Z.30s, Stereo 60 and PZ.6	High quality ceramic or magnetic P.U., F.M. Tuner, Tape Deck, etc All dynamic speakers	£26.18.0	£51.18.9
F 40+ 40 watts R.M.S. de-luxe stereo amplifier	Two Z.50s, Stereo 60 PZ.8 and mains transformer	As for E	£32.17.6	£57.17.6
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	Microphone, guitar, heavy duty speakers etc., controls	£17.8.6.	
J High pass and low pass filters	A.F.U.	D, E or F as above	£5.19.6	



Sinclair Radionics Limited, 22 Newmarket Road, Cambridge

Telephone (0223) 52731

WW—092 FOR FURTHER DETAILS

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 \pm 1dB
Distortion 0.02% into 8 ohms

Signal to noise ratio better than 70dB unweighted
Input sensitivity 250mV into 100 Kohms
 For speakers from 3 to 15 ohms impedance.
 Size 3½ x 2½ x ½ ins.



Z.50

Z.30

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**

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 \pm 1dB. 20 to 25,000Hz. Ceramic p.u. – up to 3mV; Aux – up to 3mV
- Output – 250mV.
- Signal-to-noise ratio – better than 70dB.
- Channel matching – within 1dB.
- Tone controls – TREBLE +15 to –15dB at 10kHz; BASS +15 to –15dB at 100Hz



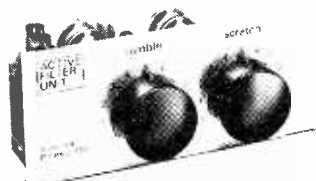
- Front panel – brushed aluminium with black knobs and controls.
- Size 8½ x 1½ x 4 ins

Built, tested and guaranteed **£9.19.6**

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.

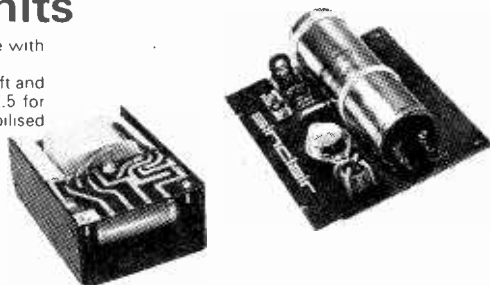
PZ-5 30 volts un stabilised £4.19.6

PZ-6 35 volts stabilised £7.19.6

PZ-8 45 volts stabilised

(less mains transformers) **£5.19.6**

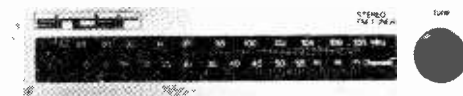
PZ-8 mains transformer £5.19.6



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.



Stereo FM tuner

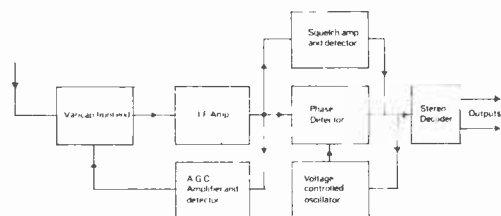


first in the world to use the phase lock loop principle

Before production of this tuner, the phase lock loop principle was used for receiving signals from space craft because of its vastly improved signal to noise ratio over other systems. Now, for the first time the principle has been applied to an FM tuner with fantastically good results. By the inclusion of other original features such as varicap diode tuning, printed circuit coils and an I.C. in the specially designed stereo decoder, the tuner has an unsurpassed specification, which also incorporates a squelch circuit for silent tuning between stations, A.F.C. and A.G.C. Sensitivity is such that good reception becomes possible in difficult areas, foreign stations can be tuned in suitable conditions and often a few inches of wire are enough for an aerial. In terms of high fidelity, this tuner has a lower level of distortion than any other tuner we know. Stereo broadcasts are received automatically as the tuning control is rotated, a panel indicator lighting up as the stereo signal is tuned in. Although the tuner is intended primarily for use with a Project 60 system, it can be used to advantage with any other high fidelity system. It is easily mounted into any cabinet as shown in the manual supplied with it

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 \pm 200 KHz
- Signal to noise ratio > 65dB
- Audio frequency response 10Hz–15kHz (\pm 1dB)
- Total harmonic distortion 0.15% for 30% modulation
- 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
- Operating voltage 25-30 VDC
- Size 3.6 x 1.6 x 8.15 inches: 91.5 x 40 x 207 mm



Price: £25 built and tested. Post free.

To: Sinclair Radionics Ltd., 22 Newmarket Road, Cambridge

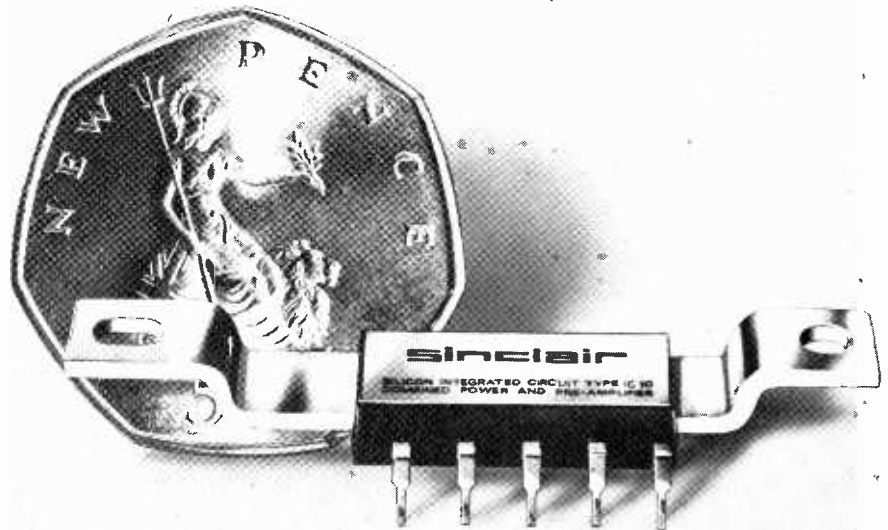
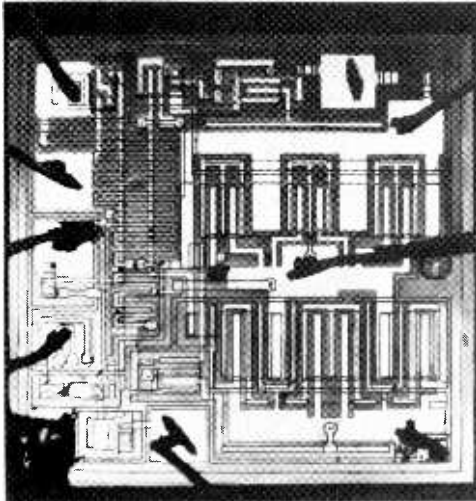
Please send

NAME

ADDRESS

for which I enclose cash/cheque money order

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 ± 1 dB
 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.

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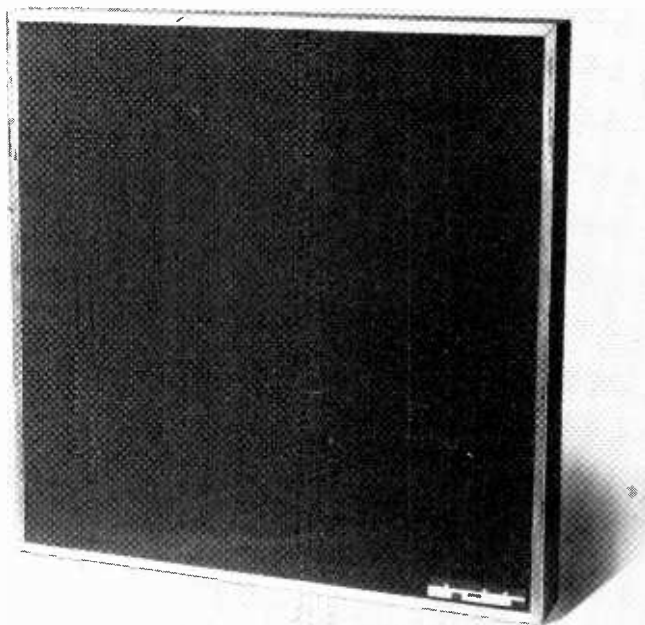
Q.16 High fidelity loudspeaker

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 $\frac{3}{8}$ " square on face x 4 $\frac{3}{8}$ " deep with neat pedestal base. Black all-over cellular foam front with natural solid teak surround.
Price:	£8 19 6.

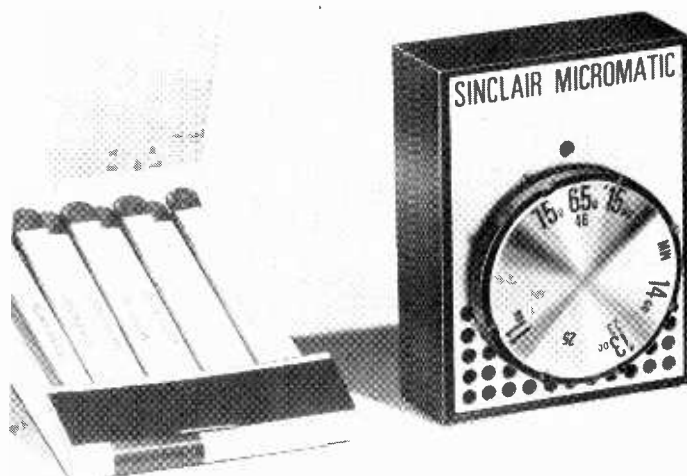


Micromatic Britain's smallest radio

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 $\frac{13}{16}$ " x 1 $\frac{7}{16}$ " x $\frac{1}{2}$ " (46 x 33 x 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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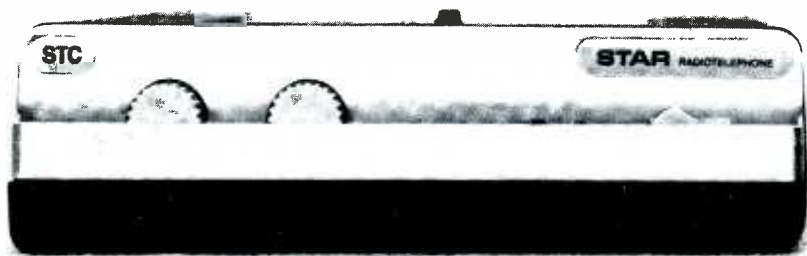
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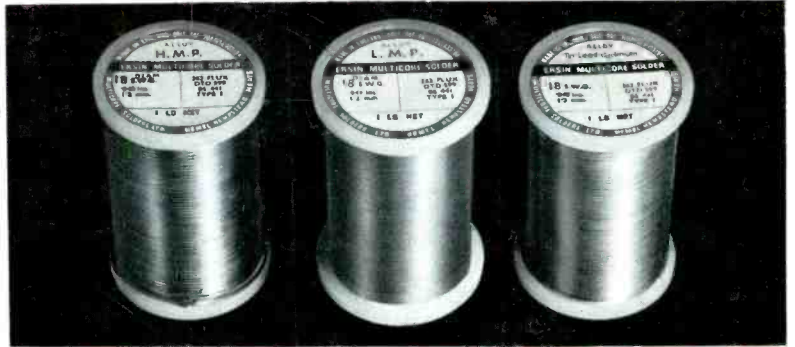
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MELTING POINT SOLDERS FOR SOLDERING

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- Silver plated surfaces
- High working temperature components

H.M.P.

HIGH MELTING POINT

For service at high temperature, or service at very low temperatures. Outstanding creep strength. Melting range 296°C - 301°C (565°F - 574°F).

Applications

A useful application of H.M.P. is the soldering of joints close to each other in such a way that the connections made first are not re-melted while later joints are made, with for example, a standard 60/40 alloy, melting point 188°C. Essential for use where high operating temperatures are experienced, for instance, electrical motors, car radiators, high temperature lamps. H.M.P. is also ideal for equipment, which is being operated in low temperatures, as it reduces the chance of the joint becoming brittle.

Specification

Multicore H.M.P. alloy complies with BS.219 Grade 5S. Supplied in a form of Ersin Multicore 5 core solder wire on 11b. or 71b. reels, incorporating Ersin 362 rosin based flux. This non-corrosive flux-cored solder wire complies with BS.441 and is available from 10 to 26 s.w.g., and in Multicore Solder Preforms. Ask for Technical Bulletin No. 1369.

L.M.P.

LOW MELTING POINT

A low melting point solder for soldering silver plated and gold plated surfaces. Melting point 179°C (354°F).

Applications

L.M.P. reduces the absorption of silver or gold into the solder alloy whilst soldering, and therefore, preserving the silver or gold plated surfaces. Also reduces the chance of a brittle joint being made.

NOTE

- a) The solution of gold into tin rises rapidly with temperature and so the use of L.M.P. Low Melting Point Solder is preferable.
- b) The solution rate of gold into tin is also reduced because L.M.P. is a ternary alloy comprising tin, lead and silver.

Specifications

L.M.P. is normally supplied in the form of Ersin Multicore 5 core solder wire, incorporating Ersin 362 rosin based flux, which complies with Min. Tech. specification D.T.D. 599A. It is available from 10 to 34 s.w.g. in 11b. or 71b. reels and Multicore Solder Preforms. Ask for Technical Bulletin 1469.

T.L.C.

EXTRA LOW MELTING POINT

Extra low melting point solder. Melting point 145°C (293°F).

Applications

T.L.C. alloy can be used whenever a soldered joint should be made with the minimum heat input. This would include heat sensitive transistors, flexible printed circuits and gold plated surfaces. The melting point of T.L.C. alloy is 38°C lower than any tin/lead alloy. Because of its low temperature application it is considered completely non-toxic in use unlike the high temperature cadmium-bearing brazing alloys.

Specification

T.L.C. alloy is normally supplied in the form of Ersin Multicore 5 core solder wire, incorporating Ersin 362 rosin based flux, which complies with Min. Tech. Specification D.T.D. 599A. T.L.C. alloy can also be supplied in the form of Multicore precision made solid solder wire. Extrusol extruded solid solder bars for solderbaths and Multicore Solder Preforms. Available from 10 to 34 s.w.g. on 11b. or 71b. reels. Ask for Technical Bulletin No. 1569.



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