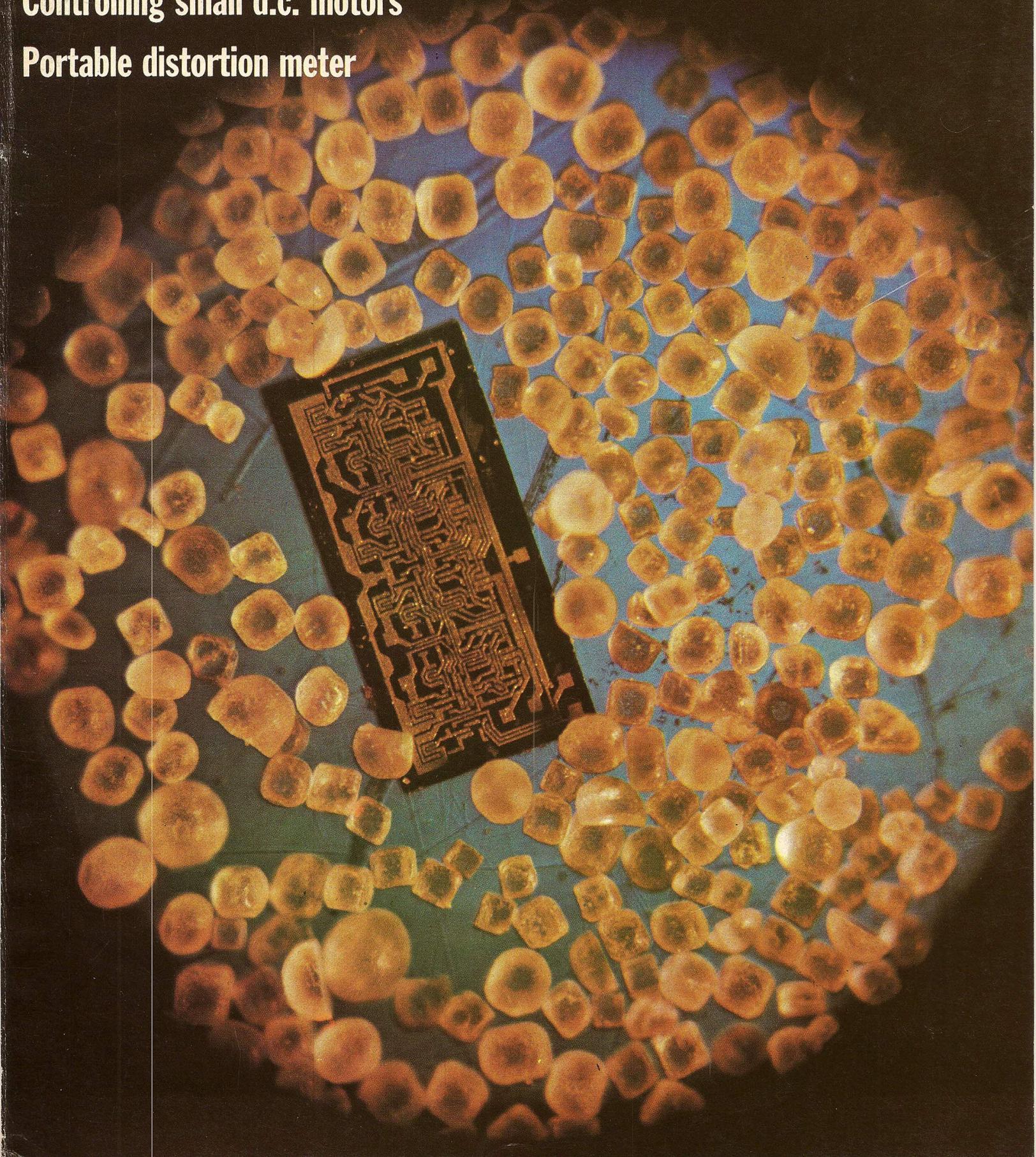


Wireless World

July 1972 20p

Controlling small d.c. motors

Portable distortion meter



Wireless World

Appreciation; apology; appeal

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First a word of appreciation. In 50% of the copies of last month's issue a questionnaire was inserted. We invited readers to answer a number of questions, to comment on the contents, and to suggest ways in which the journal could be improved — if it was not to their liking. We have received a very encouraging response from readers and we are grateful for this ready co-operation. The answers are now being analysed. We are especially grateful to those readers who went to some length to justify their criticisms or substantiate their demands. One wrote a four-page letter enclosing several contributions to the section he would particularly like to see expanded!

Secondly a word of apology. In last month's editorial on electronics and telephone exchanges we stated that the TXE4 exchange recently ordered by the Post Office is of American manufacture. This is not so. It is produced in this country by S.T.C. a subsidiary of the American I.T.T. We also apologize for the delay in publishing recent issues due to production problems.

Now, a word about symbols. Logic symbols are the elements of the new language electronics engineers have had to learn during the past decade. As with all languages, however, if it is to be meaningful it must be adaptable and therefore prepared to embrace new ideas. When, however, a new symbol is introduced for an already well established one there will be many who will be unable, or unwilling, to see the justification for such a change. A case in point is the recent adoption by the British Standards Institution of yet another almost featureless rectangle instead of the meaningful half-circle for a gate. A few years ago we were using a circle as the gate symbol, with arrows indicating the input and output. This circle became a half-circle — the semi-circle indicating the direction of flow. The half-circle was adopted as a British Standard and therefore gave, as has been the BSI's policy, formal ratification to what had become accepted practice. Many will ask, therefore, why this further change?

While we agree that, if graphical symbols are to provide an international language they must be standardized, we question the wisdom of this latest change. We understand that the U.K. was democratically out-voted at the meeting of the International Electrotechnical Commission — as it was with the B.S.I. adoption of the rectangle instead of the familiar zig-zag for a resistor.

With our international readership we must, of course, use the symbols which will be understood by the majority of our readers, nearly 25% of whom are overseas. Some B.S. recommendations we have rigidly refused to adopt — as for instance the cross-over in a circuit diagram instead of the loop-over which we have used throughout our 60 odd years. Our justification for doing this is that it is all too easy for ink blots to be introduced in the printing process and for one of them to come at a cross-over and thereby make it appear as a connection.

Another symbol which we have not adopted is the rectangle for a resistor. We, rightly or wrongly, consider that the zig-zag, which is allowed by both the I.E.C. and B.S.I. as an alternative, is more meaningful and cannot be misunderstood even by those who may now have become accustomed to the ubiquitous rectangle.

What of this latest change in the symbol for a gate? Our contributor James Franklin has used it in this month's article on logic gates (p. 350). We have used it in the letters section (p.316), although we consider it less distinctive than the one we have been using lately. This will enable readers to make their own assessment of the symbol and we shall be glad to have their reactions.

Portable Distortion Monitor

A sensitive design based on a Wien null network for t.h.d. measurements

by J. L. Linsley Hood

One of the most useful instruments in the workshop of the experimenter with an interest in audio reproduction systems is a distortion meter. It enables a rapid assessment to be made of the performance of a given piece of audio equipment. Audio amplifiers, employing class B output stages, can generate audibly objectionable amounts of crossover distortion if they are out of adjustment, and yet the distortion produced can still be too small for it to be seen on an oscilloscope trace of a sine wave. If a distortion meter, as well as an oscilloscope, is available it is very much simpler to readjust the quiescent current levels until the objectionable crossover discontinuities (visible as step waveforms or spikes) are minimized, provided that the design of the amplifier is such that this can be done within the permitted output stage dissipation levels.

However, this is not the only use to which such an instrument can be put. There are a number of other circumstances in which beneficial adjustments can be made from the knowledge of the distortion content of a given signal. For example, the positioning of a gramophone pickup cartridge in relation to the record, and the optimization of playing and bias weights, can be facilitated by the measurement of the output signal (from the main amplifier or pre-amplifier) on the replay of a nominally constant-tone track. This measurement is made more difficult, inevitably, by the eccentricity of the disc, the "wow" and "flutter" if present in the turntable mechanism, and one suspects by the speed irregularities in the original tape recording. However, in spite of this, it is still possible to determine the relative best of a series of possible adjustments.

Similarly, the meter can be used to optimize the alignment of the ratio detector in an f.m. tuner when a continuous test tone is broadcast, or to check the constancy of speed of a tape recorder or turntable.

The equipment to be described can be employed, in the range down to about 1% t.h.d., essentially as a single knob instrument. The general principle of the measurement of harmonic distortion has been covered by a number of excellent papers, of which that due to Crawford¹ is particularly helpful, and draws attention to the need for a true r.m.s. indicating meter for accuracy in measurement. However, in the particular field of audio amplifier systems in which

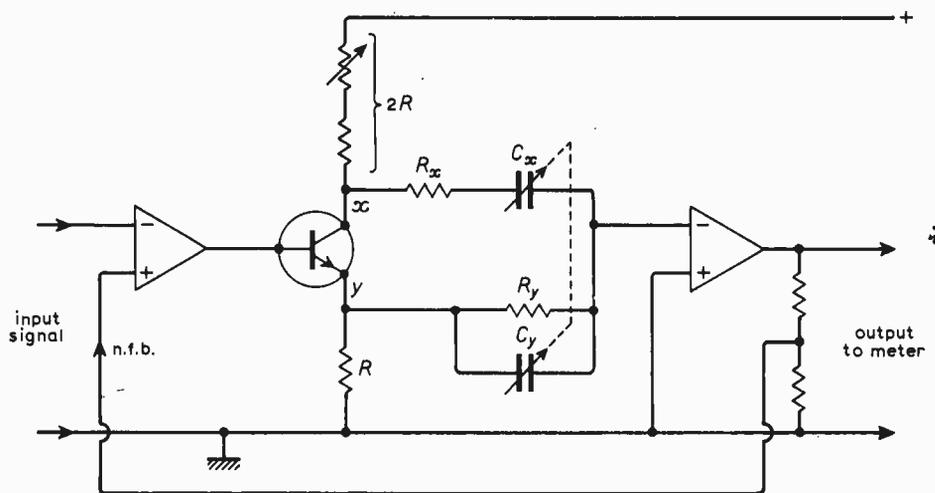


Fig. 1. Schematic arrangement of the distortion measuring system based on the Wien network.

residual distortion waveforms may be spiky, with a high peak amplitude-to-duration time ratio, the r.m.s. magnitude of the waveform may be less important than its form or peak amplitude. For this reason, among others, the choice of the display metering technique is left to the discretion of the user, with the recommendation in the case of measurements on audio amplifiers that the nature of the residual waveform should be studied on an oscilloscope as an informative exercise.

Measurement method

In principle, any circuit which can give zero transmission at some chosen test frequency could be used as the basis for a total harmonic distortion meter, since by removing the fundamental frequency of the signal only the harmonics—together with hum and circuit noise—would remain. However, in practice, it is desirable that there should not be too many knobs to adjust in producing the required null and, ideally, these knobs should be independent of one another in their operation.

Two techniques for producing such a notch in the frequency passband are commonly employed for this purpose; the parallel T and Wien networks, of which the former is the more popular, and several circuits using it have been described^{2,3}. Unfortunately, although the parallel T circuit gives a good null, it is necessary to

adjust three separate controls, which are somewhat interdependent, and this makes measurements time consuming. On the other hand the Wien network can be made to give an equally good null, with only two independent controls. This allows much greater ease in use. Some overall negative feedback needs to be applied in order to sharpen up the notch by reducing the overall amplitude response of the circuit. A suitable system for applying negative feedback in this manner around a Wien network is shown schematically in Fig. 1.

As can be seen from the analysis of the operation of the Wien network given in the Appendix, a null is obtained when the voltage at x is twice that at y, and in anti-phase. If a very high beta device, such as a Darlington transistor, is employed as the phase splitter, this condition can be very nearly satisfied when the collector load of the phase splitter is equal to twice the emitter resistance. A small variable resistor in series with the collector load will allow the necessary trimming to be carried out, and will also allow compensation for errors in the values of R_x and R_y when these are switched to change the frequency range.

The null frequency in this arrangement is then adjustable by the ganged capacitor C_x, C_y . (Although this is shown as a 470pF twin gang, a 300pF unit was employed in one of two prototypes with an appropriate increase in the values of the switched fixed

resistors, to make the RC product the same.) The trimmer capacitors across the gang sections should be retained so that the null at the low capacitance end of the swing can be set, by adjustment of these, to coincide with that obtained by the adjustment of the phase splitter load trimming resistor (Fig. 1) at the maximum capacitance end of the swing. If these adjustments are made correctly, and the values of the switched resistors are reasonably precise, the notch frequency adjustment may be made over a fairly wide frequency range without the need for alteration of the phase trimmer resistor control.

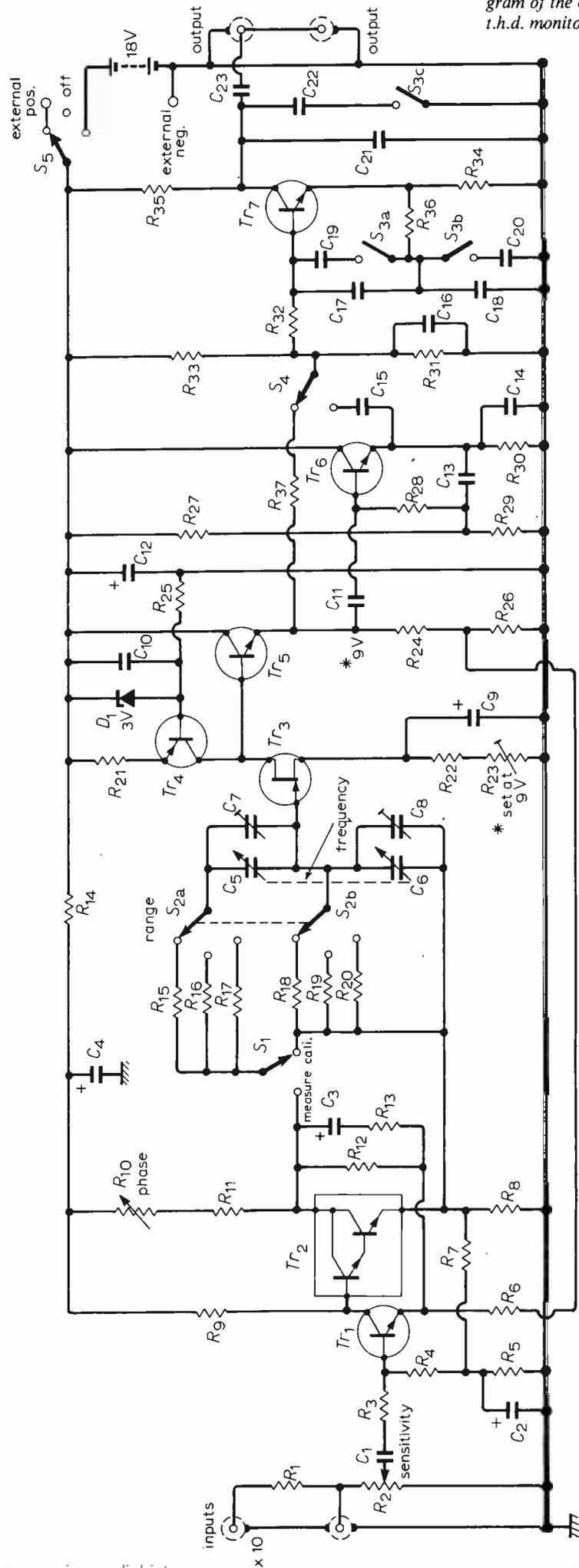
The sharpness of the null point can be determined by the amount of negative feedback applied around the loop. However, there is a practical limit to which this could or should be sharpened because the output from the phase splitter and its preceding amplifier will increase on approaching the null point, as the negative feedback is reduced, and this could cause the input stages to distort the signal. Also, the sharper the null which is generated, the more difficult it will become to set the null point on the frequency adjustment control.

Because of the high transfer impedance of the bridge system, particularly at the lowest frequency range, it is desirable that the following stage should have a high input impedance. The final circuit employed is shown in Fig. 2.

A high gain n-p-n small-signal transistor is employed as the input amplifying stage under conditions which give low circuit noise and distortion. Negative feedback is applied, from the collector of Tr_2 to the emitter of Tr_1 , to further reduce the distortion introduced by this stage. It will be appreciated that the harmonic distortion arising in this part of the circuit is indistinguishable from that of the signal being measured, and it is therefore essential to ensure that this is well below the lower limit of the measurement range. The harmonic distortion due to Tr_1 and Tr_2 , with the values shown, is less than 0.002%. This does not introduce a significant error at the lowest level to which this system is capable of measuring—of the order of 0.01% t.h.d. at 1kHz.

In practice, the limiting factor in such a measurement is likely to be the circuit noise introduced by the signal source, the equipment under test, or the t.h.d. meter stages prior to the null network. In order to reduce the extent to which the meter imposes its own lower limit the components in this part of the circuit, and in particular the transistors Tr_1 , Tr_2 and Tr_3 , should be chosen with care. In the two models of this equipment built so far the wideband noise figure due to the meter amplifier stages was approximately equivalent to 0.004% t.h.d. in one case and to 0.007% in the other. The use of subsequent passband limiting filter circuitry allows an improvement in the effective signal-to-noise ratio and with low noise sources measurements of the order of 0.01% are practicable. Because of the ability of the eye to recognize persistent patterns even when these are modified by noise, distortion components well below this figure can be seen, and correctly identi-

Fig. 2. Circuit diagram of the complete t.h.d. monitor.



fied in harmonic content, on a well synchronized oscilloscope trace.

Direct current negative feedback is applied to the emitter and base circuits of Tr_1 to stabilize the d.c. working point of the system, and the emitter voltage of Tr_2 provides the gate potential of the f.e.t. Tr_3 . The overall d.c. operating conditions are set by the adjustment of the potentiometer in the source circuit of Tr_3 , and this is set so that the voltage at the emitter of Tr_5 is approximately 9V.

For the reasons mentioned above an f.e.t. is used as the amplifier stage following the Wien network, and in order to obtain a high stage gain this is employed with a bipolar transistor constant-current load and an output emitter follower as an impedance converter stage. Overall a.c. and d.c. negative feedback is taken from a tap on the emitter load of Tr_5 , and the output signal is taken from the emitter of Tr_5 through a high-pass/low-pass filter which gives an overall gain of approximately $\times 2$. In a simple system the output could be taken directly from the output of Tr_5 . This would not make the best use of the equipment since although it is possible to design the t.h.d. meter system so that it does not introduce any significant amount of mains induced hum voltage, it is difficult to avoid some hum pick-up in practical measurements on amplifiers. It is therefore almost essential to have some optional 50Hz filtering if low orders of harmonic distortion are to be measured. The same is true in respect of high-frequency noise components since even the best amplifiers and signal generators introduce some noise, which must be removed, if possible without alteration of the harmonic structure of the distortion components.

Operation

No indicating system is included in this monitor as it is intended that it shall be used in conjunction with some suitable high-input impedance measuring device such as a millivoltmeter. If one of these is to be built, as for example the excellent design due to Waddington⁴, this would be more versatile as a separate unit.

For the purposes of measurement, it is preferred that the display instrument should possess full scale ranges of 3V, 0.3V and 30mV r.m.s. Measurement is carried out in the following manner.

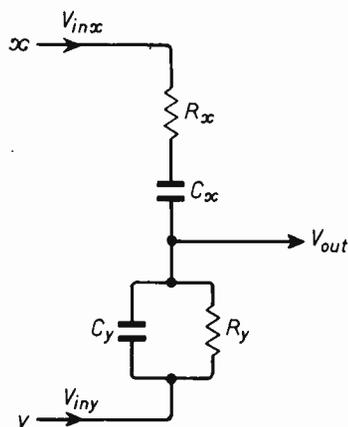


Fig. 3. Circuit of the Wien null network.

Components list

Resistors

1 - 900k	19 - 1.5M, 2%
2 - 100k, lin.	20 - 15M, 2%
3 - 22k	21 - 4.7k
4 - 47k	22 - 4.7k
5 - 15k, 5%	23 - 5k, preset
6 - 3.3k, 5%	24 - 3.3k, 5%
7 - 22k, 5%	25 - 15k
8 - 1.8k, 5%	26 - 82 Ω , 5%
9 - 470k	27 - 68k, 5%
10 - 2.5k, lin. (10 turns)	29 - 39k, 5%
11 - 3.3k, 5%	29 - 47k, 5%
12 - 120k	30 - 2.7k
13 - 39k, 5%	31 - 68k, 5%
14 - 1k	32 - 33k, 5%
15 - 150k, 2%	22 - 220k, 5%
16 - 1.5M, 2%	34 - 1k
17 - 15M, 2%	35 - 1.8k
18 - 150k, 2%	36 - 3.3k, 5%
	37 - 470 Ω

Capacitors

1 - 470n	13 - 100n
2 - 100 μ /4V	14 - 1n
3 - 22 μ /20V	15 - 10n
4 - 250 μ /25V	16 - 120p
5, 6 - 470p, ganged	17 - 68p
7 - trimmer	18 - 330p
8 - trimmer	19 - 680p
9 - 320 μ /6.4V	20 - 3.3n
10 - 100n	21 - 680p
11 - 4.7n	22 - 6.8n
12 - 250 μ /25V	23 - 220n

Diode

1 - 3V, zener

Transistors

1 - BC109 (C)	4 - 2N4058
2 - MPSA14	5, 6, 7 - BC109
3 - 2N4302	

Select a frequency range on the t.h.d. meter appropriate to the frequency of the signal being measured, and set the calibrate/measure switch to the calibrate position. Then adjust the input sensitivity control to give a 3V r.m.s. output to the millivoltmeter and set the switch to the measure position. The frequency and phase controls are then adjusted to give the best null possible (this is much easier to do if the output is viewed on a 'scope because of its very rapid response). The millivoltmeter is then switched to the 0.3V or the 30mV positions which would be equivalent to full scale readings of 10% and 1% t.h.d. If a 3mV f.s.d. scale is available on the meter, this would be equivalent to a range of 0.1% f.s.d. and the general wideband background noise of the instrument, with an open circuit input, should be less than 10% of this lowest range, reducing as the bandwidth is reduced.

In order to facilitate connection to an oscilloscope, as a guide to the specific shortcomings of the apparatus or generator under test, it is suggested that the output sockets should be duplicated. It will also be found convenient in practice to synchronize the oscilloscope from the input rather than the output waveform in order to avoid loss of timebase synchronism as the null is reached. A 'scope is also helpful in identifying the type of residual null point distortion.

In spite of the practical snags of t.h.d. measurements, because signals are contaminated with hum and residual noise, it is still much easier and quicker to perform

than the alternative method using an harmonic analyzer, which avoids these interfering signals.

Construction

No difficulty should be encountered in the construction of the meter provided that sensible precautions are observed; such as keeping the input leads within the instrument as short as possible and well away from any output circuitry. However, it is possible that the amplifier could become unstable if too high a value of collector load is employed for Tr_2 . It is advisable therefore that the trimmer potentiometer in this position (the phase control) should be set to its lowest value on initial commissioning, and then adjusted upward to the necessary operating value.

It is helpful if the whole instrument is enclosed in an electrically conducting case (a die cast box is excellent) and that the case is connected to the earth line of the instrument at the input socket only. The user can make his own arrangements for mounting the twin gang tuning capacitor and the eventual calibration of the scale of this in notch frequency, but it should be remembered that the frame and spindle of the capacitor is live and at a very high impedance point which makes this very prone to hum pick-up if the spindle protrudes through the case. A slow-motion drive to this capacitor is essential, and at least a 100:1 drive should be employed. To facilitate the isolation of the spindle a combination of epicyclic and cord drive was employed in the prototypes, with an insulated spindle coupler joining the capacitor shaft to the earthed pointer assembly.

Appendix

If an input signal is applied, in antiphase, to the opposite ends of the Wien network shown in Fig. 3, and if the magnitude of the signal applied at x is twice that applied at y , there will be a node when the impedance of the upper limb is twice that of the lower, i.e. when

$$R_x + (1/j\omega C_x) = 2 \left[\frac{R_y/j\omega C_y}{R_y + (1/j\omega C_y)} \right]$$

If $R_x = R_y$ and $C_x = C_y$ this simplifies to $[R + (1/j\omega C)] \cdot [R + (1/j\omega C)] = 2R/j\omega C$ (1)
 $\therefore R^2 - (1/\omega^2 C^2) = 0$.

The frequency at which this occurs is when

$$\omega = 1/RC \quad \text{or} \quad f = 1/2\pi RC.$$

In the more general case this is when

$$f = 1/2\pi\sqrt{R_x R_y C_x C_y}.$$

The fact that the 'imaginary' terms in (1) cancel out implies that the node is in fact a null, with no quadrature components.

References

1. Crawford, E. C., *Marconi Instrumentation*, vol. 9, no. 4, pp. 86-90.
2. Haigh, L., *Wireless World*, July 1969, pp. 317-320.
3. Bailey, A. R., *Electronic Technology*, Feb. 1960, p. 66.
4. Waddington, D. E. O'N., *Wireless World*, March 1966, pp. 111-113.

Audio Pre-amplifier using Operational Amplifier Techniques

by Daniel Meyer

It has been obvious for some time that the level of performance that can be obtained from the common two- or three-transistor pre-amplifier circuit is no longer adequate in systems of the highest quality. In a typical circuit of the type shown in Fig. 1, the open-loop gain will be a maximum of 60 or possibly 70dB. If the closed-loop gain is set at a reasonable value, say 30dB, the distortion is rather high at the lower frequencies where equalization requirements make considerable bass boost necessary. At 30Hz over 18dB of boost is required and as a result the circuit has only 10 to 15dB of negative feedback remaining. This can quite easily result in distortion in the range of 1% or more, which cannot be tolerated any longer with better grade systems.

The solution is a pre-amplifier circuit with better linearity, or more feedback; which requires a higher open-loop gain to begin with. There are many ways to approach this design problem, but one of the most interesting and rewarding seems to be to approach the circuit as a special operational amplifier. This has been done in the Motorola MC1303 integrated circuit with generally good results. Performance is more than good enough in most respects, but its noise and allowable output loading are not up to the best standards. Both characteristics are due to problems inherent in integrated-circuit fabrication. Noise tends to increase as the number of processing steps increase and it is no problem at all to find individual transistors that will give better noise figures than those on the chip.

Integrated-circuit designers seem to consider the amplifier's performance as a d.c. amplifier more important than its characteristics that would be of interest in audio-frequency circuits. Characteristics such as input bias current, input offset current and voltage, while important in a d.c. amplifier, are of little importance in an audio amplifier. In an audio system we are interested primarily in the following characteristics

- distortion generated by the circuit when used as an equalized pre-amplifier
- noise generated by the input amplifier
- dynamic range, or maximum input level
- input and output impedances.

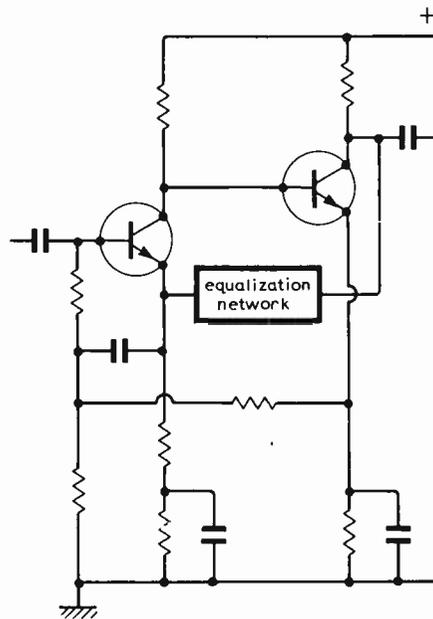


Fig. 1. With conventional pre-amplifiers of the type shown, equalization results in relatively high distortion at low frequencies — R.I.A.A. correction of 18dB boost at 30Hz means there is relatively little negative feedback at 1.f.

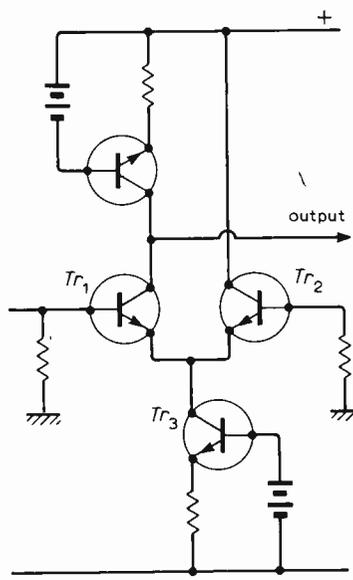


Fig. 2. Better linearity means more feedback means more open-loop gain. Use of a 'dynamic' current source as a collector load increases gain.

As the intended use is as an audio-frequency amplifier, input matching becomes unimportant. The amplifier will have unity gain at d.c. under closed-loop conditions and any small offset that may exist in the output due to mismatch in the input voltage and/or currents will be unimportant. We also do not have to be concerned with compensation that will produce a stable amplifier with gains all the way down to unity. The circuit can be designed with the knowledge that midband gain will be around 30 to 40dB.

Designing the circuit as a special-purpose operational amplifier gives us two useful characteristics. First we find that all signal input and outputs are at d.c. earth potential. This is very handy in designing a complete system because it is no longer necessary to worry about the effects of coupling capacitors charging and discharging when switches are thrown, or potentiometer wiper positions are changed. Second we have a circuit whose gain is set by the values of two resistors in the feedback loop, so performance is consistent despite variations in component characteristics.

Design details

The input stage, in common with almost all operational amplifiers, is a differential amplifier whose emitters are fed from a current source. The current source is normally included to increase common-mode rejection, but in this case it is only used to increase the circuit's isolation from the negative supply rail. The input signal in Fig. 2 sees the collector circuit of the current-source transistor Tr_3 as a high-impedance path back to the negative supply. As a result rejection of ripple or voltage fluctuations is improved considerably. Transistors Tr_1 and Tr_2 split the current from the current source between them. The standing current is chosen to give an optimum noise figure for the transistor being used, in this case 100 μ A.

To obtain the lowest possible distortion from the stage it must have as high a gain as possible. As has been pointed out by J. L. Linsley Hood* and others, the inherent non-linearity of the bipolar transistor is reduced as gain is increased. Maximum gain is obtained by using a

* "The Liniac", *Wireless World*, Sept. 1971 vol. 77 pp. 437-41.

dynamic load in the collector circuit. Such a load has substantially higher dynamic impedance than its d.c. resistance. A dynamic load may consist of a bootstrapped load resistor, an "active" load, or a current source. The current source is the least troublesome and is used in Fig. 3. The current-source stage Tr_3 is biased by the two 1N914 diodes and the current that it will pass is controlled by the value of the emitter resistor. Provided that the base bias voltage is stable the dynamic resistance will be proportional to the reciprocal of the slope of the collector current-base current characteristic of the transistor. The flatter the slope, the higher the impedance.

A fixed resistor is added in series with the current-source transistor's collector to isolate the collector-base capacitance from the signal path at high frequencies. Due to the very high dynamic impedance

presented by the current-source and the collector of the amplifier, a relatively small amount of capacitance can cause a pole in the amplifier response within the audio range. To prevent loading of the input stage collector circuit and loss of gain, an emitter follower is used as a matching device between stages. The emitter follower (Tr_4 in Fig. 3) drives the base circuit of the second amplifier stage Tr_5 . A current-source collector load is again used for maximum gain and best possible linearity.

Output impedance of this circuit is quite high in its open-loop state, but again unlike a general-purpose operational amplifier, this is not important. The circuit will always be used in a closed-loop system and as the feedback is taken from this point, the apparent impedance will be reduced by a factor proportional to the amount of feedback. In this par-

ticular case the open-loop gain will be around 110dB and the output impedance around $100k\Omega$. As the desired closed-loop gain will be 30 to 40dB, the output impedance will be reduced to a few hundred ohms. The open-loop gain of Tr_1 is about 68dB and that of Tr_5 , 45dB.

Maximum output voltage depends on the supply voltage used and the loading that is applied to the output of the circuit. With a +15 and -15V supply and a 10-k Ω load at the output, a maximum of 7V r.m.s. can be obtained before overloading or clipping occurs. In a circuit of this type the load impedance across which you can produce full output is a direct function of the value of the emitter resistors of Tr_5 and current source Tr_6 , which control the standing current of this stage. This is interesting because it can be used as the basis for an unusual type of class A power amplifier. Fig. 4 shows the maximum output that can be obtained with various values of loading using the specified supply voltages. Supply voltages can be increased to as much as +25 and -25V with the transistors specified. Maximum input level will be a direct function of supply voltage given a constant closed-loop gain, so as high a supply voltage as possible should be used if this is an important specification in the system.

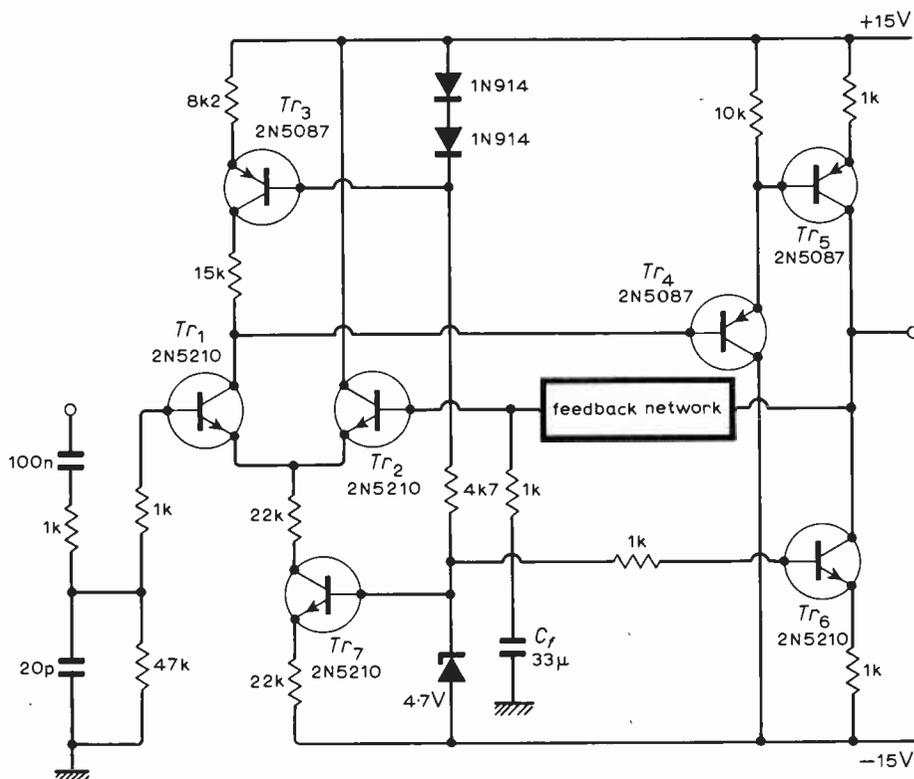


Fig. 3. A current source load, Tr_6 , is also used for the second amplifier transistor Tr_5 . Tr_1 provides 68dB of gain and Tr_5 , 45dB.

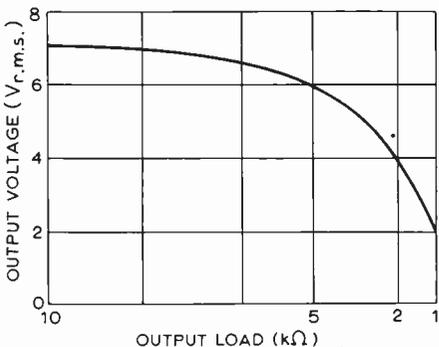


Fig. 4. Load impedance across which full output can be produced is a function of Tr_5 and Tr_6 emitter resistors which control standing current (assumes 15-V supplies).

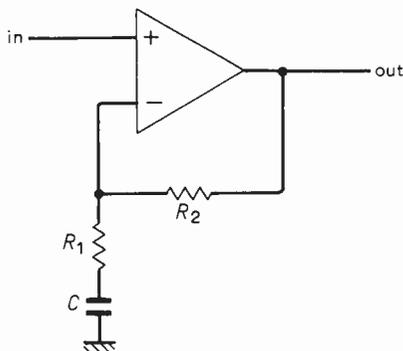


Fig. 5. Overall gain is $(R_1 + R_2)/R_1$. Resistor R_1 is 1k Ω and R_2 is in the feedback network — Fig. 6.

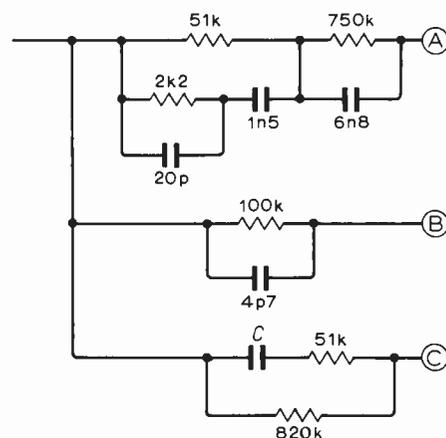
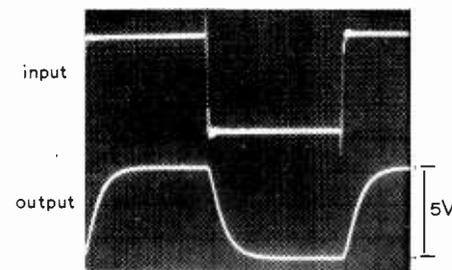


Fig. 6. Equalizing networks for Fig. 3 circuit. Values are for magnetic disc pickup A, dynamic microphone B, and tape head C. Capacitor C should be 910pF for 7½in/sec and 1.5nF for 3½in/sec. Component tolerances should be 5% or better.



110-kHz square-wave response of circuit using network B of Fig. 6. Sweep: 1 μ s/cm.

Fig. 7. Suggested circuit for pre-amplifier with Baxandall-type tone controls. Distortion at maximum lift is better than 0.01% at normal levels.

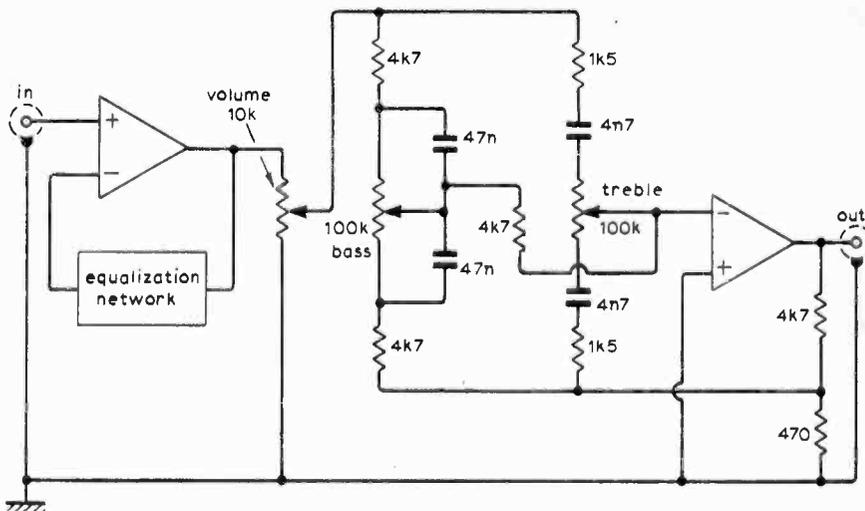
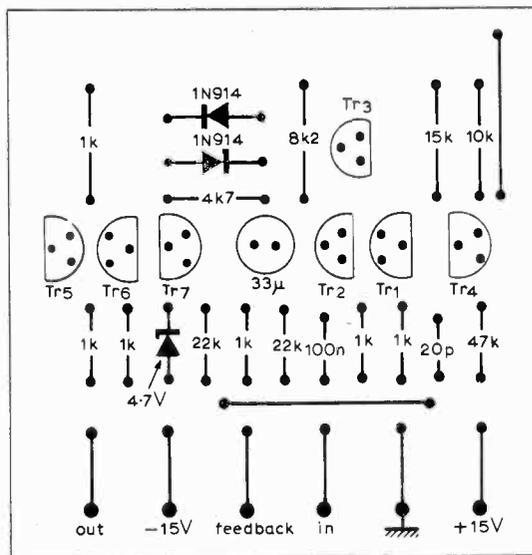
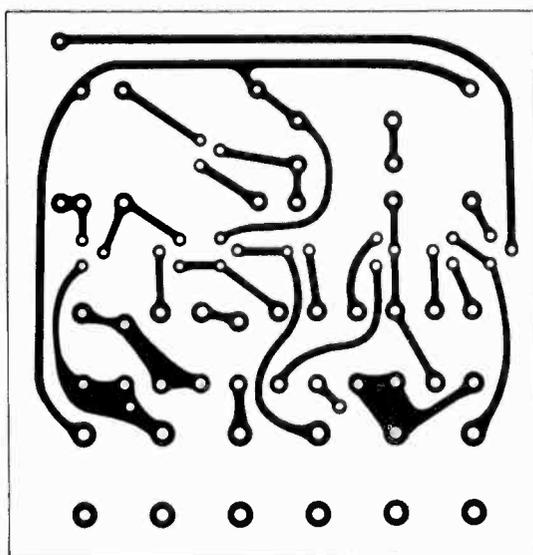


Fig. 8. Suitable printed board layout. Feedback components can be mounted on switch contacts. Components are shown mounted on the side opposite to the conductors.



Feedback networks

When a signal is applied to the non-inverting input of an operational amplifier as shown in Fig. 5, the gain is $(R_1 + R_2)/R_1$. The capacitor in series with R_1 reduces the gain to unity at d.c. Due to the large amount of feedback used, the input impedance of the circuit is well over $1M\Omega$ and a resistor must be added in parallel with the input of the circuit to properly load a magnetic cartridge, the most common value of $47k\Omega$ being shown. The small capacitor in parallel with this resistor rolls off the response at frequencies over 200kHz, ensuring that the generator impedance as seen by the circuit cannot become infinite at high frequencies.

Networks needed for R.I.A.A., microphone (flat), and tape head equalization are shown in Fig. 6. The 2.2-k Ω resistor and 20pF capacitor in the magnetic (disc) network shape the response past 50kHz. This is not normally necessary in a lower-gain, or narrow-bandwidth circuit, but is essential here if optimum transient response at higher frequencies is to be obtained. The photo shows the response of the circuit to a 100-kHz square-wave input using the flat microphone network. Rise time will be reduced if a high-

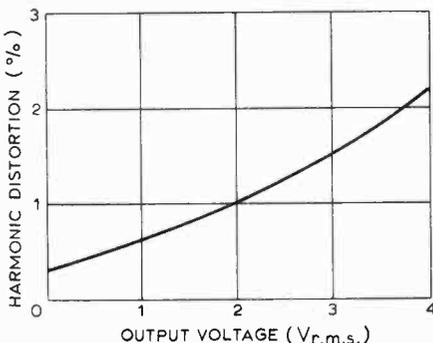
impedance signal source is used, but in no case will it be greater than 2.5 μ s.

Complete circuit

Fig. 7 shows a pre-amplifier channel with tone controls using the circuit. The circuit is an excellent driver for a Baxandall-type tone control system, the arrangement shown giving approximately 20dB of gain, and bass and treble lift and cut of approximately 20dB at 30Hz and 20kHz with respect to 1kHz. Maximum-lift distortion of this circuit, which is the worst case condition, is better than 0.01%

Performance

harmonic distortion	< 0.01% at 1-V output, 20Hz to 20kHz
hum and noise	> 80dB below 10mV input
maximum 'undistorted' output	7V r.m.s. with 15-V supplies and 10k Ω load
maximum input	100mV with 'magnetic' network
input impedance	< 1M Ω at 40-dB gain
output impedance	> 1k Ω at 40-dB gain
open-loop gain	110dB
supply voltage	+ 10 to + 25 and - 10 to - 25V
drain	10mA at 15V



Output voltage vs harmonic distortion before feedback is applied.

at normal output levels — considerably better than the widely used single-transistor driver circuit under the same maximum lift conditions.

The number and types of inputs is left to the builder. High-level inputs may be switched directly into the volume control, or reduced in level by a suitable resistive divider and applied to the pre-amplifier input. This seems to do little harm with a circuit of this quality and makes switching far simpler. In this way only two switch decks are required, one to

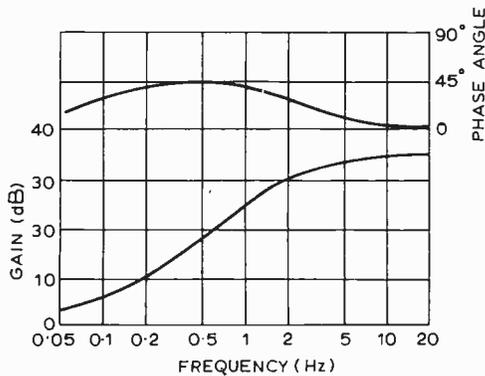
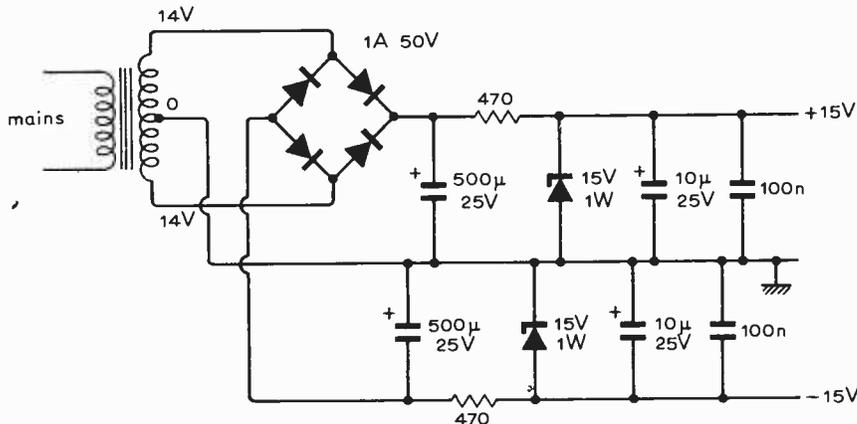


Fig. 9. Effect of l.f. roll-off capacitor C_f

Fig. 10. With this power supply, hum is 80dB below 10mV input.



switch the inputs and one to switch the equalization networks.

The circuit was designed as a plug-in unit on a small printed-circuit card. The circuit board pattern and parts locations are shown in Fig. 8. This type of construction makes screening of the circuit in a metal enclosure simple and also makes it easy to encapsulate the completed circuit in resin if desired. Leads carrying low-level signals should be shielded if more than an inch or two long and grounding loops should be avoided. Resistors and capacitors in the equalization network can usually be mounted directly on the selector switch lugs. Components should be 5% or better tolerance in this portion of the circuit. The capacitor at the input was added to ensure that no direct voltage could be applied to the input transistors from an external source. If such is not likely in your application the capacitor should be removed. The low-frequency roll-off point is determined by the value of capacitor C_f in the base circuit of Tr_2 . As can be seen from Fig. 9, the gain falls off below 5Hz and approaches unity below 0.1Hz. Phase shift reaches a maximum of approximately 55° between 0.5 and 1.0Hz and then decreases again as the input frequency approaches zero.

A suitable power supply for the pre-amplifier is shown in Fig. 10. With this supply the pre-amp will have hum and noise typically better than 80dB below 10mV input. A complete pre-amplifier and control system using the gain modules as described will have a total harmonic distortion of less than 0.01% at any frequency and under any conditions below its overload point. The extremely wide

bandwidth and freedom from ringing or overload on transients makes the pre-amplifier very pleasant to listen to for extended periods. Combined with one of the excellent transistor power amplifiers now available and a first-rate speaker system, it produces a music system second to none. The fact that seven transistors are used makes little difference as they are relatively low-cost plastic package types.

Announcements

The Electronics Division of the Institution of Electrical Engineers, in association with the telecommunications group at the Hatfield Polytechnic, is to organize a two week residential vacation school on 'Signal Processing in Modern Telecommunication Systems'. The school will be held at the Polytechnic from 10th to 21st July. Further details are available from Divisional Secretary (Electronics), IEE, Savoy Place, London WC2R 0BL.

A series of nine lectures entitled "Digital Techniques in Television Engineering" will be given on Wednesday evenings at 19.00 commencing 25th October at the University College, Gower Street, London WC1. Further details may be obtained from Royal Television Society, 166 Shaftesbury Avenue, London WC2H 8JH. Fee £5.

Three educational 8 or 16mm sound films on colour television servicing and installation, produced by Zaar Films Ltd, Montauban Chambers, 339 Clifton Drive South, St. Annes-on-Sea, Lancs FY8 1LP are: Part 1 "The Colour Signal", 30 minutes, price £70, rental £10 per day; Part 2 "The Receiver Decoder", 20 minutes, price £65, rental £7.50 per day; Part 3 "Receiver Installation", 25 minutes, price £70, rental £10 per day.

A new company, Lecson Audio Ltd, has been formed at St. Ives, Huntingdon, to manufacture high-fidelity audio equipment. The managing director is David Southward, until recently of Cambridge Audio, and the two technical directors John Greenbank, who recently left the editorial staff of *Wireless World*, and Bob Stuart, from Marconi Instruments. The first product is to be a wide-range horn loudspeaker system.

An agreement has been reached between the EVR partners, Imperial Chemical Industries Ltd of Britain, CIBA-GEIGY of Switzerland, and a Japanese consortium, to exploit the electronic video recording system world wide.

An audio advisory service to guide management of hotels, etc, through the problems of installing the right sound equipment has been set up by Audio Fidelity Ltd, Audio House, 102 Henconner Lane, Leeds LS13 4LQ, the public address systems group which includes Fane Acoustics Ltd, Futuristic Aids Ltd, Linear Products Ltd and RSC Hi-Fi Centres.

Under a marketing agreement recently concluded the Bell & Howell Video Systems Division are appointed distributors of JVC-Nivico close-circuit TV equipment for the United Kingdom and the Republic of Ireland.

Piber International Ltd, the Spanish component manufacturers who recently set up a temporary office in London, have moved to a permanent establishment at Romar House, The Causeway, Staines, Middx. Tel: Staines 56154.

Prowest Electronics Ltd, of Maidenhead, have received an order worth £40,000 from Yleisradio AB, the Finnish broadcasting corporation, to supply studio mixers and vision switching equipment for a new regional network studio being constructed in Tampere, Finland.

EMI has been awarded a contract worth over £1M from the Ministry of Defence to design and manage the re-instrumentation of a guided weapons firing range in the Outer Hebrides.

An order worth £120,000 has been placed by the Union Defence Force in the Arabian Gulf for h.f., s.s.b. "manpack" transceivers, Syncal 30 and Comcal 30, manufactured by Racal-Mobical Ltd, 464 Basingstoke Road, Reading, Berks RG2 0RY.

An order, worth approximately £1M, has been placed by the Australian Department of Supply with Amalgamated Wireless (Australasia) for field artillery computing equipment, to be built by Marconi Space and Defence Systems.

BP Tanker Co. Ltd. has placed orders worth over £262,000 for Decca radars which will equip thirteen new ships and two existing ones with dual installations of the 16in True Motion TM1629 and 12in Relative Motion RM1226. Another thirteen existing vessels are to have the TM1629.

Sixty Years Ago

July 1912. A Professor Willis L. Moore, of the United States Weather Bureau, in a paper to the International Radiotelegraph Conference proposed that the leading nations of the world should co-operate in making daily observations of weather conditions using ships at sea and transmitting them to central points by wireless telegraphy.

The report in *The Marconigraph* described the proposal as follows: "Professor Moore's project is to have all ships send daily at noon (Greenwich time) a code message of two or three words, reporting their location and observations of the weather to central stations at different points. From this information the central stations could tell the exact position of storms and other meteorological disturbances, and send out a timely warning to ships within the danger zone". This was the present World Weather Watch Network in embryo.

In the same issue of *The Marconigraph* a whole page was devoted to the amazing fact that someone had managed to pack a complete radio installation in something as small as a five ton motor boat!

News of the Month

Relativity theory test

The National Aeronautics and Space Administration is planning an experiment to test the equivalence principle, the crux of Albert Einstein's general theory of relativity.

In the experiment, a clock will be launched on a Scout-D rocket on a 3.5hr elliptical trajectory over the Atlantic. During the flight this clock will always be in a weaker gravitational field than an identical clock at a Bermuda ground station. Hence, the frequency of the clock in the probe, observed by telemetry, will always appear to be greater than that of the clock on the ground. As the clock rises from the Earth through the increasingly weaker field to its maximum altitude of 18,000km it will appear to run increasingly faster.

Its rate will then progressively appear to decrease as it returns to the stronger field at lower altitudes. During the flight, the difference between the clock rate in the probe as indicated by the telemetry signal and that on the ground will be compared with the shifts predicted by Einstein's theory.

The Smithsonian Astrophysical Observatory will develop hydrogen maser clocks for the experiment. These clocks are the most stable ones available for the period of a few hours in which the experiment will be conducted.

Topographical data on U.K. for computer use

The electricity, gas and coal industries together comprise the largest user of private mobile radio communications in the U.K. They operate some 700 v.h.f. base stations, 25,000 mobile stations, 350 point-to-point u.h.f. links and a substantial number of microwave links. When a new radio station is set up a good deal of the initial work is concerned with calculating coverage and signal strength taking into account transmitter power, aerial height and the surrounding terrain.

As a result of work initiated by Central

Electricity Research Laboratories with the University of Manchester Institute of Science and Technology, the Joint Radio Committee of the Nationalised Power Industries has developed and put into operation a computer programme which predicts the field strength and service area for a given radio site. The Central Electricity Generating Board's IBM360 computer is provided with data relating to the transmitter map reference, the height above ground and effective radiated power of the transmitter aerial, the type of mobile receiver aerial and its height above ground, and the area to be surveyed. The computer output comprises a print-out of calculated field strengths and signal levels at half-kilometre intervals for the area surveyed, and in addition if needed, a transparency plot suitable for overlay on Ordnance Survey maps as required. The plot provides a clear and easily understood picture, when overlaid on the

appropriate map, of the comparative levels of signal over the area.

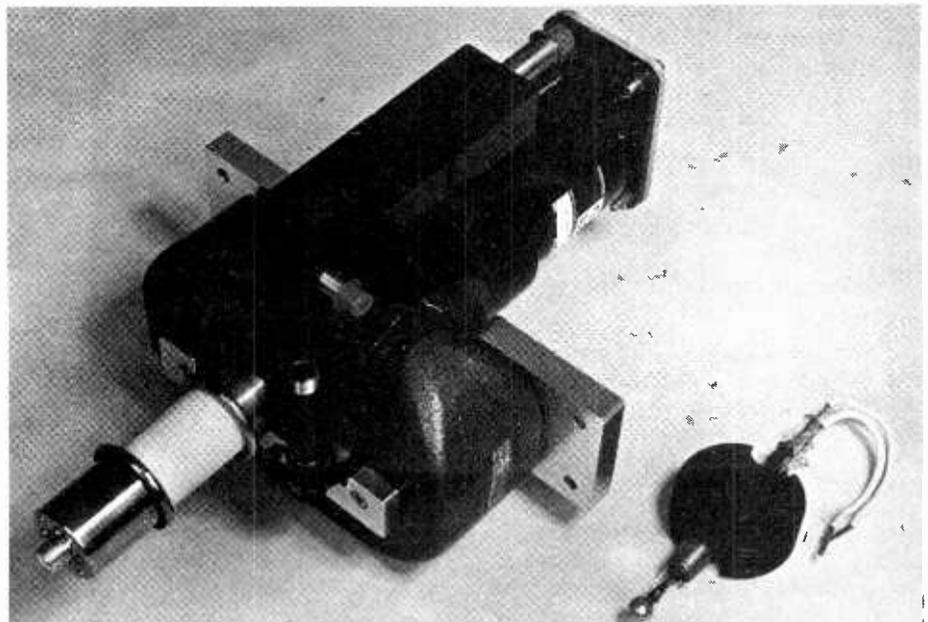
Since a knowledge of the topography of the area under survey is necessary in the calculation of radio field strengths at any given point, it has been necessary to create a data base of ground height. The work of extracting this information was carried out by PMA Consultants, of Horley, who assembled a data bank of 880,000 points at 0.5 kilometre intervals for the whole of England, Wales and Scotland (except the Highlands and Islands). The information was taken from the Ordnance Survey (1:25,000) maps, each point representing the main topographical feature in each 0.5km square.

This work was executed by a team of operators who were trained to examine map detail, and to represent as accurately as possible the information on a matrix. Each map was coded twice in this way by different operators, the results compared, and where differences occurred corrections made, and the results transferred to punched cards. Further checking processes were carried out after the cards were punched. The results have an accuracy of $\pm 3.8m$ as compared with the maps from which the information was extracted.

It is believed that this topographical data bank is the only one of its kind at present available in the United Kingdom, and other uses for it are envisaged. Copies of the data bank can be made available for use on other computers if required.

The computer method of calculating radio network coverage areas is now used by the nationalized power industries, and the service is available to other interested parties. The provision of these facilities in

It is twenty-five years since the English Electric Valve Company was founded from the wartime Marconi valve laboratories. In this time the company has expanded from employing 150 people with a working area of 50,000 sqft to the present total of more than 2,000 employees and almost half a million sqft. The photograph shows one of the company's latest 1MW tunable magnetrons on the left with one of the few remaining samples of a magnetron produced during the last war on the right.



other countries is to be made available. Enquiries for the U.K. service should be addressed to the Joint Radio Committee, c/o 30 Millbank, London SW1P 4RD, and for other countries (including Ireland) to PMA Consultants Ltd, Imperial Buildings, Victoria Road, Horley, Surrey.

Russian communication satellite

A Molniya 2 communications satellite was launched in the U.S.S.R. to extend the internal communications network. The satellite operates "in the centimetre waveband" and will carry telephone, radio and television signals. The orbit has a perigee of 460km in the southern hemisphere and an apogee of 39,300km in the northern hemisphere, inclined at 65.5° with a period of 11 hr 45min.

Hong Kong's electronic exports

Sales of Hong Kong made transistor radio sets to Britain rose 83% during last year to £4.7M. The increase was the largest in recent years. (World wide Hong Kong exports of transistor receivers increased 30% last year to £48.9M.) Hong Kong's electronics industry is second in importance to their textile industry, having 270 factories and a work force of 40,000. Last year, for the first time, more than half the value of their industry's total exports were in finished products rather than parts and components. Five years ago Hong Kong exports of tape recorders were worth under £13,700. Last year this rose to more than £412,000. In the supply of parts and components, Hong Kong's leading export line last year was parts for computers.

Ship-to-shore new v.h.f. station

The first of a new network of coast radio stations providing a v.h.f. radiotelephone service for shipping up to a range of 40 miles was brought into service on June 1st by the Post Office at Bacton, near Cromer, Norfolk. The station has a 40W transmitter. Over the next five years Post Office Telecommunications plans to build up to 13 of the new stations — and where possible aeriels will be mounted on existing radio masts on the coast, to avoid

spoiling the environment with extra structures.

The new network will greatly improve existing v.h.f. radiotelephone services and it will help to relieve pressure on the crowded medium-wave band which is used for ship-to-shore radiotelephone calls up to a range of 250 miles.

The present coast radio stations form a chain and are about 120 miles apart — with the exception of those in the far north of Scotland. With the new stations in service there will be a coast radio station every 60 miles or so.

Among the new ones to be built are stations in the Eastbourne area; at Start Point, South Devon; Ilfracombe, North Devon; Milford Haven; and the Clyde area. The others will follow as site details are settled. All the stations will be unmanned and controlled remotely by the nearest main coast radio station.

Communication through Bacton Radio is established on calling and answering frequency (Channel 16, 156.8MHz), after which the working frequencies are used (Channel 27, 161.95MHz Bacton transmit and 157.35MHz ship transmit).

Soldering aluminium alloy

Restrictions in the use of aluminium alloy because of difficulties in soldering it or in soldering other metals to it have been overcome by a simple technique which has been applied to the bonding of electrical connections. No special equipment fluxes, skills, or cleaning materials are needed. Standard cored solder, soldering irons and a small quantity of mineral lubricating oil (which acts as a deoxidant) is all that is required. The soldering procedure is as follows:

- Surfaces to be tinned are cleaned by filing or scraping.
- The cleaned surfaces are covered with a thin film of mineral oil and a further light clean given through the oil.
- The aluminium alloy is tinned through the oil film by the normal tinning procedure of rubbing the soldering bit over the surface. Care must be taken not to overheat.
- Excess oil is removed, and the tinned surfaces are degreased with a suitable degreasing agent.
- Bonding is then completed in the usual way by applying a non-corrosive flux, bringing the tinned surfaces together, and applying heat.

While the grade of lubricating oil used is relatively unimportant, it has been found that tinning is easier using the heavier oils. The procedure was described in "Techlink 1109" from the Department of Trade and Industry. Details are not available at present on the durability of joints under environmental conditions leading to electrolytic action or corrosion.

1973 television symposium

The 8th International Television Symposium is scheduled to be held in Montreux from the 18th to the 24th of May 1973. The convention building now under construction will be completed in time to house both the symposium and the associated television exhibition. The increased space available means that a larger number of exhibitors can be accommodated. The symposium will again be presided over by Prof. W. Gerber (Berne), who will be assisted by Prof. E. Borgnis (Zurich), Mr. H. Probst (Berne), and Mr. R. Jaussy (director of the Montreux Tourist Office).

Travelling scholarship

This year's winner of the £1,000 Pye travelling scholarship "for the most significant contribution during the year to the development of colour television" is C.B.B. Wood, now head of the Engineering Information Department of the B.B.C. "Woody", as he is affectionately known, was until recently head of the image scanning section in the Studio Group of the BBC Research Department. He joined the Research Department in 1946 and the award was made particularly for his work in achieving national, and subsequently international, standardization of a new set of primary colour phosphors for colour television systems. Mr. Wood, who received the award at the Royal Television Society Ball, is to undertake a tour of Australasia lecturing on colour television.

The B.B.C. celebrates 50 years

The B.B.C. will be fifty years old in November this year. To mark the occasion the B.B.C. is planning celebrations both on and off the air. One of the major public events will be an exhibition at the Langham (near Broadcasting House) which, using modern electronic techniques, will tell the story of broadcasting — a story which begins with "This is 2LO Calling" and develops beyond satellite television to the promise of the future. The exhibition, which will be free to the public, opens on 1st November.

A sister exhibition which demonstrates the engineering story of broadcasting from the improvisation of the early days to the sophisticated equipment of today — and tomorrow — will be opened the next day, 2nd November, at Mullard House, Torrington Place. The Post Office is considering demonstrating its role in broadcasting by joining the two exhibitions with a closed-circuit colour television link.

The celebrations mark the start of daily

broadcasting on 14th November 1922 from 2LO in London. The Birmingham and Manchester stations were brought into service the next day. The broadcast receiving licence was first introduced on 1st November 1922 and the fee was ten shillings. The British Broadcasting Company, which founded broadcasting in this country, was incorporated as the British Broadcasting Corporation on 1st January 1927.

The B.B.C. introduced its world-wide services in 1932 with the inauguration of the Empire Service in December of that year. Four years later, on 2nd November 1936, the B.B.C. introduced the world's first regular television service. On 1st July 1967, the first regular colour television system in Europe began on BBC-2.

Conferences and Exhibitions

LONDON

July 2-5 Imperial College
Environmental Acoustics Symposium

(Dr. R. W. B. Stephens, Dept. of Physics, Chelsea College of Science and Technology, Pulton Place, London S.W.6)

July 4-6 Bloomsbury Centre
Bio-Medical Engineering Exhibition
 (U.T.P. Exhibitions, 36-37 Furnial St. London EC4A 1JH)

July 25-27 Savoy Place
Advances in Marine Navigational Aids
 (Conference Dept., I.E.E., Savoy Place, London WC2R 0BL)

BRISTOL

July 11-13 Entertainment Centre
Wessex Electromation
 (Electromation Exhibitions, 344A Holdenhurst Rd. Bournemouth)

GUILDFORD

July 10-13 University of Surrey
Vacuum Instruments and Methods in Surface Studies
 (British Vacuum Council, 47 Belgrave Sq., London SW1X 8QX)

LEEDS

June 27-29 The University
Leeds Electronics Exhibition
 (C. S. Petch, Dept. of Electrical & Electronic Eng., The University, Leeds LS2 9JT)

PENARTH

July 22-29 Esplanade Shelter
B.A.E.C. Exhibition
 (Brit. Amateur Electronics Club, 26 Forrest Rd., Penarth, Glam.)

READING

July 19-21 The University
Defects in Semiconductors
 (Dr. J. E. Whitehouse, J. J. Thomson Physical Lab., University of Reading, Whiteknights, Reading RG6 2AF)

SHEFFIELD

July 3-6 The University
Testing in Electronics
 (Society of Electronic & Radio Technicians, 8-10 Charing Cross Rd, London WC2H 0HP)

SWANSEA

July 4-6 University College
Radio Receivers and Associated Systems
 (Conference Dept., I.E.R.E., 8-9 Bedford Sq., London WC1B 3RG)

Phase-locked-loop Stereo Decoder

A low-cost, high-performance design which does not use coils

A phase-locked-loop stereo decoder can be built with just four resistors, eight capacitors, one preset potentiometer, an indicator lamp and one integrated circuit. The circuit is given below. Briefly the i.c. works as follows. The output of a 76kHz RC oscillator (the frequency being set initially by R_4 , R_5 , C_7) is divided by four in two flip-flops to produce 19kHz. This frequency is multiplied by the incoming pilot tone to produce a d.c. component which is extracted by a low-pass filter (R_3 , C_6 , C_8) and used to control the frequency of the 76kHz oscillator. The output of the first divider (38kHz) is therefore phase-locked to the incoming pilot tone. The internally generated 19kHz is multiplied with the incoming composite signal in a second modulator and, when a pilot tone is present (stereo broadcast), its d.c. output, after low-pass filtering (C_4), operates a switch which lights the stereo beacon and allows the internally generated 38kHz to reach a switching decoder where the input signal is decoded to produce the left and right a.f. outputs. The components R_1 , R_2 , C_2 and C_3 are for output loading and de-emphasis (50 μ s).

The decoder requires a maximum input of 560mV r.m.s. and provides a stereo separation of 40dB from 50Hz to 15kHz. The audio output is 485mV r.m.s. with a 560mV input and the input impedance is 50k Ω . Other figures from the specification include a total harmonic distortion of 0.3%; 16mV pilot tone input for light-on;

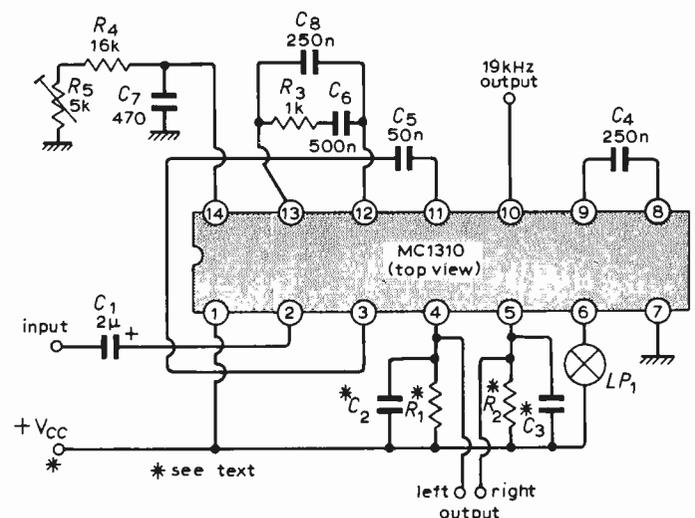
pilot (19kHz) rejection 34.4dB; sub-carrier (38kHz) rejection 45dB; and the capture range of the 76kHz oscillator is $\pm 3\%$.

The current taken by the beacon (LP_1) should not exceed 75mA (a light-emitting diode with a series resistor could be used). If it is desired to use a larger lamp a driver transistor must be employed. The values of R_1 , R_2 , C_2 and C_3 must be adjusted to suit the power supply voltage V_{CC} as follows: $R = R_1 = R_2$, $C = C_2 = C_3$, $V_{CC} = 8V$, $R = 2.7k\Omega$, $C = 18nF$; $V_{CC} = 9V$, $R = 3.9k\Omega$, $C = 13nF$; $V_{CC} = 12V$, $R = 6.2k\Omega$, $C = 8nF$.

To commission the decoder it is necessary to set the internal oscillator to 76kHz with no input signal by adjusting R_5 . The easy way to achieve this is with a high input impedance frequency meter. Failing this a stereo signal from a tuner can be applied to the decoder via a potentiometer (say 100k Ω). With maximum input adjust R_5 for a lock (indicated by the stereo beacon). Reduce the level of the input signal and again adjust R_5 . As the input signal is progressively reduced the correct setting for R_5 will become apparent.

The MC1310P stereo decoder i.c., which is manufactured by Motorola, is housed in a dual-in-line plastic package and can be obtained from Jermyn Industries Ltd, Vestry Estate, Sevenoaks, Kent, price £2.93 including p & p. Because this is a new device there might be a month or two delay before deliveries can commence.

Phase-locked-loop stereo decoder circuit using a single integrated circuit. The value of R_1 and R_2 is determined by the supply voltage and the time constants $R_1 C_2$ and $R_2 C_3$ should both be 50 μ s for the correct de-emphasis in the U.K.



Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Baxandall loudspeaker

Mr Magchielse (May issue, p.223) chooses the values for what is, after all, just a simple tuned bass-lifting circuit, by regarding the combination of this circuit and the loudspeaker as a fourth-order Tchebyshev filter. This approach I found interesting. (I assume his diagram is in the nature of a rough sketch only, as it does not show the expected equal-ripple frequency response nor is the overall curve in fact equal to the sum of the two component curves!)

I did consider the possibility of recommending an active bass-peaking circuit at the time the article on the low-cost loudspeaker* was being written, but after looking at some rather discouraging acoustic waveforms, rejected the idea for the following main reasons:

- The intermodulation distortion of the loudspeaker, already none too low on music with heavy bass components, is much increased.

- If the technique is used with a more powerful amplifier than Mr Magchielse's 10-watt one, the danger of thermally damaging the speech-coil on sustained heavy bass passages, such as sometimes occur in organ music, becomes quite serious. Many users of these loudspeakers feed them from typical present-day 20- or 30-watt amplifiers, and ordinarily experience no troubles even at quite high domestic listening levels. However, the addition of about 10dB of bass boost at 50Hz, though having little or no effect on the overall subjective volume impression, can increase several fold the power dissipated in the speech-coil on sustained heavy bass passages. It is the very fact that this occurs without a rise in the general volume level that is likely to catch the user out! (A recent measurement I made on the thermal properties of the Elac unit used, gave a speech-coil thermal time-constant of just over 5 seconds and a thermal resistance of 11°C/watt.)

With regard to intermodulation distortion, Fig. 1 shows some results obtained on one of the original models of the low-cost loudspeaker. The loudspeaker was fed simultaneously with

high level 50Hz sine-wave input and a much lower level (200mW) of 4kHz tone. The loudspeaker output was picked up on a microphone and fed via a simple third-order high-pass filter, with a cut-off frequency of 2kHz, to an oscilloscope. The filter removed all visible trace of 50Hz and harmonics, but permitted observation of the variation in 4kHz output amplitude caused mainly by the speech-coil moving in and out of the strongest region of the gap magnetic flux.

Investigations carried out by the B.B.C.¹ have shown that, on all normal musical items, even those considered to be particularly bassy, the peak levels do not actually occur at the lowest frequencies. When a bass-lifting circuit is not used, it seems that one is fairly safe in assuming that the level fed to the loudspeaker around 50Hz will never exceed 4W when using a 10-W amplifier — taken to mean one which will just give a sustained sine-wave mean output power of 10W into a 15-ohm resistance load without overloading. The intermodulation will then never be worse than that shown in Fig. 1(c), and only very occasionally will it be as bad as (b). However, with the

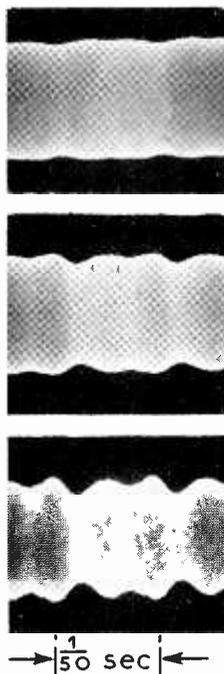


Fig. 1.

bass-lifting circuit in operation, intermodulation as bad as (c), or even worse, is likely to occur on music with large bass content, and would be expected to give "muddy" and generally displeasing sound quality.

Readers may well object that it is not the above "theoretical" arguments that count, but rather what the musical results actually sound like in practice. I therefore decided to try the scheme out using the same bass-lifting characteristic as that adopted by Mr Magchielse, though it was obtained in a rather cheaper and simpler way. This technique, which I have used a good deal during the past two years for obtaining high-pass and low-pass responses for active crossovers, is unusual and a brief description may be of interest. It is based on Sallen and Key filter ideas,² but uses the loudspeaker amplifier itself as the active element.

The most straightforward version of the scheme is that shown in Fig. 2(a), which is basically the same as Mr Magchielse's circuit, but the unity-gain feedback amplifier is here provided by the loudspeaker amplifier itself in association with an appropriate-ratio potential divider across the output. (The loudspeaker amplifier must, of course, be of the non-phase-inverting type. Most amplifiers — but not the Quad 303 — are like this.) The amplifier input impedance is assumed, for the moment, to be much greater than $2QR$, and the potential divider across the output is assumed to be of negligibly low resistance compared with $R/2Q$. If these conditions are not satisfied, the filter resistor values may, of course, be modified appropriately; it is even possible to eliminate one or both resistors by suitable choice of values in some cases.

What are the disadvantages of this technique, compared with the normal Sallen and Key arrangement as used by Mr Magchielse? Apart from a slight increase in the distortion of the loudspeaker amplifier in the region of the

(a)

0.5 W

(b)

1.5 W

(c)

4 W

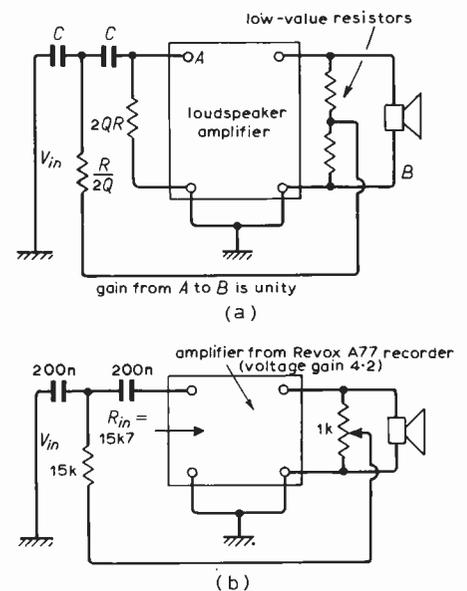


Fig. 2. Second-order high-pass active filter systems.

*"Low-cost High-quality Loudspeaker", P. J. Baxandall, *Wireless World*, August & September 1968.

cut-off frequency, which is negligible when a good amplifier is used, the only fundamental disadvantage is that the long-term stability of Q is poorer. This is because a good unity-gain feedback amplifier, as normally used, inherently possesses great stability of gain whereas, in the present scheme, the gain stability is dependent on the stability of ratio of the potential divider and the feedback resistors inside the amplifier. (A possible way to obtain a highly stable gain of unity from the loudspeaker amplifier is to feed the $R/2Q$ resistor from the emitter of the first stage. This, however, is not an ideal solution, because feed-forward, as well as feedback, is introduced; a signal is injected into the emitter and gives rise to an output from the amplifier not subject to the full filtering action. Though this may sometimes be made negligible by suitable choice of impedances, the method of using a potential divider across the output is preferred in practice, both because it avoids having to mess about with the inside of the amplifier and also because it leads to the ideas described below.)

Varying the ratio of the output potential divider in Fig. 2(a) varies the Q without altering the cut-off frequency, the latter being defined as the frequency of intersection of the horizontal and 40dB/decade (12dB/octave) asymptotes of the response curve. This introduces the convenient possibility of using equal-value resistors in the filter and obtaining whatever Q -value may be required by suitable adjustment of the divider ratio, which is best done experimentally. The cut-off frequency is simply given by $f_0 = 1/2\pi CR$, where C and R are the actual component values in the filter circuit.

The version used in the tests is shown in Fig. 2(b). From an a.c. viewpoint, the Revox amplifier has three 47k Ω resistors in parallel in its input circuit, and these, equivalent to 15.7k Ω , provide one R of the filter. It is important to set the potentiometer to a low setting before first switching on, as infinite Q , i.e. self-oscillation, is produced if the slider is raised too high!

If, as will often be the case, the loudspeaker amplifier gain is much greater than unity, say 10 or more, the long-term stability of gain of the amplifier plus potential divider will be just the same when set for a gain of nominal unity as when set for any other practical value. Indeed, for optimum overall stability of the filter characteristic, the unity-gain condition is no longer the best choice. A full analysis has been done, and checked experimentally. Some conclusions are:

(a) A unity-gain design gives minimum sensitivity of Q to variations in the actual filter R and C values, but high sensitivity to variations in the output potential divider ratio and/or amplifier internal feedback network ratio. The percentage variation in Q is $2Q^2$ times the percentage variation in either of these ratios.

(b) Minimum sensitivity of Q to variations in either of the above ratios is obtained when the right-hand filter resistor is twice the value of the left-hand one. The



Mr Baxandall's woofer.

factor corresponding to $2Q^2$ above is then $(2.828Q - 1)$.

(c) The convenient condition of equal resistor values gives very nearly as good a result; the factor is now $(3Q - 1)$.

(d) The sensitivity of Q to variations in the actual filter R and C values is much greater in (b) and (c), at high Q values, than in (a), but it is nevertheless considerably less than the sensitivity to variations in potential divider or feedback network ratios. Thus, taking all into account, the (b) and (c) designs represent a better balanced engineering choice.

In most crossover applications, the Q values required are in the region of unity, so that adequate stability is very easily obtained.

One final point. The circuit is liable to oscillate at a high frequency if the signal source, which has been assumed to have negligible internal impedance, is disconnected. This may be cured by interposing an input emitter follower, as in Mr Magchielse's diagram.

Subjective reactions to listening to music with the bass-lift circuit in operation were:

(1) It was quite successful in quiet bass passages. For example, it considerably improved the pizzicato double basses at the beginning of the second movement of Schubert's 'Unfinished' symphony.

(2) In any music with really substantial bass, such as the latter parts of

Saint-Saens's Symphony No. 3, where the organ augments the orchestra (Decca SXL6482), a very muddy, constricted and, to me, highly unpleasant sound quality was produced, leading to a desire to turn the volume down to an abnormally low level or, preferably, to cut out the bass-lift circuit.

(3) On changing over to my normal woofer, used in conjunction with the low-cost loudspeakers, on music such as in (2) above, the difference was really very great. The bass became unforced, warm and natural sounding, and one had the impression of plenty of overload capacity in hand even at a high volume setting.

Thus I think it is best to accept the bass limitations of the low-cost loudspeaker as inherent in the design, and not to try to force it to do what it cannot really do properly. Though many people seem to find the bass fairly satisfactory as it stands, the right way to improve it is certainly to add a woofer.

My woofer, which takes the form of a wide piano stool, is shown on its end (for clarity) in the photograph. It is a simple unlagged vented cabinet of $\frac{3}{4}$ in wood, with internal dimensions 31 x 17 x 11 $\frac{1}{2}$ in, containing a Goodmans unit. This unit is unfortunately not available on the retail market, but almost identical results are obtained using the KEF B139 unit. The vent is 4 x 4in and extends into the cabinet as a square pipe 6in long, giving a Helmholtz resonance frequency of approximately 35Hz. (This large-area vent has been found preferable to one of smaller area and depth, as there is much less "wind blowing" noise at high levels of bass output.) One cross-bracing member joins the centres of the larger sides. The shorter sides should protrude far enough to give a space of at least 1 $\frac{1}{2}$ in between the bottom of the cabinet and the floor.

The woofer is fed via a separate amplifier, with associated low-pass filter, from the sum of the two channels. The arrangement is as in Fig. 11 of my September 1968 article, except that, instead of using a simple shunt capacitor for attenuating the higher frequencies, the amplifier is made to give a proper second-order low-pass response ($Q \approx 0.7$, cut-off frequency 100Hz), by the technique described in this letter, but, of course, with the C s and R s interchanged. An amplifier with a maximum output of about 7 watts into 8 ohms has been found

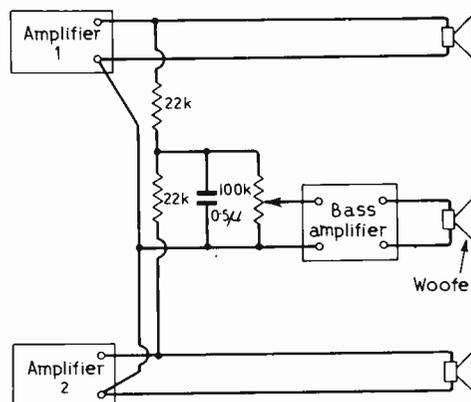


Fig. 11 from Mr Baxandall's original article.

quite adequate, though a more powerful one was used initially.

Peter J. Baxandall,†
Malvern,
Worcs.

References

1. H. D. Harwood: 'New B.B.C. Monitoring Loudspeaker — 2', *Wireless World*, April 1968, p.74.
2. F. E. J. Girling and E. F. Good: 'Active Filters — 6', *Wireless World*, January 1970, p. 27.

†Mr Baxandall recently resigned from the Royal Radar Establishment and is now undertaking consultancy work in electro-acoustics. Ed.

Ingenious tuning indicator

Readers may be interested in what I consider to be an improved solid-state tuning indicator for use with Mr Nelson-Jones' f.m. tuner.

The circuit retains the basic long-tailed pair amplifier Tr_1 , Tr_2 etc., as suggested by Mr Nelson-Jones, with the addition of a pair of switches, Tr_3 and Tr_4 , driven alternately by an astable multivibrator Tr_5 , Tr_6 etc.

When one of the switches say Tr_3 , is "on" it shunts the collector current of Tr_1 to the supply rail; D_1 serving to isolate the switch from the l.e.d. Current through the l.e.d. passes through D_2 to the collector of Tr_2 . The result is that the l.e.d. gives an indication of the current through Tr_2 on one half cycle of the multivibrator, and that through Tr_1 on the next.

When a station is correctly tuned the l.e.d. gives a steady light but any off-tuning is indicated by the diode blinking. The correct tuning capacitor position can be found far more accurately than with the two-light system, and probably with a similar accuracy to that obtained by using a centre-zero meter.

Component values indicated are those used in my prototype, the timing components giving a blink rate of the

order of 10Hz, which I found to be suitable. Too slow a blink rate fails to give a fast enough response to a tuning correction. Diodes D_1 and D_2 were 1544; transistors 3, 4, 5 and 6 were ZTX500, but many other types are suitable.

M. Gavins,
Cheadle,
Cheshire.

The author replies: I read Mr Gavins' letter with interest — it is certainly a most ingenious method of indication, not only for tuning indicators but for many other purposes. I would point out, however, that it does suffer from the difficulty of not indicating the direction of out of balance.

The circuit as shown by Mr Gavins can surely be simplified by removing Tr_3 and Tr_4 together with the two 8.2k Ω resistors, and replacing them with two diodes from the collector of Tr_5 (and Tr_6) to the collector of Tr_1 (and Tr_2) (anodes to transistors 5 and 6, cathodes to 1 and 2). If in the circuit shown, Tr_5 and Tr_6 are silicon alloy types then all is well, but if planar types are used something must be done to reduce the excessive reverse V_{eb} voltage to which these transistors will be subjected with a 12-V rail. Probably the simplest way is to connect two 2.2k Ω resistors from the collectors of Tr_5 and Tr_6 to the 12-V line, thus forming a potentiometer across each transistor and limiting the voltage on each to about 5.4V, which is usually below the zener breakdown of a base emitter diode. It may then be necessary to increase the two capacitors to 6.8 μ F, or thereabouts

My experiments with the tuning indicator circuit as published in the April 1972 issue indicate that the circuit provides a more than accurate enough method, as does the moving-coil meter. The tuner has a very wide discriminator characteristic and either tuning indicator will place the receiver on tune with the required degree of accuracy for stereo reception. Again my congratulations to Mr Gavins on an ingenious approach.

L. Nelson-Jones.

F.M. tuner instability

I was pleased to see Mr Nelson-Jones' article "F.M. Tuner Design — 12 months later" in the April issue.

This looked as if the answer to all the instability problems experienced was at hand. Various attempts to cure the instability in my particular tuner have been made, without success. The performance when tuned to the local f.m. stations has been excellent, but between stations the noise level has been high and television signals have been very troublesome.

The effect of adding the ferrite bead as suggested by Mr Nelson-Jones has only increased the number of spurious responses.

A determined attempt was then made to isolate the instability. The following additions were made to the circuit board:—

- An 80-ohm resistor was fitted between ground and aerial input tap of L_1 to terminate the aerial feeder cable. Instability appeared to be affected by moving a hand along the feeder, tending to indicate a standing wave on the feeder.
- The bypass capacitor of gate 2 of Tr_1 was increased by adding a 0.01 μ F capacitor.
- The bypass capacitor across the 220-ohm source resistor of Tr_1 was increased by adding a 0.005 μ F capacitor.
- The bypass capacitor decoupling the top end of L_2 was increased by adding a 0.01 μ F capacitor.
- The bypass capacitor decoupling the +12-V line supplying Tr_3 local oscillator was increased by adding a 0.01 μ F capacitor.

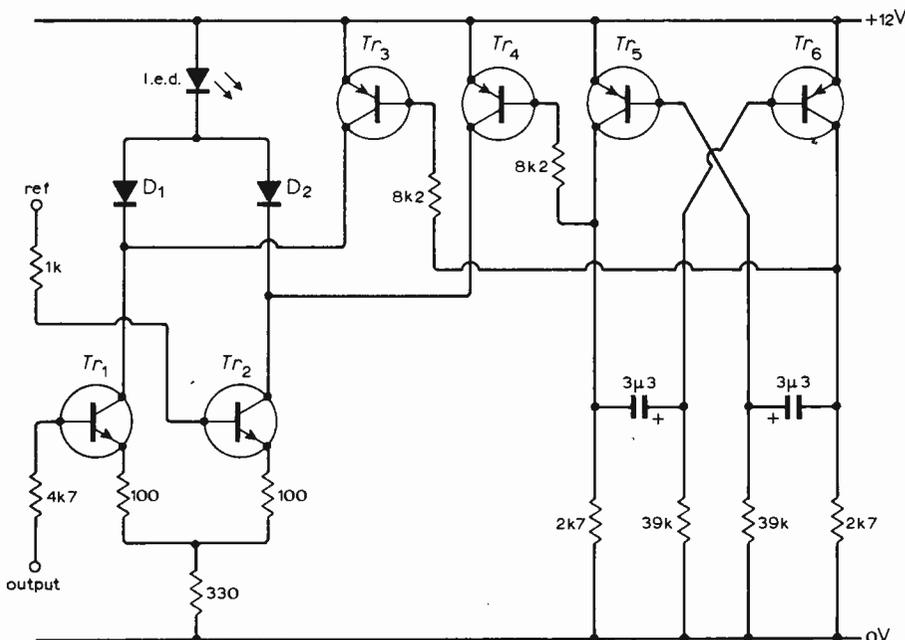
All capacitors were disc ceramic and were soldered in place on the underside of the circuit board. The original components, on top of the board, were not disturbed. The result of these modifications was stable performance with much lower interstation background noise and the ability to peak the r.f. and mixer tuned circuits from 87.5 to about 100 MHz.

There is, however, still instability at about 100MHz. This appears to be oscillation in the r.f. stage; running a hand along the aerial feeder cable still has an effect on the instability in this range.

Have other readers or Mr Nelson-Jones any further views on the causes and cures for instability in this tuner?

Roy Ellis,
Sutton Coldfield,
Warwicks.

The author replies: There has recently been some evidence of the trouble Mr Ellis describes, and this has been traced to u.h.f. oscillation of the r.f. amplifier stage. The reason why the addition of four capacitors did not completely remove the trouble is that the inductance of disc capacitors of this type is still not low enough for the frequency at which these oscillations take place, which can be up to about 1GHz. Why these oscillations in



some receivers are now more common is not clear but it may well be that the manufacture of the dual gate f.e.t.s has improved to give a higher cut-off frequency. Certainly the Texas 3N201 seems more prone to this trouble than the 40673, although on paper they are near identical products. Two readers have written to recommend a simple cure for the fault and I have myself done a deal of work to trace the cause; all three of us agree on the cure, though we use rather different values of component.

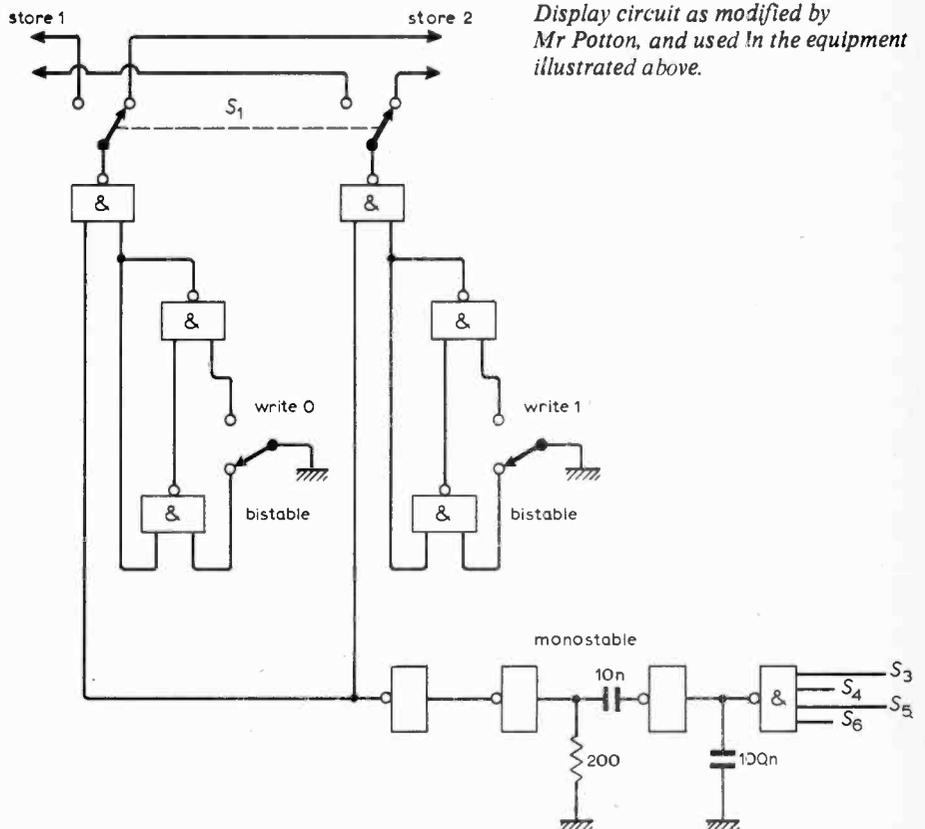
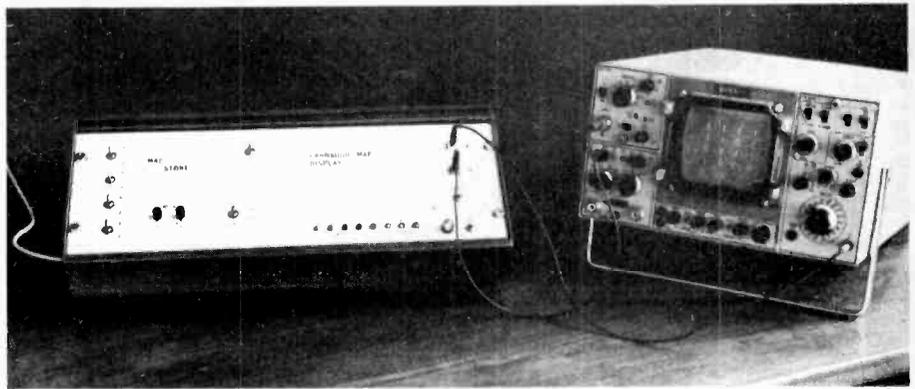
This cure is achieved by inserting an impedance in the feed to g_1 of the r.f. stage Tr_1 . I put in a 22-ohm resistor, one reader a 1 k Ω resistor and another a ferrite bead. The first and last will have very similar effects since the bead injects an effective resistance in the g_1 lead of around 20-30 ohms. The 1k Ω resistor is rather too high and will result in some loss of gain. The use of a resistor here will lower the Q of the input circuit slightly as it is in series with part of the total capacitance, but this effect has been found to be very slight. The resistor is the simplest to fit and should take the place of the link from g_1 of Tr_1 to the coil L_1 , immediately next to Tr_1 . The reason why the aerial feeder seems to take an active part is that the coupling of the feeder at the oscillation frequency is nothing like a good match, and hence gives rise to standing waves, and indeed the feeder may be the main resonant circuit element. This last point is backed up by the fact that with a direct 82-ohm resistor across the terminals of the tuner, or an attenuator box connected on short leads to the tuner, the effect is often absent. I am indebted to B. Brook, of Huddersfield, and J. R. Hooper (G3PCA), of Ilford, for their most helpful letters on this matter.

To sum up. The resistor fitted to replace the g_1 (Tr_1) to L_1 link should be of 22-47 ohms and may be a Mullard CR25 carbon film $\frac{1}{4}$ watt, or any carbon, metal or oxide film resistor of similar or smaller size. My sincere apologies to any readers who have suffered from this puzzling and very frustrating fault.
L. Nelson-Jones.

Karnaugh map display

You may be interested in some details of the Karnaugh map display equipment constructed by a member of our technician staff following publication of articles in your magazine in April and May 1971. As may be seen from the accompanying photograph, the completed unit makes an attractive addition to our teaching equipment.

The display circuit presented no problems apart from changes required in the bias conditions in the phase shift oscillator to improve the shape of the zero. This involved changing R_6 from 56k Ω to 100k Ω . The circuit associated with the 16-bit store did present problems however and several modifications were found to be necessary. These modifications were made with the object of obtaining a working



Display circuit as modified by Mr Potton, and used in the equipment illustrated above.

prototype in the shortest possible time. No doubt more elegant solutions to the problems could be found.

With the original circuit, it was found that selecting a particular cell in the map using the toggle switches and attempting to store a 1 or a 0 by operating the appropriate push button affected the contents of several cells in addition to the one selected. Investigations showed that spurious pulses at the output of the counter decode gate (SN7420 chip) were the immediate cause. These pulses probably result from the fact that the counter on the SN7493 chip is a ripple through circuit. Transient states at the counter output will occur as the ripple through action takes place. A crude but effective solution to this particular problem was a 0.1 μ F capacitor connected between the output of the counter decode gate and earth.

Two further problems remained, the first being that attempting to store a digit in a particular cell still affected the adjacent cell in addition to the one selected. The second problem was that if, for example, a 1 was to be stored, replacing a 0, and the appropriate button

was pressed, the 1 duly appeared in the map only to be replaced by a 0 again when the button was released. The modifications to the circuit required to overcome these problems are shown in the diagram above and consist essentially of (a) a simple monostable to shorten the width of the output pulse from the counter decode gate and (b) bistable pulse shapers associated with the push button switches.

A. Potton,
City of Leicester Polytechnic.

Constructors of the *Simple Electronic Multi-meter* (J. L. Linsley Hood, June issue) will find the following supplementary information useful. At frequencies above a few kHz, especially on the higher voltage a.c. ranges, the inevitable stray capacitances of the multiplier resistors are likely to impair the accuracy of the instrument. This can be avoided if a probe lead is made up, containing a suitable 1M Ω 2% series resistor (or ten 100k Ω 2% resistors wired in series), and the instrument is then used on the "a.c. current" ranges, e.g. 1mA \equiv 1kV; 30 μ A \equiv 30V; etc. With this arrangement the meter should be flat in response to 50kHz.

Reshaping Information to Suit the Channel

Advanced digital methods revealed at I.E.R.E. conference

In 1952 *Wireless World* published a report with exactly the same title as that above (November issue, pp.445-447). It was based on a symposium on communication theory held in London and dealt with signal modification, bandwidth compression and such like techniques designed to make the best possible use of available channel capacity in communication systems. Pressure of demand on channel space was building up as the post-war telecommunications industry expanded and its customers increased; and Shannon's work on information theory* had already suggested that it might be possible to approach the theoretical upper limit of channel capacity by a clever juggling with the three determining factors of information rate — bandwidth, time and signal/noise ratio. The notion of taking advantage of the limitations of human perceptions to restrict the required information rate had also been understood.

This was all in the "analogue days", as some engineers might be inclined to call them nostalgically. Since then digital computers have arrived, and with them data transmission, p.c.m. telephony, digital methods of processing signals (such as television standards converters), digital filters, digital measuring instruments, and the promise — or threat, depending on how you see it — of whole national and international p.c.m. communications networks. (There are already about 1000 Post Office 24-channel p.c.m. systems operating in the U.K.) Accordingly the techniques of "reshaping information to suit the channel" have changed markedly, though the fundamental principles remain. (It was perhaps prophetic that even back in the "analogue days" of 1948 Shannon's formulation of maximum channel capacity should be evaluated as an information rate in binary digits per second.) A wide survey of these modern techniques was presented in a three-day conference at Loughborough University of Technology on "Digital Processing of Signals in Communications", organized by the I.E.R.E. in association with the I.E.E. and the I.E.E.E. (April 11th-13th). The subjects of the sessions give some idea of the scope of digital processing

in communications: signal design (methods of encoding information); compression and expansion of coded information; adaptive equalization of data transmission circuits; digital filters; and computer simulation for designing communication systems. In this short article the forty papers cannot be adequately summarized† but the following selection of items may give some idea of the new techniques being developed.

Video telephones. Digital transmission of video information in television telephones — the Picturephone in America (just started in service) and the Viewphone in Britain — is desirable for several reasons. There is the nature of the picture signal, which need not be of such high quality as in entertainment television; the probable need for tandem switching requires digital transmission for distances over about 10 miles if acceptable performance is to be obtained economically; and, because the extensive development of this kind of service depends on the provision of wide-band transmission systems, there will necessarily be available channels suitable for pulse coded information. Since vision signals have a high information rate, compared with, say, speech signals, the need for making best possible use of available channel capacity — by eliminating picture redundancy, bandwidth compression, efficient coding, exploiting the natural limitations of human vision and so on — are particularly pressing.

†The bound conference proceedings can be obtained from the I.E.R.E., 8-9 Bedford Square, London WC1B 3RG.

In the first place the picture definition required for the small screen is not as great as for broadcast television, so the standards in use can be somewhat degraded. For example, in the first batch of experimental Viewphones being used for Post Office field trials there are only 319 lines (interlaced), the line scanning frequency is only 8kHz and the vision bandwidth is only 1MHz — all these figures being approximately half those of the u.h.f. television broadcasting standard. (The Viewphone field scan frequency is 50.15Hz and the picture aspect ratio 11:10.) The straightforward approach to digital encoding is to use p.c.m., but this would need about 7 binary digits per quantized sample (128 quantizing levels) for adequate quality, and this code is highly redundant for the application and its subjective requirements. The Post Office is therefore investigating — among other schemes — an improved code, called differential p.c.m. (d.p.c.m.) to obtain better utilization of the channel; and at the Loughborough conference a d.p.c.m. encoder was described and demonstrated by J. E. Thompson and E. A. Gerrard, of the P.O. Research Department.

The essence of differential p.c.m. is that it transmits not quantized samples of the vision waveform as in ordinary p.c.m. but, as the name implies, quantized differences between samples. Thus, when successive sample levels are the same in a part of the picture, no differences are transmitted and therefore the channel does not have to carry redundant information. In effect the d.p.c.m. encoder "predicts" that each sample of the signal will be equal to the previous one and merely transmits to the

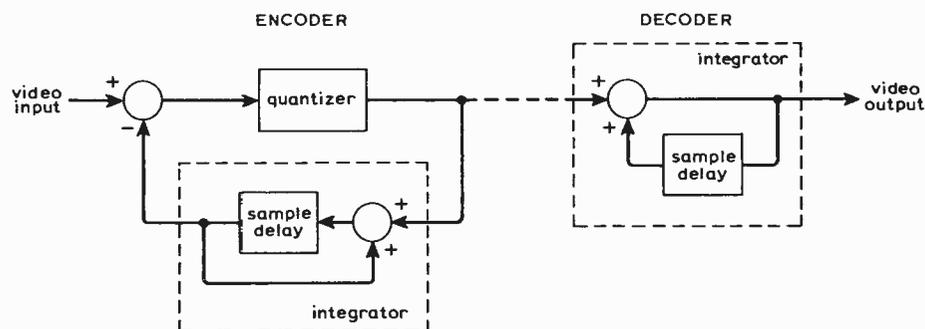


Fig. 1. Basic principle of a differential encoder-decoder as used for Post Office video telephone signals. (Binary codec for p.c.m. not shown.)

*Particularly his classic paper "A Mathematical Theory of Communication", *Bell System Technical Journal*, vol. 27 (1948) pp. 379, 623.

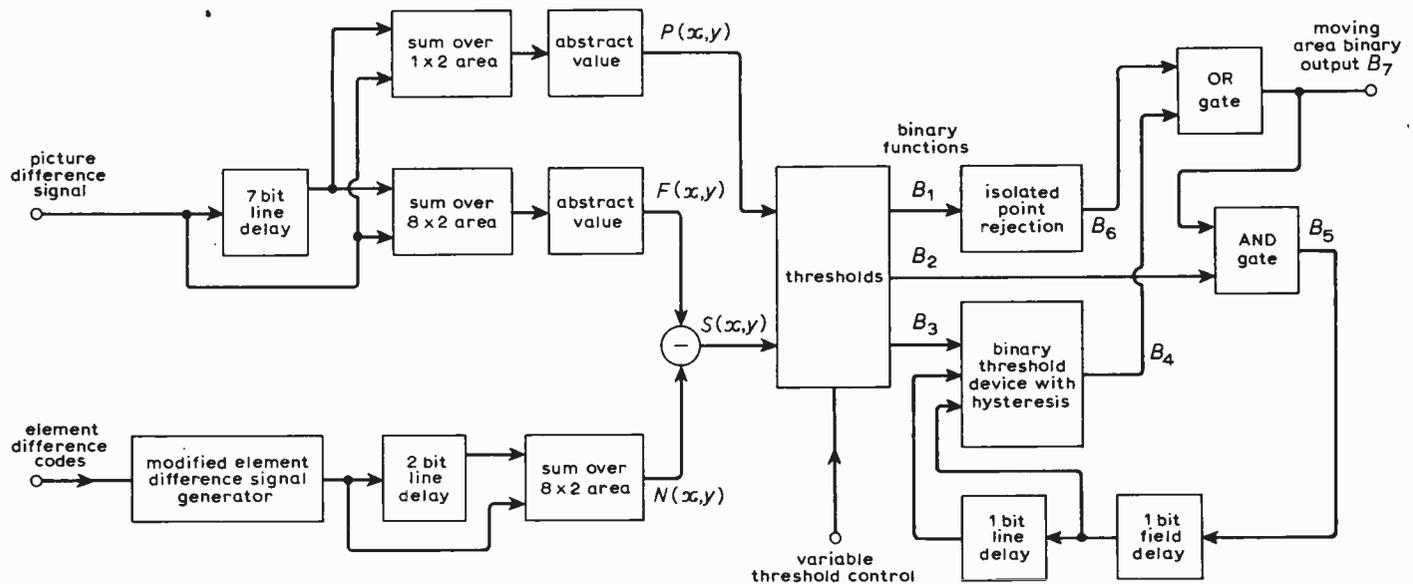


Fig. 2. Schematic of system for separating moving areas of the picture from the static background, intended for the Picturephone video telephone. Information from the present field as well as from the previous field is used.

receiver, sample by sample, the amount of error in this prediction. The three electronic operations required for this process are differentiation of the signal, "tapered" quantization (the quantized samples describing the instantaneous slope of the waveform) and integration. At the receiver decoder an integration process reverses the effect of the differentiation at the transmitter encoder. To avoid an accumulation of quantizing error by the receiver's integrator, in the encoder differentiation is performed by feedback around an integrator and subtractor, as shown in Fig. 1, and the quantizer is included in the loop. As a result the quantized difference transmitted is not the difference between two picture samples but the difference between a new picture sample and the accumulation of all the previous quantized differences sent to the receiver.

This d.p.c.m. system takes advantage of the fact that the eye is particularly sensitive to noise and quantizing contours in the low detail (gradually changing) regions of a picture, such as soft shading in the human face, but will tolerate considerable noise and amplitude distortion on samples in detailed regions and at edges. The combined operations of differentiation followed by tapered quantization have the effect of separating the areas of low and high detail and of quantizing those areas accordingly. In low-detail regions, where the sample differences are small, the system operates at the centre of the tapered quantizer characteristic and makes suitably small quantizing errors. As picture detail and the sample difference increase, quantizing errors are increased proportionately. Optimum use can therefore be made of a restricted number of quantizing levels by adjusting thresholds to minimize granularity and contouring in low-detail areas, while compromising in making the outer levels as large as possible to reduce an effect known as "slope overload" (which also occurs in delta modulation).

Hardware of the encoder is constructed

from 130 integrated circuits, of which 89 are m.s.i. arithmetic units. As it is an experimental equipment intended only for evaluating d.p.c.m. transmission, the Post Office is not giving any direct information on performance, but the authors say that it makes possible adequate picture quality with an information content of 4 bits per sample (16 quantizing levels) and a maximum sampling rate of 2.5MHz, giving a maximum information rate of 10Mbits/second. By comparison, ordinary p.c.m., to give comparable picture quality, would require about 8 bits per sample.

Another data compression system which takes advantage of intra-field (picture-element to picture-element) redundancies in a television picture was described by B. Wendland and F. May, of AEG-Telefunken (Germany). This elaboration of d.p.c.m. is described as an "adaptive" coder. It transmits only "relevant" information, which it continuously selects as the picture changes, in accordance with a "relevance" criterion based on human visual perception. For this it takes advantage of the relative insensitivity of the eye to noise and distortion at edges and in detailed regions of the picture, as mentioned above. The equipment, however, was designed not for video telephone standards but for 625-line, 50 fields/sec interlaced pictures with a bandwidth of 4.5MHz — approximately broadcasting definition.

The "relevant" information consists of only those element-to-element differences which exceed a threshold value, and this threshold varies in accordance with the amount of detail in the picture. In addition to its "relevance" detector the Telefunken encoder includes a system, using a buffer store, for smoothing out the transmitted digit rate, so that no sudden bursts of high digit rates can occur that will exceed the channel capacity. To achieve a constant data rate of binary digits out of the buffer store two expedients are adopted. Underflow of digits is avoided by generating and

feeding into the buffer what are picturesquely described as "stuffing code-words", while the overflow of the buffer store, caused by an excessive rate of incoming digits, is avoided by means of a feedback signal from the buffer store to the "relevance" detector. This feedback signal depends on the number of digits held in the buffer store and controls the "relevance" detector to limit the amount of "relevant" information as necessary.

As for performance, the authors claim that the adaptive coder gives a data compression ratio of 1:5 relative to a 7-bit p.c.m. television picture.

In addition to intra-field redundancy, television pictures also have inter-picture redundancy — for example, when the whole of the televised scene is not moving, or when certain parts of a picture, such as background, are temporarily still. This is the feature which is being considered for possible use in a data compression system for the Picturephone video telephone in the U.S.A. D. J. Connor, B. G. Haskell and F. W. Mounts, of Bell Telephone Laboratories, described an inter-picture coder working on this principle which they have simulated on a computer — an example of the computer simulation techniques mentioned above — and showed a film of the results obtained. The coder is designed to accept the output of a differential p.c.m. coder of the kind outlined above, so that, overall, both intra-field and inter-picture redundancies are reduced. The final data rate is then low enough so that the output can be transmitted with satisfactory picture quality in a 2Mbit/second channel.

An interesting feature of the system is that the incoming picture to the inter-picture coder is separated into moving areas (e.g. a person's face) and background areas. The moving areas are then transmitted using a number of data compression techniques. During periods of slow movement clusters of picture-to-picture differences in the moving areas are

transmitted. For moderate movement, picture differences are sent only in every other field period, the moving areas of intervening fields being transmitted by an interpolation technique. With violent movement the buffer store of the coder (see above) is likely to become filled and then pictures are repeated.

A block diagram of the system which separates the moving areas from the stationary background areas is shown in Fig. 2. In order to separate the picture-to-picture brightness changes caused by movement from those caused by the noise from an element difference quantizer, advantage must be taken of the properties of the unquantized signal (i.e. the picture differences produced by movement) and of the noise (i.e. the picture differences caused by the element difference quantizer). The important properties of the unquantized signal are that it is both spatially and temporally correlated. The important properties of the noise are that it is uncorrelated spatially; it is dependent on the magnitude of the quantized element difference signal; and the magnitudes of individual noise spikes are proportional to the spacing of the representative levels. It is by exploiting the different properties of the signal and the noise, that the moving/stationary separation is achieved.

The spatial correlation of the signal and the lack of correlation in the noise are exploited by forming the sum of the picture difference signal in a 1×2 units area to give $P(x, y)$ in Fig. 2 and in an 8×2 units area to give $F(x, y)$. The last two properties of the noise lead to the generation of an artificial signal called the modified element difference signal, this being proportional to the level spacing for the outer quantized levels and being zero for the inner levels.

If the magnitude of the modified element difference signal is summed over an 8×2 units area, a function, shown as $N(x, y)$, is obtained that is proportional to the expected picture difference noise due to quantization. If it is subtracted from $F(x, y)$, the function which results, shown as $S(x, y)$ is effectively an average picture difference signal with compensation for the detail-dependent variations of the picture difference noise. It is a very sensitive indicator of spatially correlated picture differences and is insensitive to quantization noise.

The functions $S(x, y)$ and $P(x, y)$ are sent along independent paths and either of them can cause a point to be included in the moving area, i.e., cause the binary output function, B_1 , to be a "1".

Function $P(x, y)$ is used to detect slowly moving, high contrast edges which are characterized by large picture differences. It is converted to a binary function B_1 using a high threshold. B_1 is further processed by rejecting any binary 1s that are flanked on both sides by two or more 0s. The resultant function, B_6 , is combined with B_4 in an OR gate to obtain the output binary function B_7 .

The function $S(x, y)$ is best suited to detecting large-area movement. It is converted into two binary functions, B_2

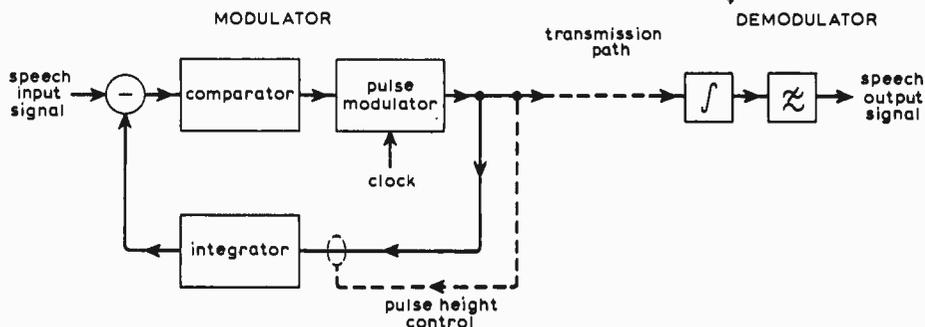


Fig. 3. Basic delta modulation system, as used for speech. An additional feedback loop for adaptive modulation, using pulse height control, is shown as a broken line loop.

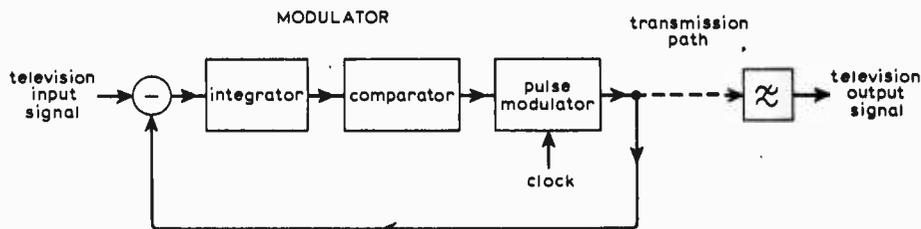


Fig. 4. Delta-sigma modulation system as applied to television signals.

and B_3 , by comparing it with thresholds, T_2 and T_3 , where $T_2 < T_3$. The function B_3 and the values of B_2 from the adjacent lines in the previous field are combined in a threshold logic device so as to exploit the temporal (and spatial) correlation of a moving area. This hysteretic device monitors the number of binary 1s in an 8×3 area around a point. If the number exceeds a threshold (about 9 out of a possible 24), the device switches to its "on" state, the output function B_4 is a 1, and the point is included in the moving area. The number of 1s must then fall below a second threshold (about 4 out of a possible 24), before the device switches to its "off" state.

The function B_5 provides the temporal correlation for the system and is obtained by using the function B_7 to gate the function B_2 . Since B_2 was obtained by operating on $S(x, y)$ with a lower threshold than for B_3 , it is a more sensitive motion detector. However, it also introduces more noise. In areas where no motion is detected B_2 is inhibited by the AND gate. In areas where motion is detected, B_2 is passed, in effect sensitizing the binary threshold device in the next field.

If the $S(x, y)$ and $P(x, y)$ thresholds are fixed, they must be set quite low in order to detect very slow motion. Given the level of quantization noise from a differential p.c.m. coder, such low thresholds inevitably lead to the inclusion of some background points in the moving area. If some measure of the speed of motion is available (e.g. buffer store contents), the thresholds can be made dependent on speed. For even moderate motion the separation is then virtually free from errors.

The differential p.c.m. method of information transmission outlined above is rather like delta modulation¹ in that it sends pulses describing the instantaneous slope of the original waveform. Delta modulation is now being used as an

inexpensive method of digitally encoding speech signals, particularly in military communication systems. The basic principle of the modulating and demodulating system is shown in Fig. 3. The modulator has a feedback path comprising an integrator, the output of which is made to follow the input signal. The modulator produces a train of positive pulses, which are fed into the integrator until its output voltage exceeds the incoming signal voltage. The comparator detects this condition and causes the modulator to produce negative pulses until the output voltage from the integrator becomes less than the incoming speech signal voltage. In this way, the modulator follows the incoming pulse signal. The pulse train feeding the integrator is also the transmitted pulse signal. At the receiving end, the demodulator uses a similar integrator, followed by a low-pass filter, to reconstruct a signal similar to the original.

A simple delta modulation system cannot transmit the d.c. component of the input signal, and the signal-to-quantization noise ratio, and thus the dynamic range, vary inversely with frequency. Moreover, the integrator at the receiving terminal causes an accumulation of error voltages when the system is subjected to transmission disturbances. This is when the system is mainly used for speech signals which are without a d.c. component and contain less energy at high frequencies than at low.

For television signals, which do contain a d.c. component and have a more uniform energy spectrum, a modified version of the system, known as delta-sigma modulation, is preferable. This is obtained by altering the position of the integrating network in the feedback loop, as shown in Fig. 4. Since there is now an integrator in the forward path of the sending terminal, no integrator is required at the receiving end. Consequently, the d.c. level of the signal is maintained and transmission errors are not

accumulated. Moreover, the de-emphasis introduced at the transmitting end results in integrated quantizing noise power being independent of signal frequency.

A disadvantage of delta modulation systems is that quantizing noise is almost independent of the level of the input signal, and so is extremely objectionable at low signal levels. This restricts the range of input levels that can be handled and simple systems typically have a dynamic range of only 15dB. When the original input signal has a greater range of power levels, some form of compression and subsequent expansion (companding) must be used. The companding action can be obtained by using a similar expanding characteristic at the receiving end and in the feedback path at the transmitting end. This can be achieved by means of an additional feedback loop to control the height of the pulses fed to the integrator circuit, as shown by the broken-line f.b. circuit in Fig. 3.

Recent companders use additional feedback circuits of this kind based on digital logic. When the feedback signal in the modulator is closely following the input signal, the output pulse train does not contain a preponderance of positive nor of negative pulses. If the number of positive or of negative pulses predominates, this indicates that the feedback voltage is not following the input signal sufficiently accurately and that the height of the input pulses to the integrator should be increased. This can be controlled by storing the last few transmitted pulses in a shift register and determining the number of positive and negative pulses contained in the store.

Such a controlled system has become known as adaptive delta modulation. M. J. Hawksford, of Essex University, presented a study, by himself and Professor J. E. Flood of Aston University, of the positive and negative pulse groupings in the output of a delta-sigma modulator and their frequency of occurrence with respect to the input signal. The results of this study, performed by computer simulation, have led to a proposal for an adaptive delta-sigma modulator, using high-speed logic, which could be used for encoding television signals.

Another use for delta modulation may prove to be in digital filters. It is notorious that digital filters, although they have the

advantages of being stable in operation and giving a truly linear phase characteristic, are very expensive to build. This is because they are usually based on p.c.m. techniques, in which signal samples are encoded as groups of binary digits and linear difference equations are performed on these groups by digital arithmetic circuits to achieve the required filtering characteristic. Moreover, if an analogue output is required the output from the digital filter must be passed through a further process, of digital to analogue conversion. Perhaps the coming of l.s.i. semiconductor devices will make all this possible at about 3p per chip by 1975, but meanwhile G. B. Lockhart, of Leeds University, has suggested that delta modulation provides an inexpensive alternative to the p.c.m. method and has shown by computer simulation that "delta modulation filters" could be relatively simple devices.

The essence of the delta modulation filter is shown in Fig. 5. It consists of a delta modulator feeding into a binary transversal filter², which is basically a tapped shift register with means for individually weighting the outputs from the taps and combining the weighted outputs. The pattern of weighting determines the impulse response required. A number of digital second-order resonators based on this configuration have been simulated on a computer, and Dr Lockhart said they revealed no essential difference from the performance of conventional analogue filters. He commented that they were particularly suited to applications where analogue-digital interfacing is normally required. One speaker thought that in practice they might even prove cheaper to build than conventional filters.

Although speech signals do not occupy as much bandwidth, nor require such a high digit rate in bits/second, as television signals, the number of speech channels used in public telecommunication systems is much greater than the number of television channels, and is likely to continue so. Consequently there is plenty of motivation for applying data compression techniques to speech. As an example, J. R. Sergio, of Bell Labs, U.S.A., described a system of data companding for p.c.m. encoded speech signals which he had simulated on a digital computer and also outlined a practical

data compressing encoder which he said could be made with standard emitter-coupled logic gates and other devices.

The basic principle of compression is to eliminate sequences of levels in the quantized speech signals (the levels produced by sampling, prior to binary coding, in p.c.m.) when those levels occur at equal spacing — for example, when a part of the signal corresponds to a linear variation of amplitude with time. Thus, bits are transmitted only during those intervals when the original speech waveform does not have linear variation with time. In the encoder the original speech signal is applied to a sample-and-hold circuit, the output of which, $S(t_s)$, where t_s is the sampling time, is applied to one input terminal of a comparator. The second input terminal of the comparator receives a sequence of levels produced by quantizing the output, $R(t)$, of a linear ramp generator: the time sequence of these levels is the time sequence of the p.c.m. levels to be used. The comparator produces a binary 1 during each interval Δ_i and these digits are fed into a binary counter which counts up from zero until $R(t) \geq S(t_s)$. The corresponding binary digital representation of $S(t_s)$ is then read out of the counter at the same time as the next input signal sample is applied (each t_s seconds), and the counter is zeroed. Thus, binary digits from the comparator are counted only when the signal level differs from the ramp level. When they increase together in step — the signal varying linearly with the ramp — there is no difference and no digits are counted and read out of the encoder. Compression is produced by appropriate spacing of the voltage levels of $R(t)$.

One of the earliest and best known speech coding and information compression devices is the "vocoder", which achieves bit-rate compressions of the order of 10:1 by transmitting the speech spectral envelope, plus "pitch" information to allow fine spectral detail to be re-inserted at the receiver. Relatively simple compared with this is a technique based on signal interruption which was described by J. S. Severwright, of Loughborough University (his co-author being G. B. Lockhart of Leeds University). Although this scheme provides bit-rate compressions of the order of only 3:1, it is claimed to avoid many of the performance problems encountered in more efficient coding schemes. It operates by regularly transmitting short sections of signal. Typically 10ms sections of speech might be transmitted, separated by intervals of 10 to 20ms, to obtain bit-rate reductions of 2 or 3:1. Demonstrations by Mr Severwright showed that this "interrupted" speech is highly intelligible in itself; and, in fact in an early paper describing the effects of interruption on speech, Millar and Licklider³ reported intelligibility results above 80%. But the interrupted speech sounded extremely distorted, its "pitch" and individual character being largely eliminated.

The interrupted signal may be considered as a multiplication of the speech signal $s(t)$ with a periodic low frequency rectangular

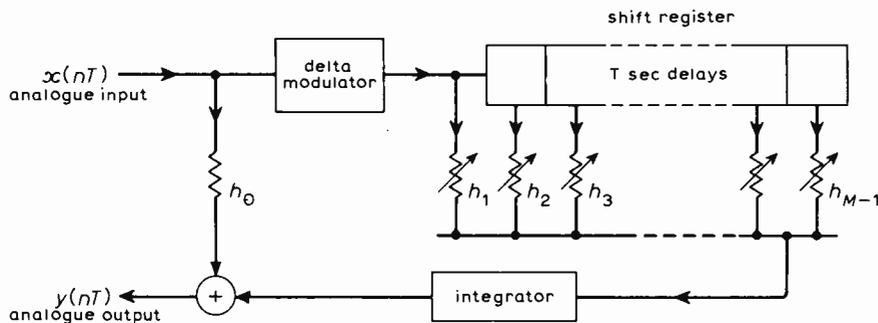


Fig. 5. Digital filter based on delta modulation in binary transversal filter technique. The resistors h_1, h_2, \dots, h_{M-1} represent different weightings applied to the tap outputs of the shift register. The input and output signals are functions of n , the number of samples, and of T the delta modulator clock period (in seconds).

"interrupting" function $m(t)$. Interruption therefore results in mathematical convolution of the speech spectrum with the relatively narrow spectrum of the interrupting function $m(t)$. The resulting spread of the speech spectrum is limited, having little effect on the characteristics of the spectrum which apparently give intelligibility, but seriously modifying fine spectral detail and thereby degrading the quality.

In order to reproduce a speech signal of adequate quality the receiver must generate suitable signals to fill the interrupted periods.

The schemes being investigated at Loughborough for this purpose use a speech waveform predictor which generates signals for the interrupted periods from those sections of speech which are transmitted. Prediction techniques using pitch period, growth/decay of excitation function amplitude, and change in short-term spectrum are being examined. At the receiver the outputs during interrupted periods are predicted directly from the transmitted sections of speech, without intermediate processing. As a result certain

speech characteristics may be reliably retained during processing, allowing faithful reproduction.

References

1. F. de Jager, "Delta modulation, a method of p.c.m. transmission using the one-unit code" *Philips Research Reports* vol.7, p.442 (1952). See also *Wireless World*, October 1952, p.427.
2. "Transversal filter", *Wireless World*, May 1968, p.101.
3. G. A. Millar and J. C. R. Licklider, "Intelligibility of interrupted speech," *J. Acoust. Soc. America*, vol.22, p.162 (1950).

Experiments with Operational Amplifiers

2. Measurement of input bias currents and offset voltage

by G. B. Clayton, B.Sc., F.Inst.P.

A differential input op-amp draws small currents, called bias currents, at each input terminal; and for the amplifier to give zero output voltage a small differential input voltage, called the input offset voltage (V_{io}), must be applied to it (because if the input terminals are shorted together the op-amp is found to give non-zero output). Amplifier bias currents and offset voltage can be measured directly with the amplifier used open loop, but it is usually more convenient to take measurements on a closed loop test circuit. An example of such a closed loop test circuit is illustrated in Fig. 2.1.

Offset voltage and bias current effects,

which may be represented by equivalent generators at the amplifier input, are multiplied by the closed loop gain, $(1 + R_2/R_1)$, and measurements are made at the low impedance output terminal (6) of the amplifier. The voltage at the amplifier output terminal is measured under the following conditions: S_1 and S_2 closed, output voltage e_{o1} ; S_1 closed S_2 open, output voltage e_{o2} ; S_1 open S_2 closed, output voltage e_{o3} ; S_1 open, S_2 open, output voltage e_{o4} .

The effect of opening a switch causes bias current flowing through R_3 to apply a voltage at the input. Resistor R_1 is made sufficiently small to allow $I_b^- \cdot R_1$ to be neglected.

The following equations hold for the four measurements:

$$e_{o1} = \left[1 + \frac{R_2}{R_1}\right] V_{io} \quad (2.1)$$

$$e_{o2} = \left[1 + \frac{R_2}{R_1}\right] [V_{io} - I_b^+ \cdot R_3] \quad (2.2)$$

I_b^+ is the bias current drawn by the non phase inverting input terminal.

$$e_{o3} = \left[1 + \frac{R_2}{R_1}\right] [V_{io} + I_b^- \cdot R_3] \quad (2.3)$$

I_b^- is the bias current drawn by the phase inverting input terminal.

$$e_{o4} = \left[1 + \frac{R_2}{R_1}\right] [V_{io} + I_{io} \cdot R_3] \quad (2.4)$$

$I_{io} = (I_b^- - I_b^+)$ is the input offset current.

A set of typical readings using the component values of Fig. 2.1 are:

$$e_{o1} = -0.12V, \quad e_{o2} = -9.8V,$$

$$e_{o3} = +9V, \quad e_{o4} = -0.75V.$$

Substitution in equations 2.1 to 2.4 gives

$$V_{io} = -0.12mV, \quad I_b^+ = 98nA,$$

$$I_b^- = 91nA, \quad I_{io} = -6nA$$

The discrepancy in the measured value of I_{io} is due to inexact matching of resistors R_3 .

The experiment can be repeated for different values of the supply voltages or under different temperature conditions.

Note: The measured parameters will be found to have different values for different amplifiers of the same type, and it may be found necessary to select alternative values for the resistors if the amplifier output approaches saturation in any of the measurements.

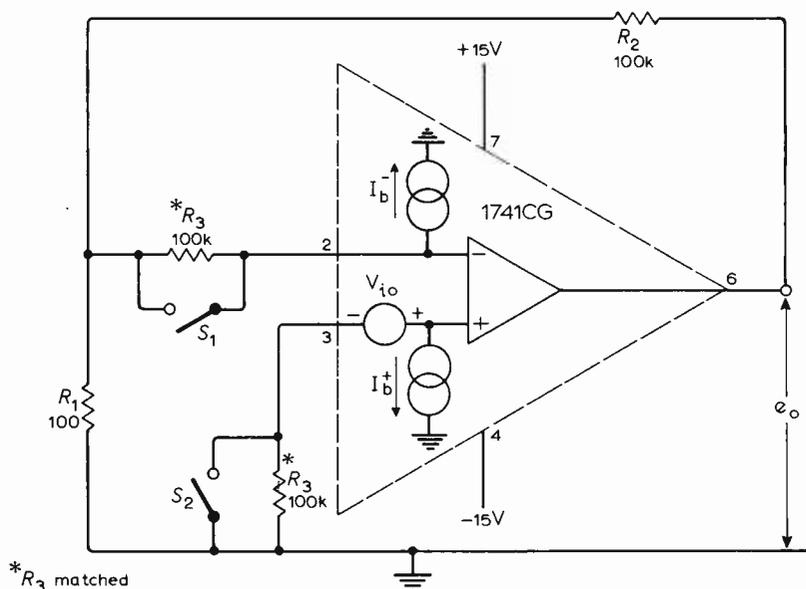


Fig. 2.1. Closed loop test circuit for measurement of input bias currents and input offset voltage.

10-80 Metre Amateur Transceiver

2: The receiver

by D. R. Bowman, G3LUB

At this point the author wishes to emphasize that a project of this complexity is not for the inexperienced. Any potential constructor would be expected to have had considerable previous experience in building complex equipment as the capital outlay necessary for the project is large. The availability of at least a valve voltmeter with an r.f. probe and some type of simple signal generator would be wise.

Fig. 2 is a block diagram showing the required gain distribution throughout the receiver. The three basic blocks which are common both to transmit and receive (local oscillator synthesizer, 9MHz crystal filter and crystal controlled b.f.o.) will be dealt with as part of the receiver.

The layout used in the prototype can be seen in the photographs, Figs 3 and 4. The more detailed photographs, which appear later, should be viewed in conjunction with Figs 3 and 4 and last month's front cover.

Radio-frequency amplifier

The requirements for the r.f. amplifier are:

- Very good immunity from cross-modulation and blocking at all signal levels.
- A low noise figure (this is secondary in importance to the above).
- Enough gain to ensure that the noise of the r.f. amplifier overrides that of the following mixer (the mixer noise should contribute only about 1dB to the overall receiver noise figure).
- Good stable circuit design using amplifying devices with low feedback capacitances for low noise performance.
- The a.g.c. voltage requirements must be compatible with the rest of the receiver.

The introduction of dual-gate field-effect devices has eased the cross-modulation problem considerably. These devices, which are basically two cascade connected single gate f.e.t.s, exhibit both a very wide a.g.c. range (Fig. 5) as well as an approximately square law transfer characteristic. The cross-modulation that does still occur is a function of the deviation from square law of this transfer characteristic. This deviation is small and mainly occurs with very large signal, thus the justification for the aerial attenuator which is to be described later. These devices fulfil all the requirements noted above with the exception of the last point which can be overcome using a simple circuit.

The author has considerable experience in using 3N140 dual gate devices and as a result of a large number of destroyed transistors all constructors are strongly advised to purchase only the gate protected types. The extremely high internal resistance of the 3N140 gate electrodes means that static

charge can easily accumulate until the 50V rating is exceeded. A list of the available protected devices is to be found in the components list. The manufacturers have incorporated minute zener diodes across the gate-to-source junction and these devices have been found particularly robust.

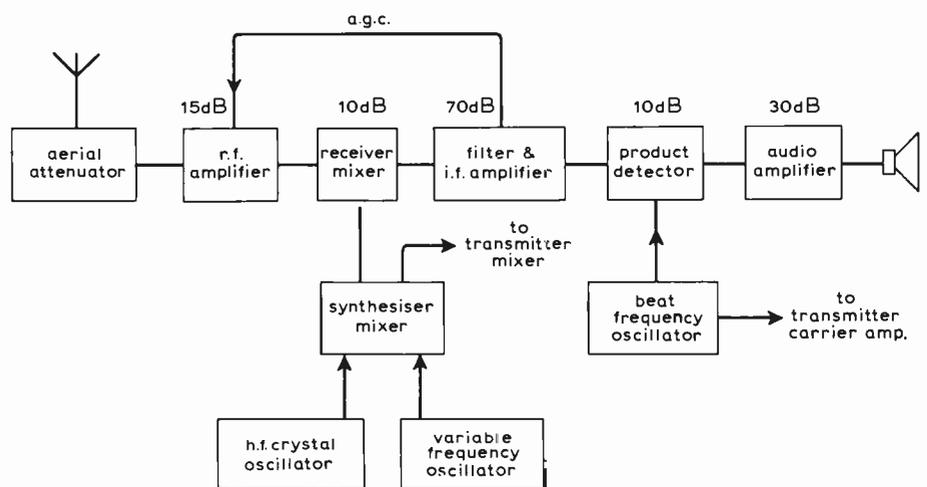


Fig. 2. Block diagram of the receiver section of the transceiver showing gain distribution.

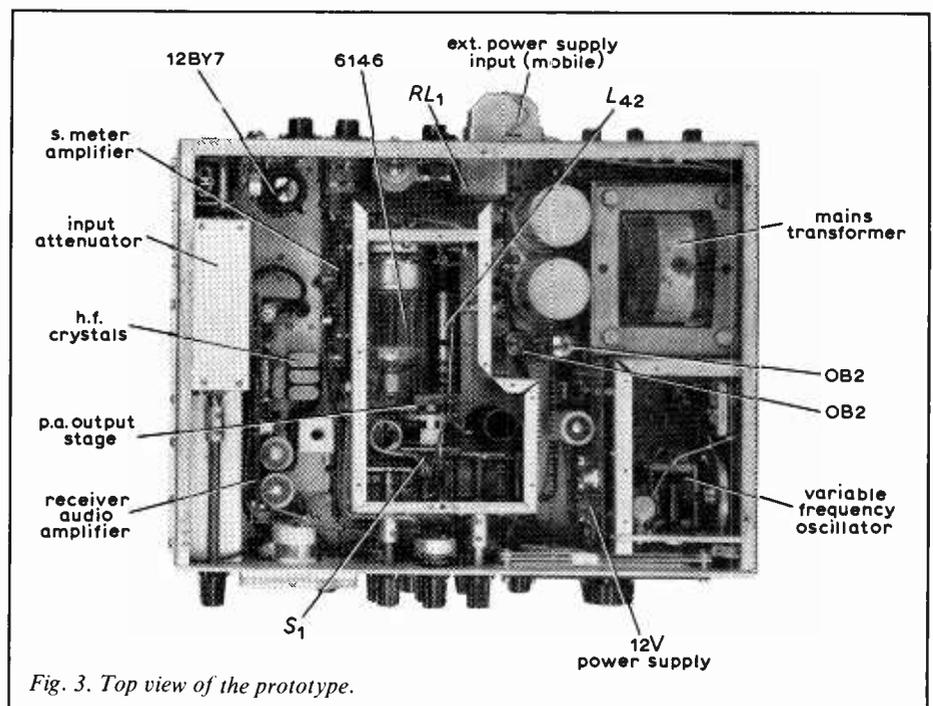


Fig. 3. Top view of the prototype.

Transistor Tr_1 in Fig. 6 is the r.f. amplifier. The circuitry is reasonably straightforward, the f.e.t. being connected as a common source amplifier. The system encountered in a number of Japanese transceivers of using one main coil for 40m and then switching in parallel C or L to resonate the r.f. circuit to other bands is to be found in both the input and output circuits of the amplifier. This reduces the number of coils and switch wafers without appreciably degrading the performance. The two diodes connected across the input circuit are there to protect the r.f. stage from damaging overload, but if gate-protected f.e.t.s are used they may not be necessary. The author included them having in mind the close proximity of the 50W p.a. which is isolated only by the aerial changeover relay.

The dual-gate f.e.t.'s a.g.c. requirement of -2 to $1V$ can easily be shifted to 1 to $4V$ by introducing a zener diode biased (D_3) artificial earth line. All the circuitry surrounding Tr_1 is referred to this rail instead of to real earth.

One of the problems inherent in using a high i.f. is the difficulty of obtaining an adequate i.f. breakthrough rejection. The basic rejection is not much greater than 40dB when the receiver is tuned to 40m and therefore it has been necessary to include a two-stage 9MHz rejection filter between the aerial attenuator and the receiver's first tuned circuit. The filter (L_1, C_1 and L_2, C_2, C_3) adds at least 30dB to the i.f. breakthrough rejection.

Construction: The layout of the r.f. amplifier is quite critical. It was found necessary to position Tr_1 halfway through a carefully cut hole in the screening wall, thus isolating the input from the output. Even with this precaution Tr_1 will self oscillate if the a.g.c. voltage is allowed to rise too high. It was found that more than adequate stable gain variation was available with the a.g.c. voltage restricted. The dynamic range is reduced only slightly and the required

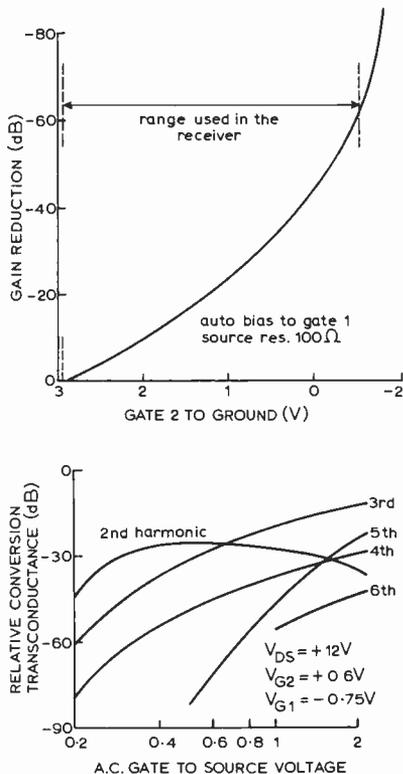


Fig. 5. Showing a.g.c. performance and the relative conversion gain of several harmonics as a function of oscillator voltage for the 3N201.

gain control range is still obtainable. It must be stressed that careful earth point and coil layout is essential if a stable amplifier is to be constructed. Some readers may say why not neutralize the stage? This cannot be done due to the considerable variation in stray capacitance caused by the range changing switches and the variation in device characteristics with a.g.c. changes.

Aerial attenuator

In view of the extremely high signal strengths experienced on some of the h.f. amateur bands it was decided to include a switchable

attenuator ahead of the r.f. amplifier. A six-position switch is used which provides attenuation ratios of 0, 6, 12, and 18dB. The fifth position short circuits the receiver input whilst the sixth position is used to bring in an auxiliary aerial socket, thus bypassing the aerial changeover relay. This allows either an external converter or a separate aerial to be used for reception. The attenuator consists of a number of resistance pi sections, the ratios being chosen to suit the 6dB per S point system (explained later). The attenuator is remarkably accurate, but carbon resistors with their low inductance must be used.

Signal mixer

The dual-gate f.e.t. has a definite advantage over single-gate devices when used as a mixer. The single-gate device relies on the non-linearity of the transfer characteristic at low drain currents. This process is akin to the valve bottom bend mixer while the dual-gate device is similar in operation to the multigrad valve. The isolation between the two gate electrodes curbs the tendency for the local oscillator to feedback to the aerial via the r.f. amplifier. This same isolation presents a constant, frequency-independent, load to the local oscillator. This is almost impossible when using mixer circuits where both signals are injected into a common electrode.

In Fig. 5 it can be seen that the conversion gain is a function of the level of the local oscillator voltage. The diagram also shows that the spurious generated within the mixer are l.o. amplitude dependent. As mixer gain is of secondary importance to the reduction of unwanted signals the level of oscillator drive is not allowed to exceed 500mV r.m.s. The stability of the potentials (d.c.) applied to both gate electrodes is important and for this reason the potential divider network and gate 1 resistor are returned to the source electrode and not to earth (Tr_2 , Fig. 6).

Intermediate frequency amplifier

The i.f. amplifier has to provide 70dB gain to bring the signal level up to the amplitude necessary to drive the detector and a.g.c. amplifiers. It must obviously be stable throughout the 80dB a.g.c. range and sufficiently frequency selective as not to degrade the signal-to-noise ratio of the signal present at the output of the mixer. The application of a.g.c. must have no effect upon the frequency response which must always be determined by the crystal filter.

Many circuit configurations were considered for use in the i.f. amplifier. The common-emitter transformer coupled technique was avoided as the poor isolation would have necessitated either very low gain per stage or neutralization. The use of dual-gate f.e.t.s was considered, but although the performance would have been excellent the price of these devices is still much higher than good bipolar types. The inherent superior cross modulation performance would not have been such a great asset as the amplifier follows the narrow eight-pole crystal filter. For cross modulation to occur two signals at least have to be present in the amplifier and the filter should

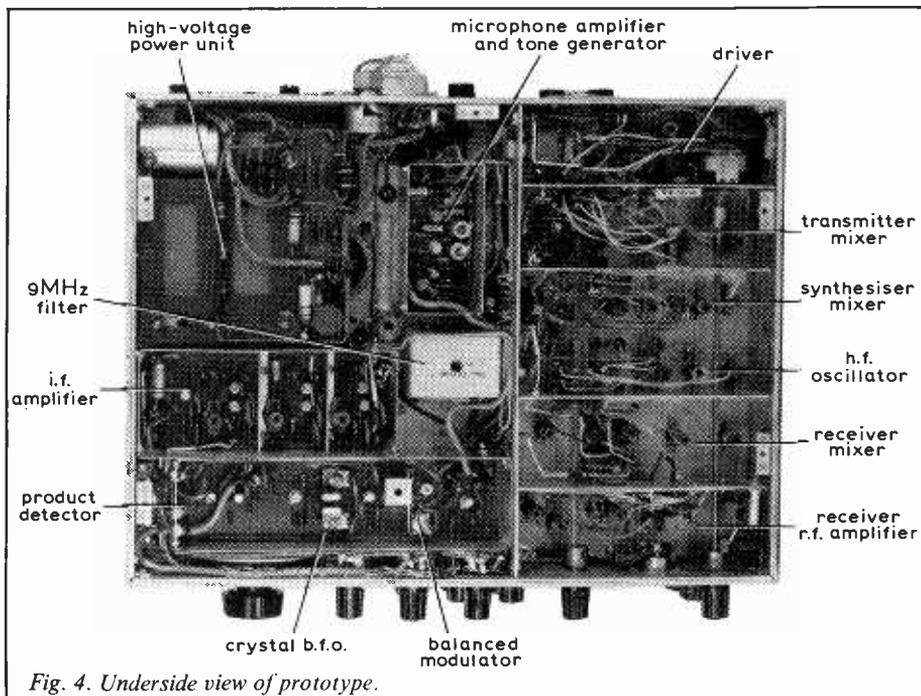


Fig. 4. Underside view of prototype.

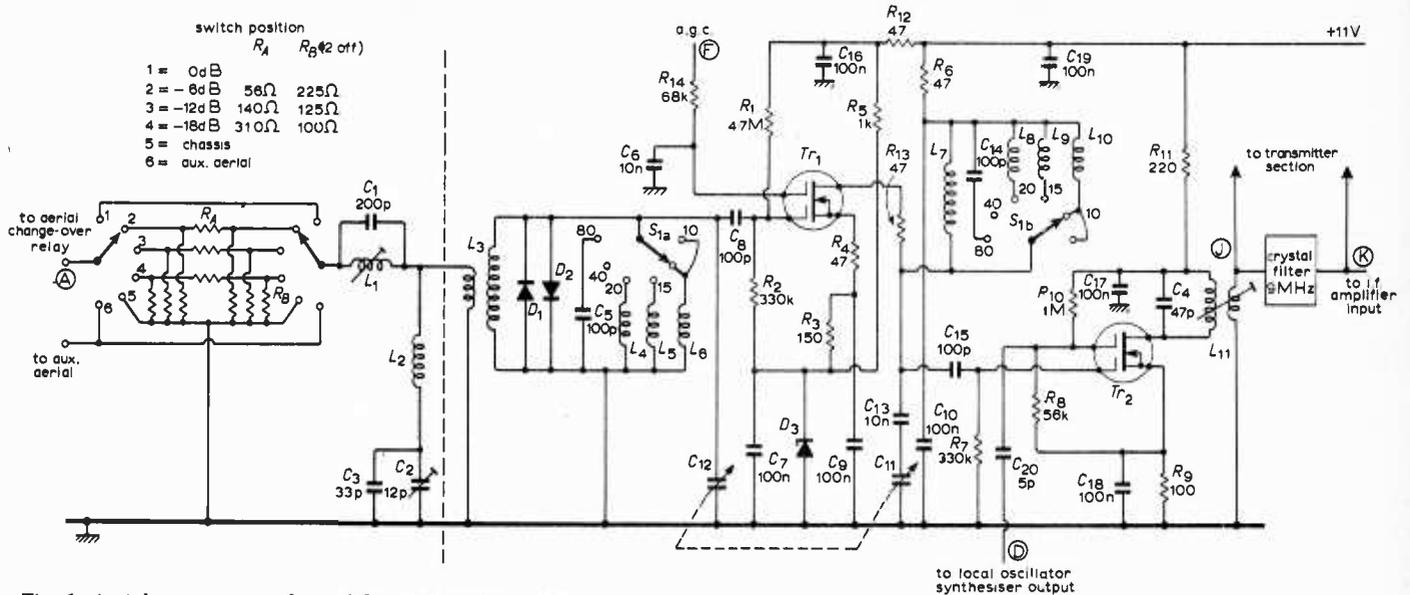


Fig. 6. Aerial attenuator, r.f. amplifier and receiver mixer.

have eliminated all unwanted signals before they reach the main i.f. amplifier.

The author has had considerable experience of the cascode connected bipolar amplifier and this arrangement (Fig. 7) has much to recommend it. The gain is high, being approximately the same as that of a common source neutralized stage. The considerable intended mismatch present between the first common emitter device and the common base connected second unit enhances stability considerably. The common base electrode is available for a.g.c. control.

It was decided to use single tuned transformers between stages as this would maintain a broad overall bandwidth and thereby allow the final shape to be determined by the filter characteristic even with considerable a.g.c. in operation.

Although two high-gain sections could

be designed to provide the required overall gain, previous experience suggested that three stages, incorporating a total of six transistors, would be better for stability reasons.

The required gain is spread equally between all three stages, but a.g.c. is applied only to the first two. The last stage is expected to cope with quite high signal levels and if a.g.c. were applied some distortion could occur. The use of capacitive impedance matching between stages was investigated, but found to exhibit rather large stray capacitive effects and the use of low impedance link coupling was found to reduce the earth path lengths and thereby help to stabilize the amplifier. During tests of the i.f. amplifier the a.g.c. performance was found to be as follows. With a change of input signal of 50dB below 200mV the output dropped by only 3dB and for a change

of 80dB the output varied less than 10dB. The amplifier had a maximum gain of 85dB (becoming unstable above this figure).

Construction: The mechanical layout is quite critical and it is suggested that the printed circuit layout (Fig. 8) be followed exactly. All component and, especially, coil leads must be as short as possible. It is advisable to use screened leads between the crystal filter and the i.f. amplifier input. The interstage screens shown in Figs 8 and 9 must be used. It is also advisable to mount the i.f. board either within an aluminium close fitting box or fit sheet screens around the amplifier.

Automatic gain control amplifier

The stage from which the a.g.c. is derived uses a single transistor (Tr_7 , Fig. 7) biased so that, with no signal, it is very nearly

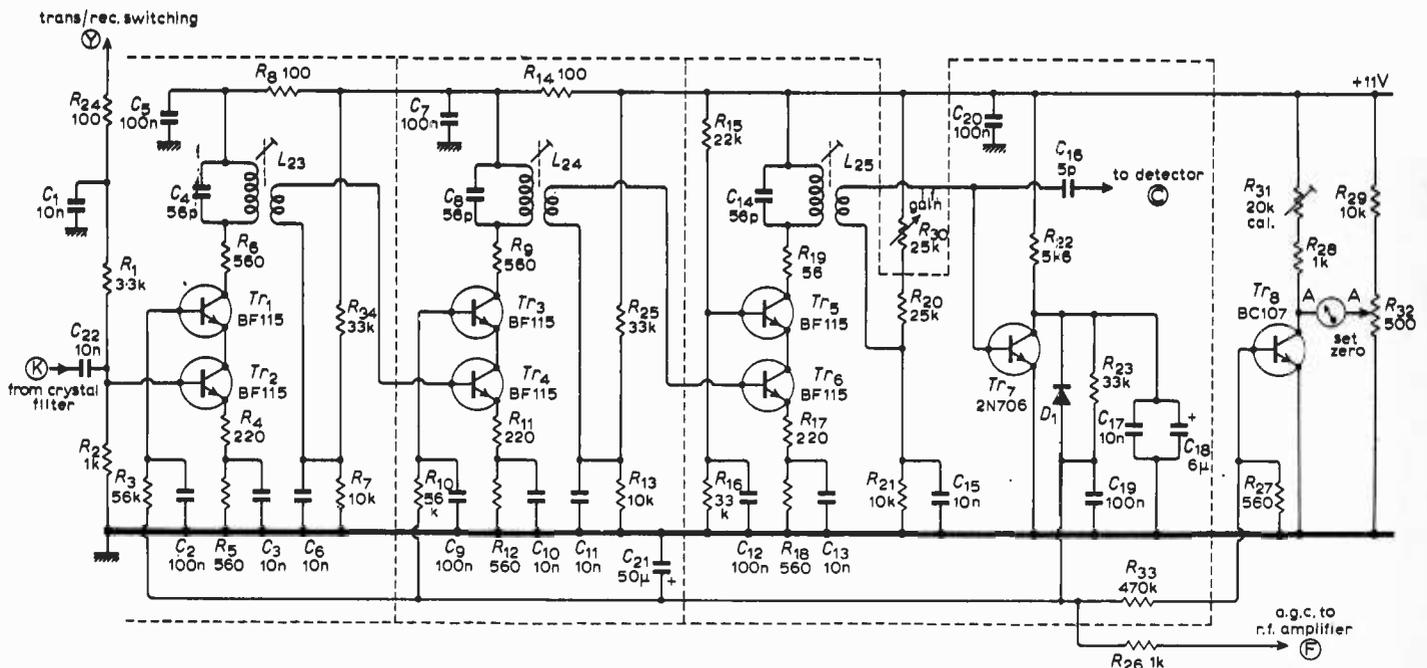


Fig. 7. The i.f. amplifier, a.g.c. amplifier and S meter. The leads to the meter (marked A) are connected to switches so that other parameters can be monitored. The circuit of the switching will be given in a later issue.

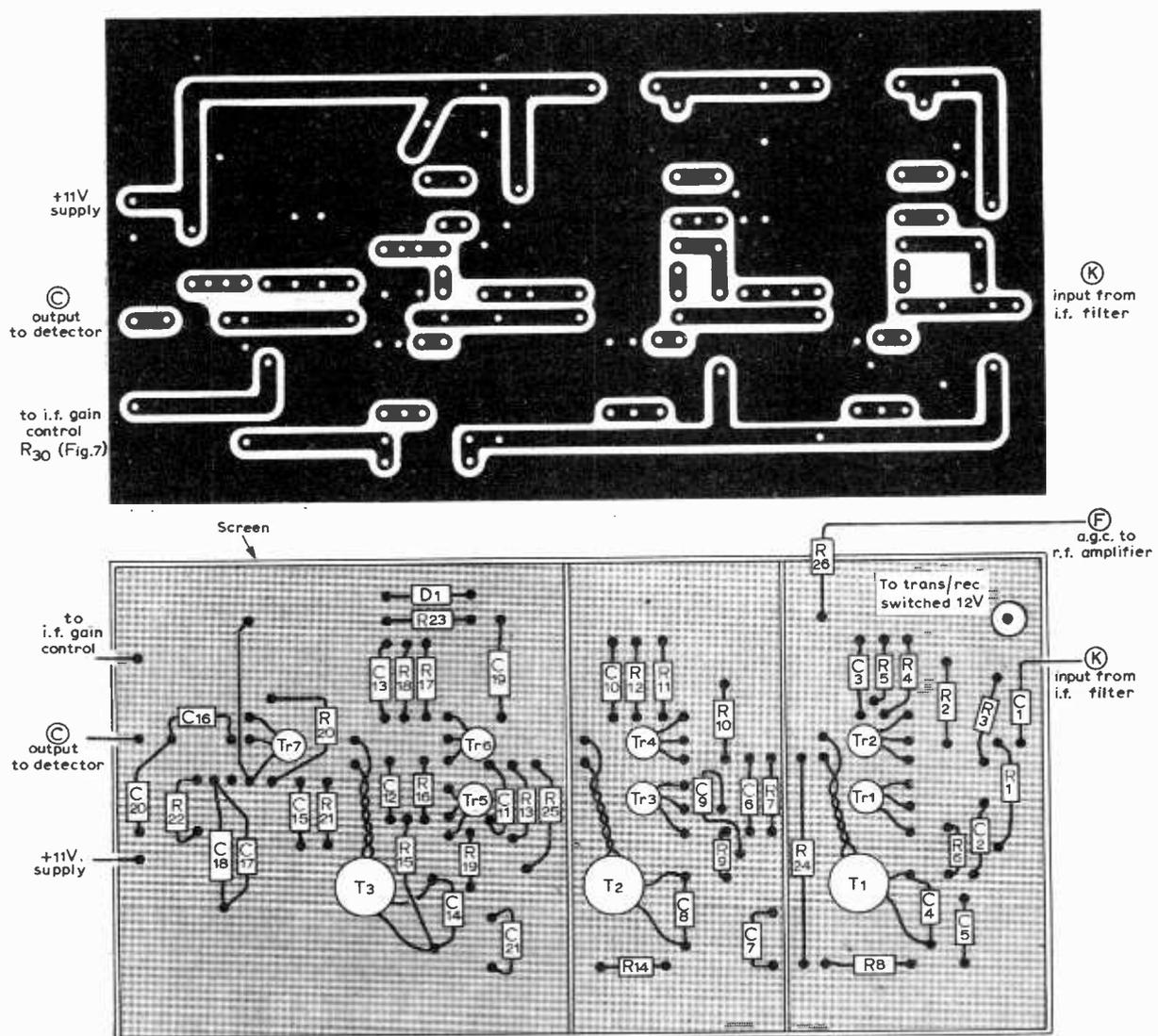


Fig. 8. Printed circuit board and layout for the i.f. amplifier shown actual size (130mm long).

switched off. As the signal increases so the average collector current also increases and therefore the mean collector voltage change is approximately proportional to the output of the i.f. amplifier.

For the reception of a single sideband transmission the normal fast attack, fast recovery, a.g.c. characteristic is useless. Because the transmission has no steady carrier wave the fast attack a.g.c. system tries to follow each syllable. One method of using a.g.c. with s.s.b. is to tailor the response to fast attack, slow decay. This has the effect of reducing the receiver's gain almost instantaneously, but delaying the release for the order of a second or so. This is achieved using an RC network where the R is replaced by a diode (D_1 , Fig. 7) connected in parallel with a resistor (R_{23}). The rise time is determined by the diode's on resistance and C_{19} , while the release time constant is $R_{23}C_{19}$ as the diode is reverse biased. With the values given in Fig. 7 the attack time is about 50ms and the decay time is about 1.65s.

During the discussions concerning the i.f. and r.f. amplifiers a note was made of the need to restrict the maximum gain to a stable usable value. This is most easily

achieved by adjusting the value of the resistor connected in series with the i.f. amplifier gain control R_{30} . The larger the value the lower the gain, but do take care to set R_{30} to minimum resistance when this adjustment is carried out.

Signal-strength meter

Although a signal strength meter is not absolutely essential it does add considerably to the operating convenience as well as helping with the alignment. If accurately calibrated it can be used to measure sideband and carrier suppression together with many aerial measurements. Numerous systems of calibration have been used over the years but the one originated by G. R. B. Thornley (G2DAF) and described by him in his booklet 'Communications Receivers' (R.S.G.B.) is preferred. This system takes as its reference a signal of $1\mu\text{V}$ across the 75Ω input impedance of the receiver as S1 and S2 as $2\mu\text{V}$, S3 as $4\mu\text{V}$ and so on, i.e. 6dB per S point. It must be realized that the signal voltage received at the aerial is as much a function of propagation conditions as the power originating from the transmitter. Therefore a signal strength report of S9 in the scale used by most amateurs has

no absolute value and is only S9 with reference to the background noise or alternatively to other signals.

The S meter circuit used in this receiver is a simple single-transistor voltmeter (Tr_8 , Fig. 7) with potentiometers for calibration (R_{31}) and setting zero (R_{32}). The voltmeter is connected to the main a.g.c. line and its output via the meter switch (not shown in Fig. 7), as the meter is used to monitor many other transceiver parameters.

Detector

The circuit (Fig. 10) used to demodulate the 9MHz s.s.b. waveform consists of two transistors (Tr_2 , Tr_3) in a balanced arrangement. The amplitude of the i.f. drive required is only about 10mV. This allows the use of a high-impedance (small value) feed capacitor which helps to reduce the amount of b.f.o. signal flowing back into the i.f. amplifier where it is liable to paralyse the a.g.c. system. The level of oscillator drive to the base of Tr_2 should not exceed 200mV. The oscillator and signal voltages at the collectors of the two stages are out of phase since the input voltages are cross-coupled to the respective bases and emitters. The third order intermodulation distortion

Continued on p. 330

Coil winding details for the receiver

All coils are wound on 0.25 inch (6mm) formers type CR2 (except where otherwise stated) with adjustable iron dust cores available from Home Radio Components Ltd, 240 London Road, Mitcham, Surrey, CR4 3HD.

The author found that a satisfactory method of anchoring the wire ends to the former was to wind several turns of ordinary sewing cotton over the wire. To complete the job a small quantity of shellac can be applied to the cotton. Enamel copper wire must be used in all cases where the coils are close wound. The centre taps are prepared by carefully scraping the enamel from the wire where appropriate and then soldering a short length of tinned copper wire to produce the centre tap.

i.f. trap

L_1 , 9MHz: 12 turns of 34 s.w.g. close wound with 200pF in parallel.
 L_2 , 9MHz: 46 turns of 34 s.w.g. with 40pF in

parallel.

signal frequency gate 1

L_3 , 40m: primary, 60 turns of 34 s.w.g. close wound; secondary four turns wound over earthy end of primary.
 L_4 , 20m: 28 turns of 34 s.w.g. close wound.
 L_5 , 15m: 15 turns of 28 s.w.g. close wound.
 L_6 , 10m: 19 turns of 22 s.w.g. close wound.

signal frequency drain

L_7 , 40m: 60 turns of 34 s.w.g. close wound.
 L_8 , 20m: 28 turns of 34 s.w.g. close wound.
 L_9 , 15m: 15 turns of 28 s.w.g. close wound.
 L_{10} , 10m: 19 turns of 22 s.w.g. close wound.

i.f. transformer

L_{11} , 9MHz: primary, 40 turns of 28 s.w.g. close wound on 0.25in (6mm) screened former No. CR12 or CR13; secondary, 7 turns of 28 s.w.g. wound over earthy end of primary.

variable frequency oscillator

L_{12} , 5MHz: 30 turns of 28 s.w.g. enamel copper wire close wound.
 L_{13} , 5-5.5MHz: primary, 50 turns of 34 s.w.g. close wound; secondary, 16 turns bifilar wound.

h.f. crystal oscillator

L_{14} , 11MHz: 30 turns of 34 s.w.g. close wound.
 L_{15} , 25MHz: 18 turns of 28 s.w.g. close wound.
 L_{16} , 32MHz: 9 turns of 26 s.w.g. wire spread to a winding length of 12.5mm.
 L_{17} , 32.5MHz: same as L_{16} .

i.o. synthesizer

L_{18} , 5.25MHz: 38 turns of 34 s.w.g. bifilar wound with 33pF and 4.7k Ω in parallel.
 L_{19} , 16.25MHz: 20 turns of 28 s.w.g. close wound bifilar with 15pF in parallel.
 L_{20} , 30.25MHz: 12 turns of 22 s.w.g. close wound with centre tap and 20pF in parallel.
 L_{21} , 37.25MHz: 11 turns of 28 s.w.g. spaced to 12.5mm wound with centre tap and 20pF in parallel.
 L_{22} , 37.75MHz: same as L_{21} .

i.f. amplifier

L_{23} , L_{24} , L_{25} , 9MHz: primary, 35 turns of 34 s.w.g. close wound; secondary, 8 turns wound over h.t. end of primary.

product detector

L_{26} , r.f.c., 10 μ H: 30 turns of 34 s.w.g. close wound with an iron dust core.

Components list

Aerial attenuator, r.f. amplifier, receiver mixer (Fig. 6)

C_1 200p, SM	C_{11} 6-75p, V
C_2 12p, T	C_{12} 10n, DC
C_3 33p, SM	C_{13} 100p, SM
C_4 47p, SM	C_{14} 100p, SM
C_5 100p, SM	C_{15} 100n, DC
C_6 10n, DC	C_{16} 100n, DC
C_7 100n, DC	C_{17} 100n, DC
C_8 100p, SM	C_{18} 100n, DC
C_9 100n, DC	C_{19} 100n, DC
C_{10} 100n, DC	C_{20} 5p, SM

SM - silver mica, DC - disc ceramic, T - trimmer, V - twin-gang variable 6-75p, Jackson type U101.

2.	R_A 56	R_B 225	} 0.5W moulded carbon
3.	140	125	
4.	310	100	

R_1 4.7M	R_8 56k
R_2 330k	R_9 100
R_3 150	R_{10} 1M
R_4 47	R_{11} 220
R_5 1k	R_{12} 47
R_6 47	R_{13} 47
R_7 330k	R_{14} 68k

Tr_1 } 3N201 (RCA) available from Henry's
 Tr_2 } Radio. Alternatives 3N187 or 40673
 D_{1-2} 1N916 or BAY38
 D_3 3.9V zener (BZY88 3.9V or BZY88 C3V9)

K.V.G. type XF-9B 9MHz crystal filter

i.f. amplifier, a.g.c. amplifier, S meter (Fig. 7)

C_1 10n, DC	C_{12} 100n, DC
C_2 100n, DC	C_{13} 10n, DC
C_3 10n, DC	C_{14} 56p, SM
C_4 56p, SM	C_{15} 10n, DC
C_5 100n, DC	C_{16} 5p, SM
C_6 10n, DC	C_{17} 10n, DC
C_7 100n, DC	C_{18} 6 μ , E
C_8 56p, SM	C_{19} 100n, DC
C_9 100n, DC	C_{20} 100n, DC
C_{10} 10n, DC	C_{21} 50 μ , E
C_{11} 10n, DC	C_{22} 10n, DC

SM - silver mica, DC - disc ceramic, E - electrolytic

R_1 3.3k	R_{18} 560
R_2 1k	R_{19} 56
R_3 56k	R_{20} 25k
R_4 220	R_{21} 10k
R_5 560	R_{22} 5.6k
R_6 560	R_{23} 33k
R_7 10k	R_{24} 100
R_8 100	R_{25} 33k
R_9 560	R_{26} 1k
R_{10} 56k	R_{27} 560
R_{11} 220	R_{28} 1k
R_{12} 560	R_{29} 10k
R_{13} 10k	R_{30} 25k, P
R_{14} 100	R_{31} 20k, P
R_{15} 22k	R_{32} 500, P
R_{16} 33k	R_{33} 470k
R_{17} 220	R_{34} 33k

P - potentiometer

Tr_{1-6} BF115 (Alternatives, TIS48, 2N706).
 Tr_7 2N706 (TIS48).
 Tr_8 BC107 (BC108, BC109).
 D_1 OA91 (CV448 or any Ge point contact).

b.f.o., detector, a.f. amplifier (Fig. 10)

C_1 150p, SM	C_{13} 1n, DC
C_2 10n, DC	C_{14} 1n, F
C_3 10n, DC	C_{15} 2.5 μ , E
C_4 10n, DC	C_{16} 80 μ , E
C_5 1n, F	C_{17} 80 μ , E
C_6 1n, F	C_{18} 125 μ , E
C_7 10p, SM	C_{19} 1,000 μ , E
C_8 10n, DC	C_{20} 1,000 μ , E
C_9 10n, DC	C_{21} 25 μ , E
C_{10} 12 μ , E	C_{22} 2-10p, T
C_{11} 10n, DC	C_{23} 2-10p, T
C_{12} 1n, F	

SM - silver mica, DC - disc ceramic, F - feedthrough, E - electrolytic, T - trimmer.

R_1 47k	R_{14} 470
R_2 22k	R_{15} 0.22, W
R_3 220	R_{16} 0.22, W
R_4 2.2k	R_{17} 33k
R_5 2.2k	R_{18} 1k
R_6 22k	R_{19} 2.2k
R_7 2.2k	R_{20} 1.8k
R_8 6.8k	R_{21} 100
R_9 1k	R_{22} 680
R_{10} 1k	R_{23} 3.3k
R_{11} 2.2k	R_{24} 500, P
R_{12} 22k	R_{25} 5k, P
R_{13} 6.8k	

W - 1W wire wound, P - potentiometer

R_1 10k	R_{12} 33k
R_2 10k	R_{13} 10k
R_3 1k	R_{14} 470
R_4 560	R_{15} 330
R_5 100	R_{16} 180
R_6 22k	R_{17} 1M
R_7 47k	R_{18} 56k
R_8 430	R_{19} 220
R_9 1k	R_{20} 3k, P
R_{10} 680k	R_{21} 25k, P
R_{11} 25k	

P - potentiometer

Tr_{1-3} BC107 (TIS48)
 Tr_{4-5} 3N201 (3N187, 40673)
 D_{1-2} 1N916 (1N914, BAY38)
 D_3 BA142 (see text).

local oscillator synthesizer consisting of h.f. crystal oscillator, v.f.o. and mixer (Fig. 11)

C_1 6-75p, J	C_{15} 100n, DC
C_2 5-30p, T	C_{16} 5p, SM
C_3 10p, N	C_{17} 10n, DC
C_4 100p, SM	C_{18} 47p, SM
C_5 150p, SM	C_{19} 25p, SM
C_6 3n, SM	C_{20} 10p, SM
C_7 1n, SM	C_{21} 50p, SM
C_8 10n, DC	C_{22} 100n, DC
C_9 50p, SM	C_{23} 100n, DC
C_{10} 10n, DC	C_{24} 10n, DC
C_{11} 10n, DC	C_{25} 100n, DC
C_{12} 2-10p, T	C_{26} 5p, SM
C_{13} 68p, SM	C_{27} 56p, SM
C_{14} 68p, SM	

SM - silver mica, DC - disc ceramic, J - Jackson U101 single gang variable capacitor, T - trimmer, N - negative temperature coefficient, adjust value for minimum frequency drift.

Note. Capacitors wired across the mixer coils (L_{18} , L_{19} , L_{20}) are detailed in the coil winding table.

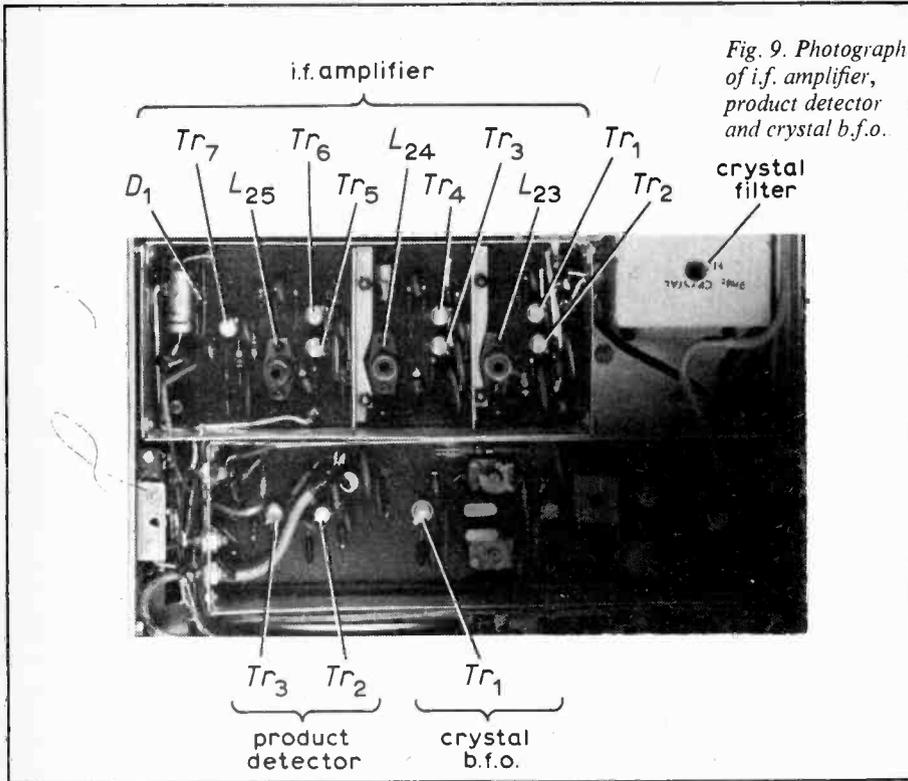


Fig. 9. Photograph of i.f. amplifier, product detector and crystal b.f.o.

switch to be mounted away from the b.f.o. From the circuit it will be noted that a sample of the b.f.o. output is taken to the carrier amplifier which will be described later in a section dealing with the transmitter circuitry.

Audio amplifier

The circuit (Fig. 10) consists of a transformerless push-pull transistor output stage (Tr6, Tr7) driven by two class A common emitter stages (Tr4, Tr5). A good deal of negative feedback is applied over the last two stages to stabilize the d.c. conditions as well as to reduce output non-linearity. There are no temperature compensating diodes or networks but with silicon transistors, thermal runaway is unlikely unless the transceiver is used in extremely high temperatures. The prototype transceiver is very compact and runs quite hot without any ill-effects.

It is essential that the following procedure is carried out to set the bias level of the output stage, or considerable cross-over distortion may occur. The no-signal idling current flowing through the output transistors is determined by the values of resistors R22 and R24. A general purpose multimeter set to read about 50mA f.s.d. should be connected in series with the 11V supply to the amplifier. With no input signal and R24 set to minimum resistance a note of the current should be made. The resistance of R24 should now be progressively increased until the current increases by about 1 to 3mA. It is most important not to exceed this current as thermal runaway may occur. The amplifier is capable of delivering about 200mW, which is more than adequate.

Local oscillator synthesizer

The local oscillator synthesizer consists of a v.f.o. with a 500kHz tuning range feeding a mixer/amplifier. An h.f. overtone crystal oscillator also feeds the mixer. The v.f.o. has a range of 5 to 5.5MHz (as discussed in Part 1) with a particularly clean output

should be about -50dB with the fifth order less than -70dB relative to the required output.

The r.f. choke connected in the collector circuit of the detector is not critical in value, only being required to exhibit a high reactance at 9MHz and a low reactance over the audio range. The author wound 35 turns of 30 s.w.g. wire on to one of the 6mm (0.25in) formers used elsewhere in this project and screwed an iron dust core fully into the coil.

Common carrier and beat frequency oscillator

This is the first of the stages which are common to both transmit and receive. The

crystal controlled b.f.o. produces two slightly different frequencies, allowing the two sidebands to be demodulated. The frequencies chosen lie about 1.5kHz each side of 9MHz, the final adjustment being carried out using the small trimmers C22 and C23 in Fig. 10. The final frequencies are set to 20dB down either side of the crystal filter. As appropriate crystals are supplied with each KVG filter no extra crystal details are necessary. Anyone intending to order separate crystals should specify series resonant units and in the author's case the frequencies were 9.001MHz and 8.998MHz. The rather elaborate diode switching system used for sideband selection is to allow the

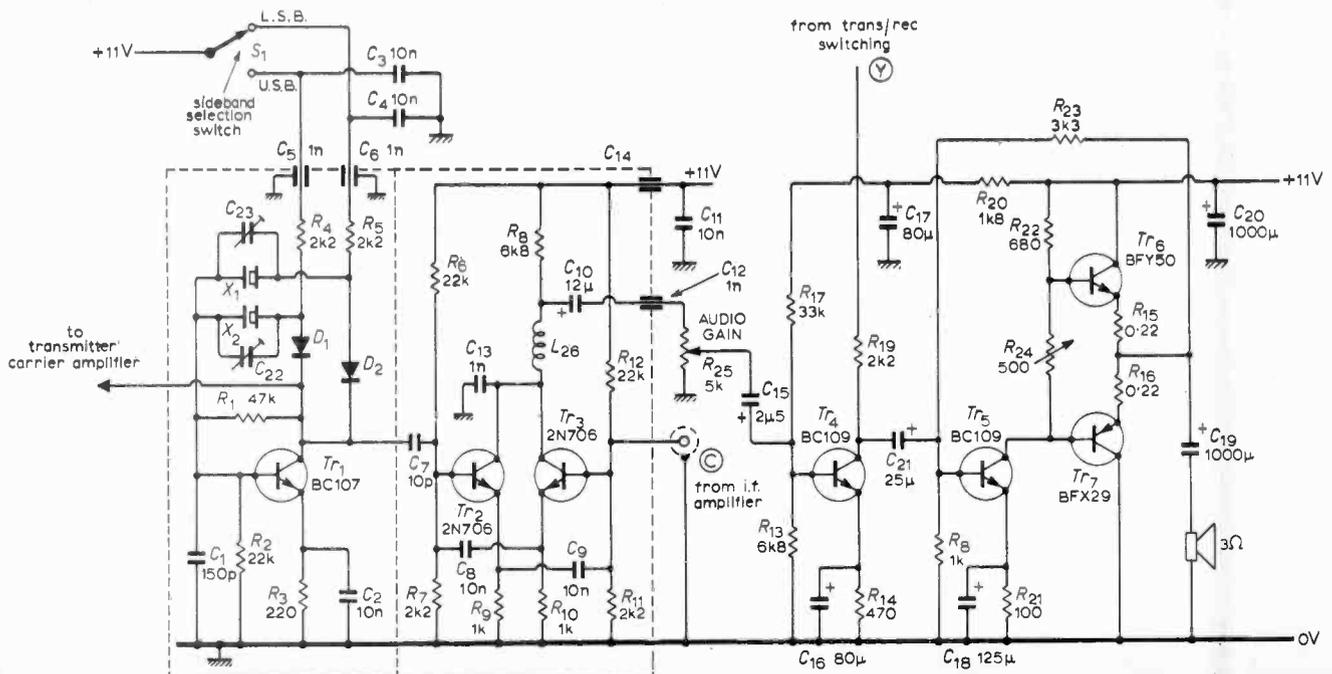


Fig. 10. Circuit of b.f.o., product detector and audio amplifier.

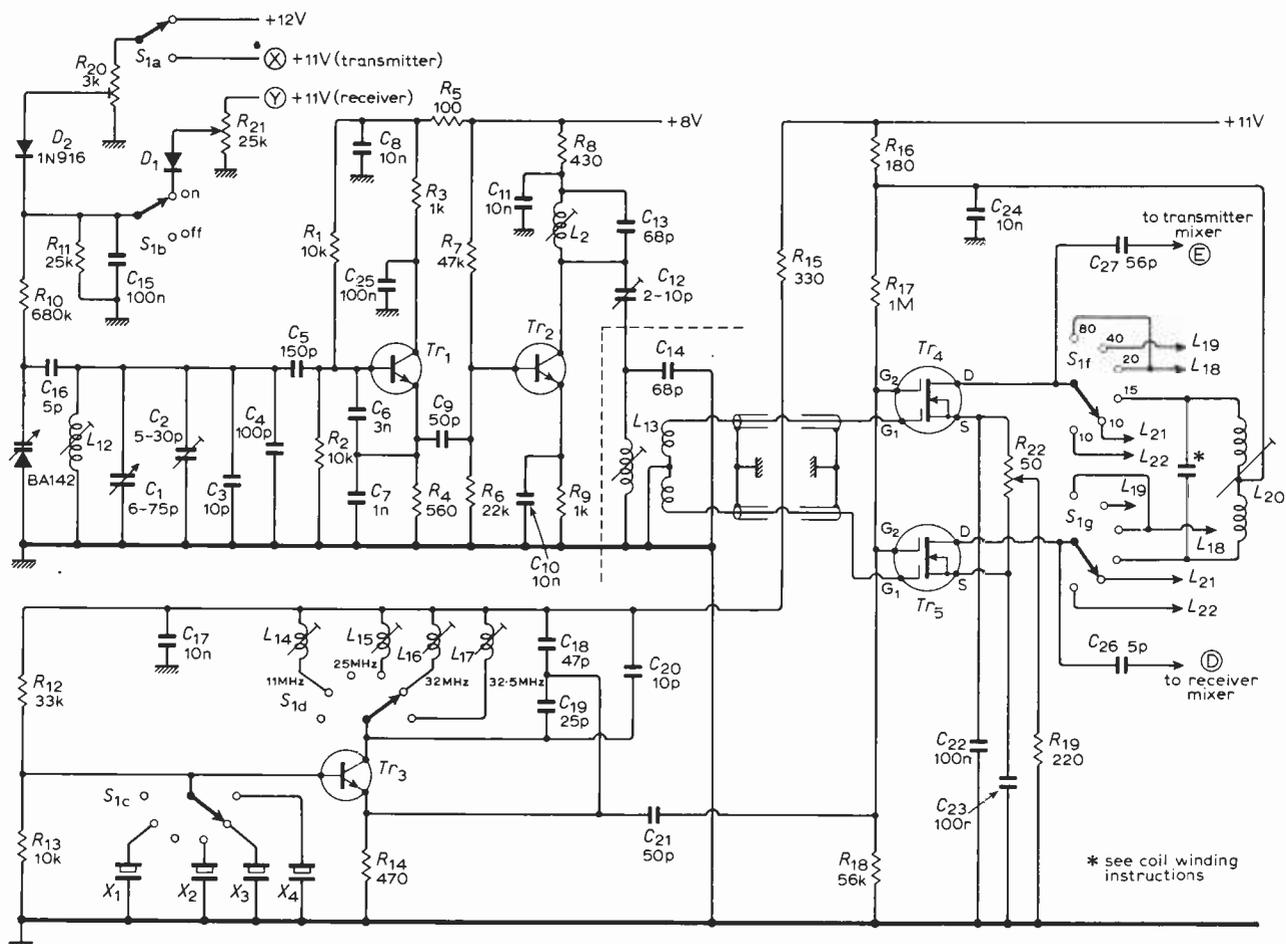


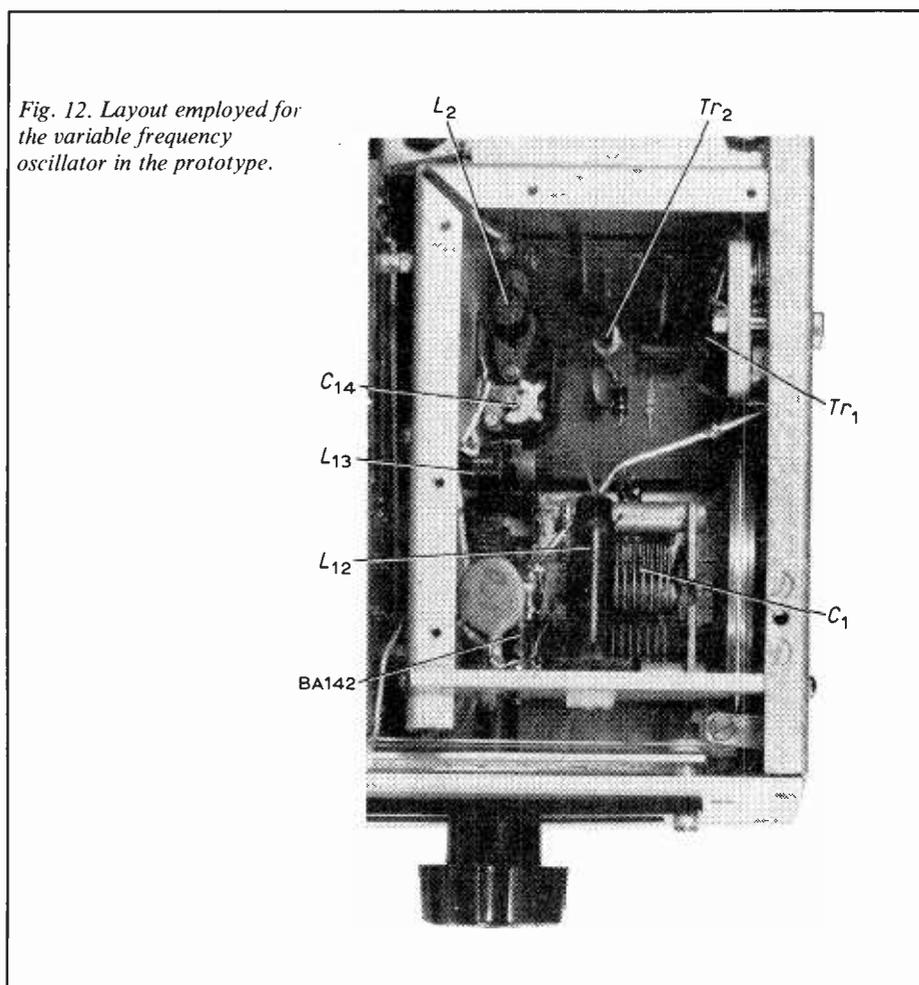
Fig. 11. The local oscillator synthesizer consisting of a variable frequency oscillator, a crystal oscillator and a balanced mixer.

waveform. The v.f.o. output is passed to the mixer, which doubles as an amplifier when the transceiver is in operation on 80 and 20m. On 40, 15 and 10m it reverts to a balanced mixer and is additionally fed from the h.f. crystal oscillator. The crystals used are third overtone series resonance types with the exception of the 40-metre unit which is a fundamental series resonant crystal of 11MHz. We will now look at each section of the local oscillator synthesizer in turn. The circuit is given in Fig. 11.

Variable-frequency oscillator: The v.f.o. consists of three main items. First the basic linear frequency oscillator (Tr_1) followed by an isolation amplifier (Tr_2), which feeds a wideband coupler the push-pull output of which is connected to the two gate electrodes of the mixer.

The oscillator circuit (Tr_1) is quite straightforward, using a single bipolar silicon transistor in a Colpitts arrangement. The L to C ratio is chosen to incorporate relatively large values of C in order to swamp the capacitance variations which occur within the transistor. This, with the addition of a small amount of negative temperature coefficient capacitance, reduces the overall frequency drift to much less than 1kHz. A reasonably linear tuning range is achieved using a semicircular vanned linear tuning capacitor. The linearity can be improved by cutting away a small section of the moving vanes at the minimum capacitance end. The prototype was accurate to 1kHz over the 500 kHz of the range.

Fig. 12. Layout employed for the variable frequency oscillator in the prototype.



When S_{1b} is closed R_{21} (the i.r.t. control) is connected via D_1 . Potentiometer R_{21} allows the operator to shift the receive frequency by about 5kHz by varying the bias on the BA142 varicap diode. Almost any other variable capacitance diode can be used (a reverse biased 1N916 might do) if the value of the 5pF coupling capacitor is altered accordingly. Switch S_{1c} disconnects R_{20} from the permanent 12V line and connects it to the transmitter supply line. This means that when on receive the varicap obtains its bias via D_1 and on transmit via D_2 . The variation of capacitance with reverse voltage is nonlinear, but if a log potentiometer is used for R_{21} an approximately linear frequency range is obtained. By carefully adjusting R_{20} it is possible to set the transmit and receive frequencies to coincide when R_{21} is set to the middle of its travel.

The alignment of the wideband coupler will be dealt with in part 4 of this article. A photograph of the v.f.o. appears in Fig. 12.

Mixer: This stage acts as a mixer when the h.f. crystal oscillator is in operation and as an amplifier for the v.f.o. when the transceiver is operating on the 80 or 20m bands.

A balanced arrangement of dual gate f.e.t.s is used in preference to single gate devices (Fig. 11) because of the extra local oscillator isolation and wider dynamic range available. The d.c. conditions were chosen to suit mixer operation (the gain in this mode is about 10dB less than when the devices are used as straight amplifiers to level the output from band to band).

It will be noticed that a balance potentiometer (R_{22}) is used in the source circuit of the mixer. This was intended to optimize the rejection of the l.o. in the output, but in view of the small effect it has, some constructors may decide to replace it with a short-circuit, i.e. both source electrodes and the top end of the source bias resistor connected together.

The synthesizer output to the receiver and transmitter signal mixers are taken from the two drains.

High-frequency crystal oscillator: This oscillator is required to produce a number of switched, highly stable oscillations within the range 11–38MHz.

In the circuit the 11MHz crystal is a fundamental type while all others are of the third overtone variety. Circuit operation is slightly different from some other circuits, namely in that the transistor will give an output which locks only when the free running and crystal frequencies coincide. One word of warning! The transistor type seems to be quite critical and the very high f_T devices such as BFY90 must not be used as they oscillate so eagerly that there is difficulty in getting them to lock to the crystal frequency. Suitable transistor types are BC107 and TIS48 with other 400MHz f_T devices as possible alternatives.

When ordering crystals, type of operation (i.e. fundamental or overtone) and final output frequency must be stated.

Construction: The frequency synthesizer is the most important section of the whole

transceiver from the stability point of view. Frequency stability depends on good solid mechanical construction. Thick, rigid short lengths of connecting wire must be used; in the prototype most of the wiring was in printed circuit form. The printed boards being attached to the chassis using many substantial nuts and bolts.

The dial mechanism is also very important and it is advisable to use a reputable manufactured unit, for example a full-vision Eddystone type. The dial used by the author was home made with the exception of the worm gear and cog which were kindly donated by a fellow enthusiast from an old TA12 transmitter. Similar gears are available from Bond's Model Engineering Shop, 186 Tottenham Court Road, London.

Experience has shown that almost any coil core material can be modulated by large magnetic fields and for this reason it is advisable to mount the v.f.o. coil well away from the mains transformer. Even with this precaution ferrite cores must not be used. Iron dust will work well, but if it is found to be necessary to mount the coil close to the transformer then the number of turns on the coil must be increased and an 0 B.A. brass slug used for minimal final frequency adjustment. The magnetic effect shows up as an unexplained hum or frequency modulation at 50Hz superimposed on the suppressed carrier and in severe cases it can be audible on signals being received.

Temperature stability is vital and as in some cases it is impossible to maintain the v.f.o. at a constant temperature the only alternative is to use negative or possibly positive coefficient capacitors. The idea is that when the v.f.o. tends to drift upwards with an increase in temperature then an amount of negative temperature coefficient capacitance shunt connected across the tuned circuit will tend to correct the drift.

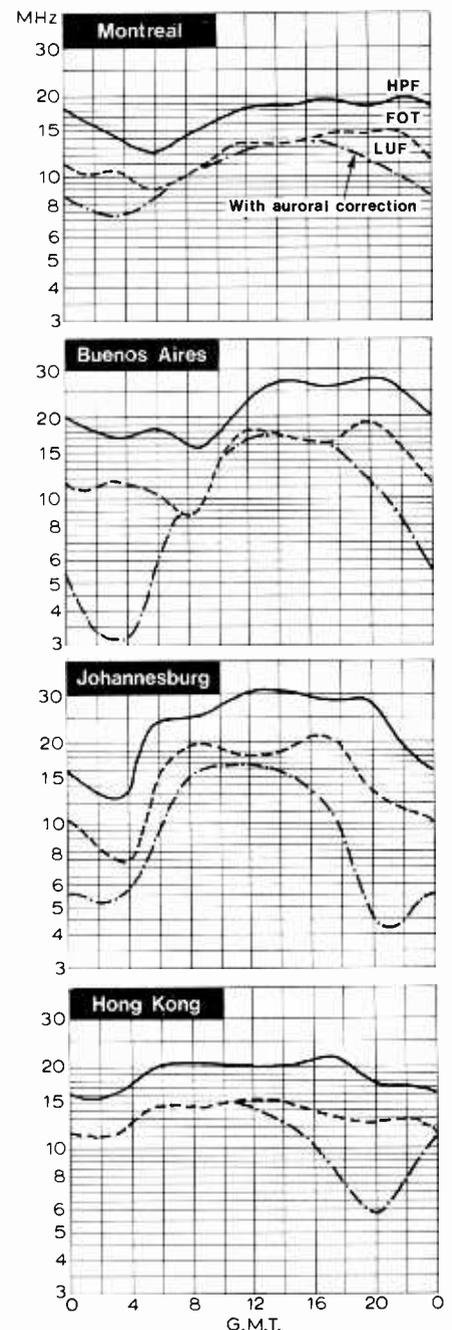
The h.f. crystal oscillator is less critical but even here it is advisable to construct solidly.

The balanced mixer must maintain its balance and for this reason a symmetrical layout is used. The l.f. coils are bifilar wound to assist the maintenance of balance, but on the higher frequencies this type of construction is more difficult and therefore normal centre-tap coils are used. Some misbalance results as the output coupling capacitors have different values and are connected to alternate drain electrodes, but this asymmetry is small and does not warrant further circuit complications.

(To be continued)

H.F. Predictions — July

Measured values of the solar index IF2 for March and April are 100 and 89, respectively. These are slightly higher than the corresponding 1971 values and double those predicted three months in advance. As a result the charts for months since then have been rather pessimistic in their predicted HPFs (highest probable frequencies) and FOTs (optimum traffic frequencies). This month's charts assume that the high level of solar activity will continue. Magnetic disturbed day rating was reached on twelve occasions during May/June indicating that the next disturbed period would probably be at the end of July.



Towards Better Control of Small D.C. Motors

by P. L. Hollingberry, M.A., A.M.I.E.E.

The circuit was designed for controlling model locomotives but can be used in any application which requires that the speed of a small d.c. motor remains constant under varying loads. The supply to the motor is pulsed; the rate the pulses are applied is determined by comparing a potentiometer-set voltage specifying required speed with the motor's back e.m.f. The back e.m.f. is measured between pulses when there is no supply on the motor. The idea is based on a circuit which appeared in *Control* (Oct. 1967).

The use of electronics to improve the control of very small permanent-magnet motors has intrigued the model railway fraternity ever since transistors first became available to the hobbyist. As the scale of the models has diminished, so the scale of the problem has increased and called for greater ingenuity. The electronic regulator to be described was developed for an N-gauge model railway (scale-1:148), but the improvement in performance the circuit gives over conventional circuits has been so substantial that the idea is likely to be of value to all who need to control the speed of small d.c. motors under varying loads.

Earlier solutions

Rheostat control: The classical method of controlling a small motor is to wire a variable resistance in series with the supply. This allows continuous control of

the armature current, and hence of torque. Trouble occurs at low speeds as the torque is substantially independent of back e.m.f. (whose mean value is proportional to motor-speed) and the speed therefore becomes very dependent on load.

Pulse controllers: A popular and simple improvement over the variable resistance method consists of allowing unsmoothed pulses of mains ripple from a rectifier to reach the motor. This technique gives better low-speed performance because the alternating component, rich in harmonics, contributes greatly towards overcoming the high 'stiction' of small motors. The model can consequently be run at a lower speed without stalling than is possible if a smooth voltage were applied. Pulses can be generated by more sophisticated means than this, and a great deal has been written about the best shape and p.r.f. for

the job. However, this system does not provide any improvement in load regulation.

Emitter follower: A different line of attack has been the use of Darlington or compound emitter followers to provide a low impedance variable-voltage source to power the motor. This is very effective for motors with low armature winding resistance, because when the back e.m.f. becomes less than the source voltage a large current will flow in such a direction as to tend to make the two voltages equal. Unfortunately the armatures of many small motors for models are deliberately wound with many turns of thin wire having a high resistance in an effort to cause them to run slowly, so that more than half the applied voltage is dropped across the armature resistance. The value of an emitter-follower controller is, in this instance, much reduced.

Negative resistance: An ingenious and moderately successful attempt to deal with the armature resistance was described by H. M. Butterworth¹, who arranged to balance it out in a Wheatstone bridge circuit and was able to isolate the back e.m.f., compare it with an e.m.f. proportional to the desired speed, and amplify the difference to drive the motor in such a direction as to make the two e.m.f.s equal. This was equivalent to connecting an equal negative resistance in series with the armature resistance. The circuit has therefore to be matched to a particular motor, and on small two-rail systems where electrical pick-up is made from the running-wheels its performance is reduced by the extra resistance due to dirt and corrosion on the rails. It is also surprisingly difficult to measure the effective armature resistance by a direct method because in the normal type of 3-pole motor the commutator segments are arranged to switch in alternately two coils in parallel, followed by two in series with each other and in parallel with the third, as the rotor rotates.

A solution

The circuit of Fig. 1 takes advantage of the best aspects of earlier work in this field and overcomes all the problems which have been mentioned above. Drive to the motor is by voltage pulses from a free-running multivibrator, and during each pulse the full voltage of the supply is switched across the motor. In the interval between each pulse and the next, the back e.m.f. is monitored and compared with a reference; the result of the comparison determines whether the subsequent pulse is accelerated or retarded, thereby tending to correct any error in the motor speed.

K. C. Johnson² has described a similar method for controlling an electric drill but using unsmoothed a.c. and a thyristor. In this application the multivibrator was preferred because of its greater flexibility in respect of p.r.f. and duty cycle. Because the back e.m.f. is measured when no current is flowing, the regulator will accommodate wide variations in circuit

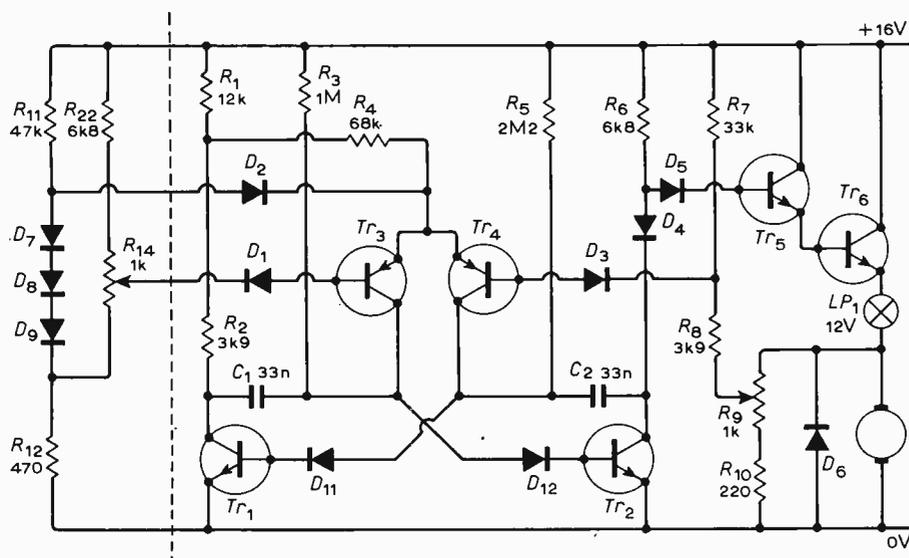


Fig. 1. Circuit of the d.c. motor speed controller. Potentiometer R_{14} is the speed control.

resistance, without requiring to be matched to the load.

In operation the circuit has added strikingly to the realism of a model railway, allowing trains to be accelerated and decelerated as smoothly and gradually as any operator could wish. Because of the closed-loop nature of the circuit, models exhibit no tendency to stall or race on gradients, and when left running trains maintain a steady speed regardless of track conditions and load (this feature may not appeal, of course, to those enthusiasts who favour realistic prototype steam locomotive behaviour!). Although one preset is provided the circuit is not at all critical and once set up will certainly accommodate any model of the same gauge without adjustment.

In Fig. 1 Tr_1 and Tr_2 are the multivibrator transistors. Capacitor C_1 determines the duration of the pulse and C_2 the interval between pulses. During the interval between pulses Tr_3 and Tr_4 form a long-tailed pair differential comparator, comparing the reference voltage set on the potential divider R_{12} , R_{14} and R_{22} (R_{14} is the speed control) with the back e.m.f. of the motor (now acting as a tachometer generator). The back e.m.f. appears at the cathode of D_3 after being attenuated by R_9 and R_{10} , and level-shifted by R_7 and R_8 . During the interval Tr_2 is conducting and Tr_1 is cut off, C_2 is discharging through R_5 and Tr_4 ; if the back e.m.f. is too low Tr_4 takes most of the tail current and discharges C_2 quickly, thereby increasing the mark-space ratio. If the back e.m.f. is too great, the reverse happens, Tr_4 takes little of the tail current and C_2 discharges slowly through R_5 .

During the pulse the voltage across the motor is greater than the back e.m.f. and so D_3 and Tr_4 are non-conducting and all the tail current passes through Tr_3 , discharging C_1 . However, since the tail (R_4) is not taken to 16V, but to the junction of R_1 and R_2 , the voltage across it during the pulse is much less than during the interval, and is dependent to a much greater extent on the reference voltage set on R_{14} . Resistors R_1 and R_2 are chosen so that at the maximum setting of R_{14} , no current flows in R_4 , and C_1 discharges slowly through R_3 . At low settings of R_{14} , C_1 discharges about twice as quickly, partly through R_3 and partly through Tr_3 ; consequently the pulse duration is varied directly by the speed setting in anticipation of a greater mark-space ratio, preventing excessively wide variations in p.r.f. over the operating range.

When R_{14} is set to minimum D_2 conducts, causing enough current to pass through Tr_3 to discharge C_1 very quickly, generating very short pulses. The object of this is to prevent the locomotive from inching forward when the speed is set to zero. The short pulses may just be heard as a low buzz from the motor. Resistors R_7 and R_8 have been chosen so that at this setting Tr_4 does not conduct, and the p.r.f. is minimum.

The diode D_4 is needed to prevent the multivibrator from synchronizing with the mains ripple on the supply or switching off

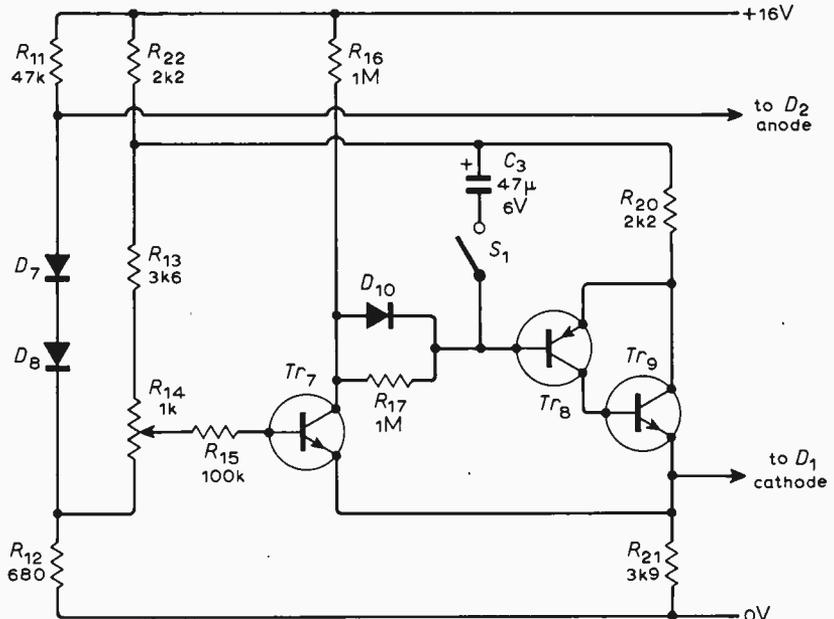


Fig. 2. Add-on circuit which gives a motor a controlled rate of acceleration and deceleration.

prematurely owing to the effect on the supply of the sudden load following the onset of a pulse. In the absence of this diode any fall in voltage at either end of R_6 during the pulse would be transferred through C_2 to the base of Tr_1 , switching the latter off and causing the pulse to disappear. If other circuits are to be fed from the same rectified supply, a second diode should be similarly located in the Tr_1 collector circuit.

The pulse is coupled to the load via the Darlington emitter follower Tr_5 , Tr_6 , and the lamp LP_1 , which provides both overload protection and indication. An incandescent lamp filament does not obey Ohm's Law, but conforms approximately to a square law, and so at half its rated current drops only one-quarter of its rated voltage, leaving the remainder to power the locomotive.

The quenching-diode D_6 is provided to 'catch' the negative-going spike at the end of each pulse caused by the inductance of the armature winding, which tries to maintain the current set up during the pulse.

Acceleration and deceleration control

It will not be denied that the circuit, Fig. 1, is a good deal more complicated than most model train controllers, but its performance really does justify the extra expense. In any case the cost of the circuit in Fig. 1 is still unlikely to be much more than that of the mains transformer needed for any such unit. However, for those with a positive passion for realism an auxiliary circuit is described (Fig. 2) which will control the rate of acceleration and braking to simulate the inertia of the more massive, under-powered real thing. A simple resistor and capacitor will not do this because of the exponential characteristic, but by amplifying the curve and using a small part of it, such a network may be used to effect nearly uniform acceleration and deceleration.

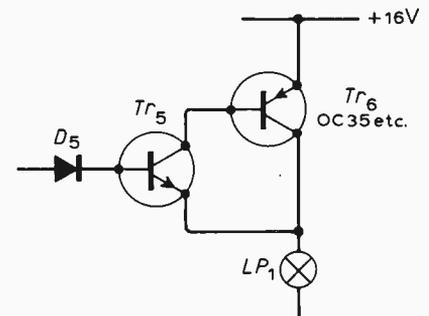


Fig. 3. If wished the n-p-n output transistor of Fig. 1 can be replaced with a p-n-p type.

Much satisfaction is to be derived from setting up the speed on the controller and watching the train slowly accelerate away as if under its own volition.

The circuit uses a composite transistor, Tr_8 , Tr_9 , to avoid loading the relatively small timing-capacitor C_3 . Any change in the setting of R_{14} causes Tr either to bottom or to turn off, and C_3 then discharges or charges very slowly through R_{17} or R_{16} and D_{10} . The slowly changing voltage appears on the emitter of Tr_8 , and a similar voltage changing slowly in the opposite direction is present at the emitter of Tr_9 , because nearly identical currents flow in R_{20} and R_{21} . When the voltage on R_{21} reaches about 600mV, below that set on R_{14} , Tr_7 resumes its initial state of conduction. If switch S_1 is left open, the speed will respond immediately to the setting of R_{14} .

Practical points

Transistors Tr_1 , Tr_2 , Tr_3 and Tr_9 must be very high gain silicon transistors having an h_{fe} of not less than 200 (BC109 or 2N930); Tr_3 , Tr_4 , Tr_7 and Tr_8 should all be silicon, but no other special characteristics are required.

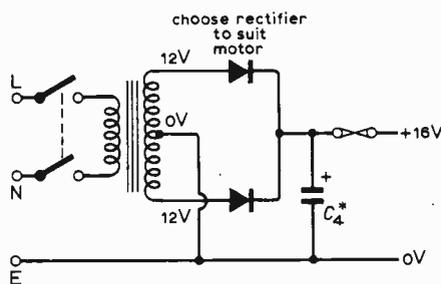
The choice of Tr_6 and the lamp are closely linked with the maximum load intended. The lamp should be for 12V

operation and the range of motor car lamps available give one a wide choice. For small models taking about 250mA when running uphill with a train, use a 2N1613 or one of the ZTX300 series (or any 500mA transistor) with a 5 or 6W lamp (BFS254, BFS980, side-lamp bulbs). For locomotives requiring 500mA the BFY50 series of transistors (rated 1A) is available, and a 12W lamp is called for (which might be difficult to obtain). If this is the case use a 21W lamp (BFS380, BFS382) and a power transistor although, because it is either off or dropping only a volt or two, no heat-sink will be required. For those wishing to use a p-n-p transistor such as OC35, a compound emitter follower circuit is given in Fig. 3. Few trains will require as much as 1A even with double-heading, but it is as well to bear in mind that as motors age and lose their magnetism they consume more current.

The diodes should all be silicon small-signal devices, such as 1N914.

Resistors R_7 , R_8 , R_{12} , R_{13} , and R_{22} must be 5% tolerance or better, but the remaining resistors may be 20%; R_9 and R_{14} should be linear, and preferably wirewound.

A suitable power supply circuit is given in Fig. 4; C_4 should be rated at least 20V and have a value not less than that given in the table.



max mean motor current Amps	T_{r_6} I_{CM} rating Amps	lamp rating Watts	C_4^* μF
0.25	0.5	5 or 6	500
0.5	1.0	12	1,000
1.0	2.0	21	2,000

Fig. 4. Circuit of a suitable power supply.

The use of the preset R_9 will become obvious when the unit is tried out, and no instrument is needed to set it up. It merely allows the working range of the speed control R_{14} to be adjusted to suit the user and the general characteristics of the gauge of model in use. Larger gauges,

though still nominally 12V, will tend to produce larger back e.m.f.s than smaller ones. Some operators, too, are known to run their models at a maximum speed which would put the advanced passenger train to shame; but most will want to restrict maximum speed to safe limits and R_9 allows both to have their way.

The value of C_3 , which determines acceleration and braking rates, is a compromise between pedantic realism and what is practical on a small layout. To obtain differing rates of acceleration and braking, vary R_{17} or R_{16} .

Specification

Dynamic pulse duration: variable between 10 and 20ms

Idling pulse duration: 1.0ms

Minimum pulse interval: 2.5ms

Maximum pulse interval: 50ms

Maximum back e.m.f. accommodated: variable between 1.5 and 12V

Acceleration and braking (Fig. 2 only): start to maximum and maximum to stop in 12s

Nominal operating voltage: 12V

REFERENCES

1. H. M. Butterworth, "Speed Control for D.C. Model Motors", *Wireless World*, September 1967, p.440.
2. K. C. Johnson, "Thyristor Speed Control for Electric Drill Motors", *Wireless World*, July 1967, p.328.

Fixed Aerial for Satellite Reception

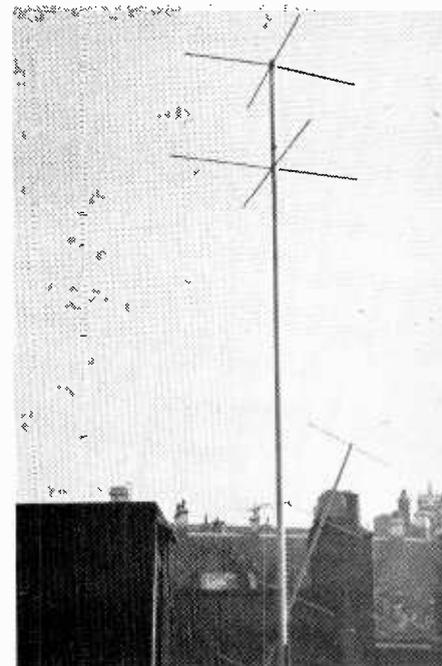
by J. M. Osborne*

By a suitable choice of components and dimensions, acceptable signal strengths are obtained with this aerial from a weather satellite at an elevation of more than 20°. Although a fixed aerial of such low directional properties is inevitably less efficient than a conventional highly directional tracking aerial, it opens up an entirely new field for inexpensive, simple-to-man, weather picture stations.

The fixed aerial not only removes much of the expense but makes the choice of site much less demanding as the aerial is as easy to erect as a domestic television dipole. With a fixed aerial, it is even conceivable that the equipment could switch on automatically when the satellite signal strength reaches an acceptable level. Provided time information was recorded the geographical position of pictures could be determined subsequently from a previously prepared table. It should be possible to make such a station automatic. It would not, of course, replace the existing meteorological tracking stations essential for obtaining long-distance

weather information but the fixed aerial might prove a useful standby, for example for monitoring a second satellite frequency when the main station was in use.

From a study of signals from weather satellites which radiate elliptically polarized signals (commonly called circular) around 137MHz, useful signals can be obtained with a suitably orientated single dipole when the satellite's elevation is greater than 20°. A pair of crossed dipoles, with a quarter-wave delay in one element, avoids the necessity of rotating the system as it accepts rotating polarization. Such an aerial with the dipoles horizontal gives a fairly good polar diagram. A further improvement is effected by putting half-wave parasitic reflectors at 0.3 wavelength below the dipoles. The polar diagram is modified, broadening the lobe at low elevations at the expense of vertical gain (Fig.1). It also gives a null in the horizontal plane thus discriminating against ground stations and interference sources. The reflectors can be compared to the radials of a ground plane aerial in providing a virtual earth so that the polar diagram is similar to that for



The aerial described, erected on the roof of Westminster School, London.

* Westminster School, London.

a dipole of 0.3 wavelength over a perfectly conducting earth.

The elements are of $\frac{3}{8}$ in outside diameter aluminium tube and are mounted on the top of a 4m $1\frac{1}{2}$ in o.d. aluminium tube which forms the mast. This is wide enough to allow all connections to be made inside the tube for protection against the weather. Holes in the side of the mast take rubber grummets which support and insulate the dipoles. The screened copper braid of the dipole lead is laid bare at a



Fig. 1. Polar diagram of the aerial.

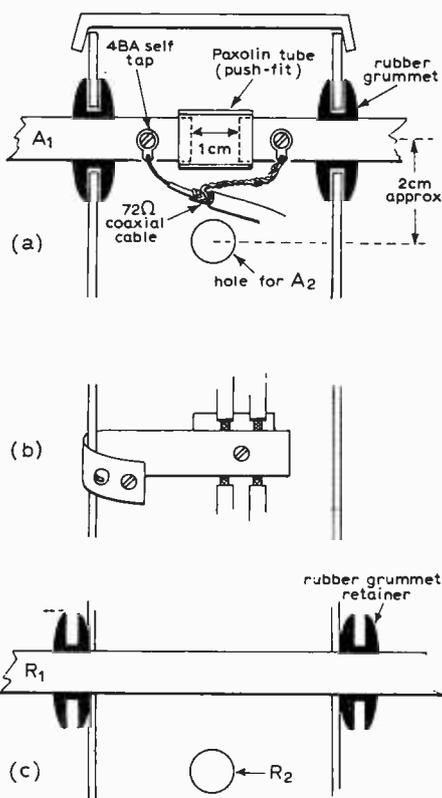


Fig. 2. Aerial construction; (a) showing dipole assembly; (b) the strapping of the braid of the feeder to the mast to form a balun; (c) the assembly of the reflectors 0.3 wavelength below the dipoles.

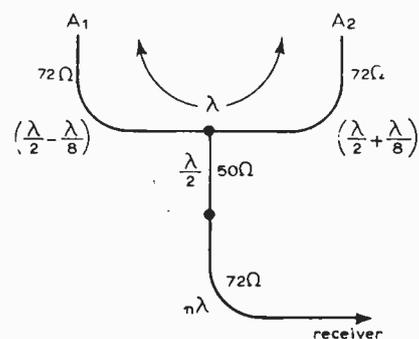
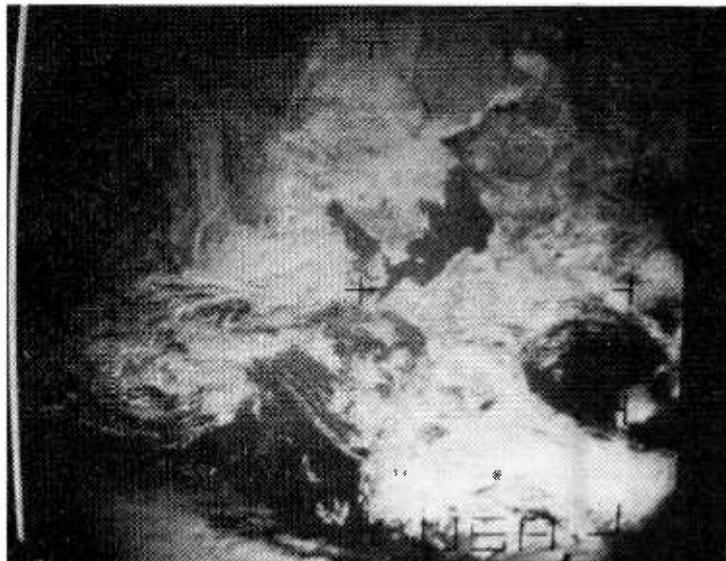


Fig. 3. The feeder system employed.

This picture centred on the Copenhagen area, showing the Baltic Sea, was received using the fixed aerial described.



quarter wave (55cm) below the dipoles and electrically connected to the mast at this point to form a balun (Fig.2). The matching section and down lead continue down inside the mast and emerge at the bottom. The reflectors, 0.3 wavelength (66cm) below the dipoles, are fitted through diametrically opposite clearance holes in the mast and are retained by grummets pushed over the reflectors from opposite sides. The element lengths (not optimized) are (a) dipole arms 52cm each, making 105cm between ends with a 1cm centre gap and (b) reflectors 110cm. The coaxial centre conductor and braid terminate in solder tags which are fixed to the dipole arms with self tapping screws. To make the dipoles mechanically rigid a short length of about $1\frac{1}{2}$ cm of Paxolin tube sleeving makes a push fit over the dipole arms where they nearly meet inside the mast.

The two dipoles have to be connected in parallel and a quarter-wave delay inserted in the right sense in one lead. As half-wave sections of feeder minimize matching errors whereas quarter-wave sections tend to aggravate mismatch, it was decided not to add a single quarter-wave in one arm only but to have one eighth short in one arm and one eighth extra in the other (Fig.3). The two dipoles are connected in parallel with an exact one wavelength section of coaxial feeder which is tapped off centre, 68cm from one end and 113cm from the other, the difference being 45cm (exactly quarter-wave).

The dipoles being in parallel across this tapping point we have approximately 36 ohms here from the two centre-fed dipoles with 72-ohm coaxial feeder. This point is matched to the main feeder, also 72 ohms by a quarter-wave section of 50-ohm coaxial feeder. The main feeder is also cut for a whole number of half wavelengths to minimize mismatches.

The dimensions given are for the frequency of the ESSA 8 satellite, 137.62MHz. The aerial is also suitable for the reception of Nimbus 4 on 136.9MHz and, indeed, the whole of the satellite band.

The lengths given for the coaxial sections take into account the velocity factors

of the particular cables used. The 72-ohm cable used was Radio Spares "Economy" (75-ohm to be precise) which has a velocity ratio of 0.82. The 50-ohm cable was Radio Spares "Uniradio 67" with a velocity ratio of 0.67. The latter cable is rather cumbersome and any low-loss type of the appropriate impedance could be used.

For making the measurements I use a noise bridge, which employs the receiver for detecting balance. This technique has the advantage of making the measurement on the actual electrical length at the working frequency. It makes cutting of the long length of main feeder to $n\lambda/2$ easy as one does not need to measure the length or to know n . However, given the manufacturer's specifications of the cables, such measurements are not necessary.

To make the best use of the signals a low-noise receiving setup is used. Two low-noise pre-amplifiers (using f.e.t.s in a conventional cascode circuit) are followed by a m.o.s.f.e.t. 2-m converter (by Solid State Modules) changing 137.62 MHz to 21.62 MHz i.f. This is tuned by an EC10 (Eddystone) receiver modified to use an external i.f. This is a commercial broadband 455-kHz f.m. strip from mobile equipment but the discriminator has been replaced by a phase-locked loop integrated circuit (Signetics NE561B).

High-quality tape recorder p.c.bs

Printed circuit boards for J. R. Stuarts' tape recorder design are available from Hart Electronics. Full kits of parts including ready-wound coils and tape heads are also stocked for the design, originally detailed in the November and December 1970 issues. Hart Electronics, Penylan Mill, Oswestry, Shropshire SY10 9AF.

The Junction F.E.T. as a Voltage-controlled Resistance

— with particular reference to communications receivers

by R. M. Lea*, M.Sc.

The wide range of drain-source channel resistance of junction f.e.t.s, under control of the gate-source voltage, makes them useful in high-performance a.g.c. systems. This two-part article first outlines shunt and series a.g.c. systems, discusses design problems and then in part 2 gives practical circuits for a series and shunt a.g.c. system, volume compression and expansion, squelch and battery saving. Circuits are especially suitable for mobile communications receivers.

The current in a typical *n*-channel, depletion mode f.e.t. is transported by electron flow between the source and drain terminals by a channel of *n*-type material which is restricted in dimension by the depletion region of the reverse-biased gate junctions, Fig. 1. The interesting condition is when the two depletion regions merge, for then the current flowing through the device is limited only by carrier injection from the source region of the channel and is virtually independent of drain voltage variation.

Thus a junction f.e.t., operating in the saturation region of its output characteristic, may be considered as a resistance between the drain and source, R_{ds} , which is critically dependent on the gate-source voltage, V_{gs} , and independent of the drain voltage, V_d . This latter property renders the channel resistance, at a particular control voltage, V_{gs} , remarkably linear for signals applied to the drain terminal of the device.

The channel resistance characteristic of Fig. 2 illustrates the suitability of a variably-biased junction f.e.t. for controlled resistance applications. Within the saturation region of the channel resistance characteristic, the device exhibits:

- large range of channel resistance from $1k\Omega$ to several $M\Omega$.
- critical dependence of R_{ds} on the control voltage V_{gs} ; a 0.5-V change in V_{gs} can change R_{ds} from $1k\Omega$ to $1M\Omega$.
- high degree of linearity for constant V_{gs} .
- very low operating current; drain current $< 30\mu A$.

As the curves show, a disadvantage of

operation within the saturation region is the relatively large change in the drain voltage, V_d , over the channel resistance range. Within the linear region of operation ($R_{ds} < 1k\Omega$) the change of the drain voltage ceases to be a problem, but the above advantages of the saturation region do not apply to the linear region.

A.G.C. systems in general

There are two basic circuit configurations for a.g.c. of audio amplifiers—the shunt and series systems represented in Fig. 3—whose effect is shown in Fig. 4.

For output signals below V_{o1} , zero a.g.c. is delivered so that the overall gain of the

†A. S. Grove, "Physics and Technology of Semiconductor Devices", Wiley, 1967, pp. 243-57.

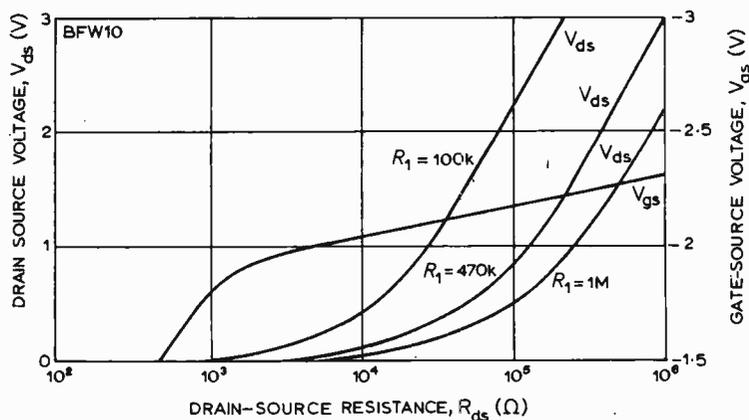


Fig. 2. Drain-source channel resistance characteristic of BFW10 for very low drain currents ($< 30\mu A$). R_1 is a resistance in the drain circuit.

*Electronics Dept., Southampton University.

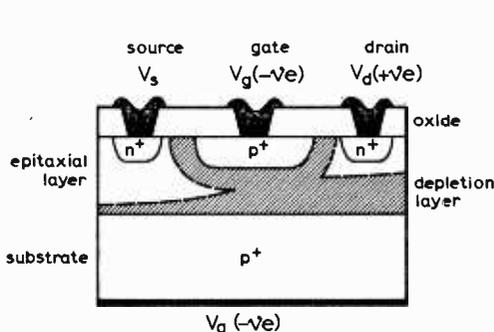


Fig. 1. When the two depletion regions meet in an epitaxial planar junction f.e.t., current in the device is independent of drain voltage variation.

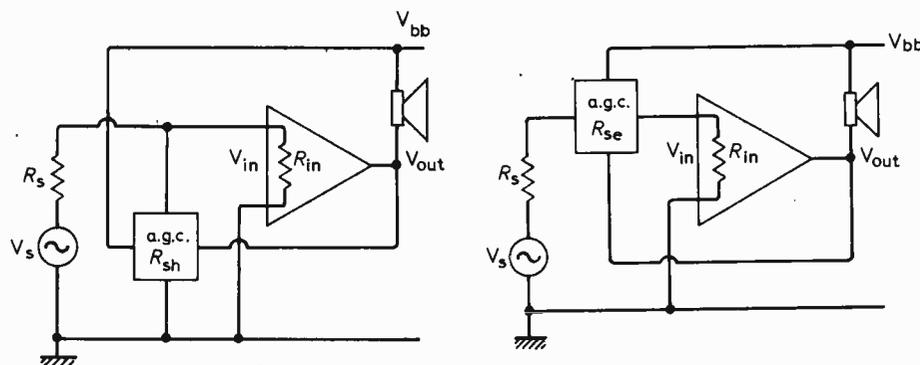


Fig. 3. Shunt a.g.c. system (a) has greater constraints on component values than series circuit (b).

amplifier is unaffected. For increasing V_s , above V_{s1} , the input to the amplifier is attenuated to prevent a corresponding increase in V_{out} . The limit of a.g.c. is shown for values of V_s approaching V_{s2} when maximum attenuation is supplied. Thus for effective a.g.c.

$$\bullet \frac{V_{o2} - V_{o1}}{V_{o1}} \ll \frac{V_{s2} - V_{s1}}{V_{s1}}$$

- Attack time must be so short that sudden peaks of input signal can be controlled.
- Recovery rate must be slow enough for the distortion of 'volume compression' not to be detectable.
- For a given level of attenuation the a.g.c. element should behave as a linear circuit component.

In a shunt system, a.g.c. effectively shunts the input to the audio amplifier with a variable resistance, R_{sh} , controlled by the amplifier's output audio level V_{out} —Fig. 3(a). Using the terminology of Figs 3 and 4, the amount of attenuation provided by the shunt a.g.c. element is

$$\frac{V_{in}}{V_s} = \frac{R_{sh} \cdot R_{in}}{R_s \cdot (R_{in} + R_{sh}) + R_{sh} \cdot R_{in}} \quad (1)$$

There are two cases of shunt operation. With no shunt a.g.c. applied, $R_{sh} \gg R_{in}$, so equation 1 becomes

$$\frac{V_{in}}{V_s} \approx \frac{R_{in}}{R_s + R_{in}} \quad (2)$$

Thus for an output audio level less than the threshold a.g.c. level (V_{o1}) then R_{sh} should be very large compared with R_{in} and R_s should be very small compared with R_{in} for maximum signal transfer.

With shunt a.g.c. applied, $R_{sh} \ll R_{in}$, equation 1 becomes

$$\frac{V_{in}}{V_s} \approx \frac{R_{sh}}{R_s + R_{sh}} \quad (3)$$

Thus for appreciable attenuation of input signals greater than the threshold level then R_{sh} should be very much smaller than R_{in} and R_s . Indeed, for maximum attenuation the source impedance, R_s , should be very large indeed.

From these considerations there are two opposing requirements for the value of R_s . It can be shown by partial differentiation of equations 2 and 3 that there is an optimum value of R_s necessary to obtain the maxi-

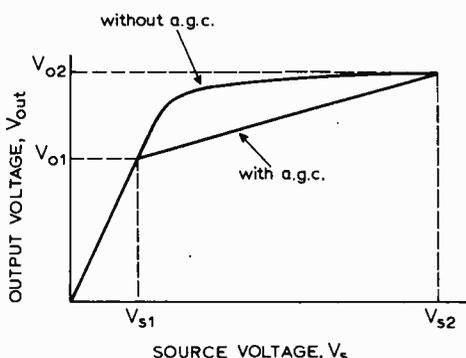
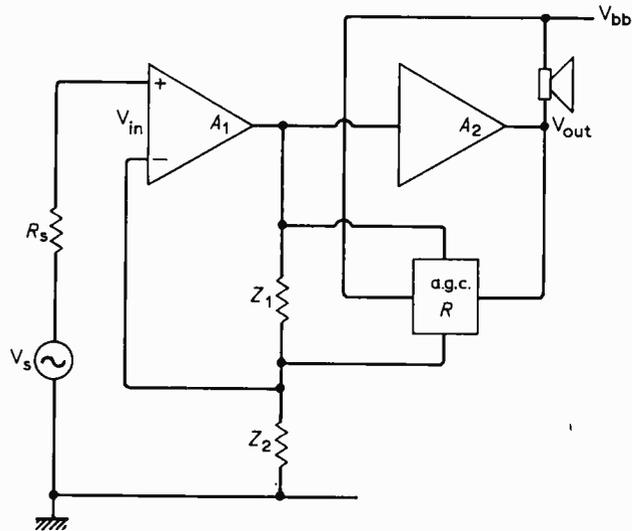


Fig. 4. Effect of a.g.c. on amplifier characteristic.

Fig. 5. Application of shunt circuit to feedback loop of first stage only, to avoid operating near knee of f.e.t.



imum attenuation range of the a.g.c. system, i.e. $R_s = R_{in}$. Thus a limitation of the shunt system is that, for a wide range of input signal attenuation, the choice of source impedance or input impedance of the amplifier is restricted.

For example, a shunt system having a minimum R_{sh} value of $1k\Omega$ and designed to deliver 40dB of input signal attenuation, is useful only for an audio amplifier having R_s and R_{in} equal to and greater than $100k\Omega$.

The series a.g.c. stage provides a variable series resistance, R_{se} , in the input circuit of the amplifier, controlled by the amplifier's output audio level, V_{out} , Fig. 3(b). For larger audio output the series a.g.c. resistance increases, whereas the shunt resistance would have reduced in value.

Using the terminology of Figs 3 and 4, the amount of attenuation provided by the series element is

$$\frac{V_{in}}{V_s} = \frac{R_{in}}{R_s + R_{se} + R_{in}}$$

Thus for series a.g.c. there is no restriction on the value of the source impedance—for maximum signal transfer it should be as low as possible. But there is a restriction on the value of the input resistance of the amplifier. For an output audio level less than the a.g.c. threshold level the value of R_{se} should be very much less than R_{in} for unimpeded signal transfer. As there must be a lower limit to the value of R_{se} , it is beneficial for the value of R_{in} to be fairly high. For appreciable attenuation of input signals greater than the threshold value R_{se} should be very much larger than R_{in} . For maximum attenuation it would be beneficial for the value of R_{in} to be fairly low.

For example, a series system having a minimum R_{se} value of $1k\Omega$ applied to an audio amplifier having a $10k\Omega$ input impedance and driven from a $1k\Omega$ source impedance would have the following characteristics

- for $V_{out} \leq V_{o1}$ the input signal would be attenuated by 2dB;
- for $V_{out} > V_{o2}$ the input signal could be attenuated by as much as the a.g.c.

biasing circuitry allowed.

Hence, an advantage of the series a.g.c. system is that very large attenuation may be delivered as there is theoretically no limit on the maximum value of R_{se} . In practice the audio amplifier and associated circuitry of the series system limits the extent of available attenuation.

Feedback a.g.c.

A better system for a.g.c. would be to incorporate the controlled resistance element in the a.c. negative feedback loop of the audio amplifier. A shunt system of this type is shown in Fig. 5, where the a.c. negative feedback over the first stage of the amplifier is rapidly increased by shunting the impedance Z_1 with R , for output voltages above the required a.g.c. threshold, V_{o1} .

This type of a.g.c. system improves linearity, noise and amplitude response, but suffers from a limitation in signal handling capability.

Linearity of a controlled resistance arises because in saturation the drain of the f.e.t. is essentially isolated from the channel. However, large variations in drain voltage, especially when operating near to the knee of the output characteristic of the device, may introduce non-linear distortion. Thus there is a limit to the signal amplitude that can be applied to the drain.

Hence the system shown in the circuit of Fig. 5 incorporates shunt a.g.c. in the feedback loop of only the first stage of the amplifier.

Another way of overcoming this limitation is to apply series a.g.c. to the feedback loop. In applications where the d.c. and a.c. feedback loops are separated the f.e.t. could replace the impedance Z_2 of Fig. 5. This would allow the significant advantage of increased amplifier gain for low input signals thereby increasing the dynamic range of the audio amplifier.

Design considerations

A simple circuit is shown in Fig. 6 to illustrate the design requirements of a junction f.e.t. acting as a v.c.r.

The drain-source channel of the device provides the variable resistance, R , which is controlled by the voltage $V_{gs} = V_g - V_{so}$. It is convenient to keep the source voltage, V_{so} , constant so that the channel resistance, R , is rendered more sensitive to the gate voltage, V_g . This is effected by the potential divider incorporating resistors R_2 and R_3 which are chosen to give the required value of V_{so} and hold it constant against all drain-source current variations. When V_g is at earth potential, the value of V_{so} determines the maximum value of the channel resistance, and the a.g.c. threshold level, V_{o1} .

Resistor R_1 limits the drain current of the device to a convenient value and tends to speed up the change of the channel resistance with change of gate voltage. A large value of resistance must be chosen for R , as it shunts the channel resistance.

A stage is required to give a direct output voltage related to the output audio level, V_{out} , of the amplifier under a.g.c. Of the many circuits that can provide this function, a simple diode pump arrangement has been chosen as an example. Basic circuits for both shunt and series a.g.c. applications are shown in Fig. 7.

Although these driver circuits may be designed empirically, the process is much simplified by using a computer program for large signal analysis of simple non-linear systems. Such a program, specifically written for diode-pump applications, was used for the circuit shown in Fig. 9 (next issue).

Driver stage for shunt systems

Fig. 7(a) shows a basic pump circuit configuration for the driver stage of a shunt a.g.c. system. The transistor acts as a rectifier of controllable forward resistance and also as a high input impedance buffer stage, necessary to ensure that the audio output of the amplifier is not loaded by the a.g.c. circuitry.

It should be chosen for appreciable current gain at very low collector currents, very low leakage current and low collector series resistance. Hence a silicon epitaxial planar transistor such as the BC108, or possibly the BC109, would be suitable.

The resistors R_1' and R_2' bias the transistor off in its quiescent state and they must be sufficiently high-valued to avoid loading the output of the audio amplifier. It is important to bias the transistor just off to change the capacitor C to the maximum extent.

The requirements of the capacitor C are conflicting. It is necessary to find the best compromise between: a low value for a fast a.g.c. attack rate; a high value for a minimum of a.c. ripple on the output d.c. level of the driver stage; a low series impedance for minimum a.c. ripple; and a low leakage current, for the long period of charge storage necessary for slow a.g.c. recovery. These requirements suggest use of medium-valued solid tantalum electrolytic capacitors. However, the higher cost of these components would probably lead to their rejection in favour of the wet aluminium electrolytic capacitors. Although the use of these components is acceptable, the above requirements should be carefully considered.

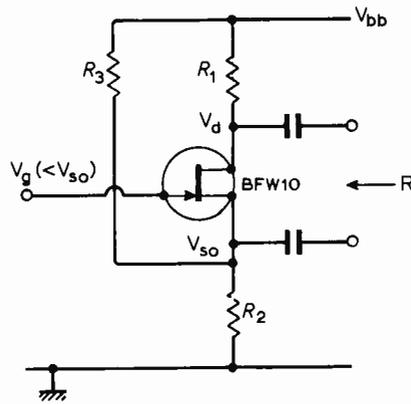


Fig. 6. Basic circuit for using a junction f.e.t. as a voltage-controlled resistance.

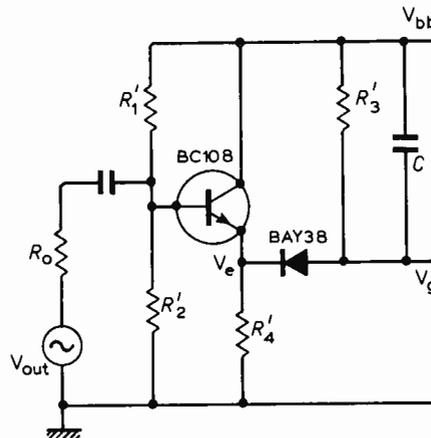
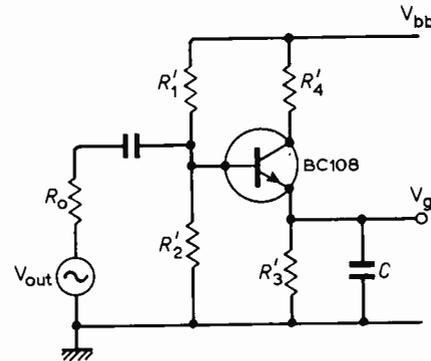


Fig. 7. Basic circuit of 'pump' driver stage for shunt system (a) and for series system (b).

Resistor R_4' determines the forward resistance of rectification and hence controls the charging rate of the capacitor C . Thus, the smaller the value of R_4' the faster the attack of the a.g.c. system.

Resistor R_3' can be chosen according to the required a.g.c. recovery time, the product CR_3' being the recovery time constant.

Driver stage for series systems

Fig. 7(b) shows a basic diode-pump circuit configuration for the driver stage of a series a.g.c. system. The circuit is almost identical to that shown in Fig. 7(a) for a shunt system and the same design considerations apply, except for two points. A diode is used in place of a transistor, for the essential non-linear element providing rectification. This

diode should be chosen for very low leakage current and low series resistance. A silicon epitaxial planar diode such as the BAY38 would be suitable. The transistor in the circuit provides the same buffer impedance as in the shunt a.g.c. driver circuit. Also the transistor is biased so that the quiescent emitter voltage, V_e , provides the upper limit of the output d.c. level, V_g .

Parasitic a.g.c. action due to voltage spikes

Ideally there should be no transfer of signal between the gate and the drain of the f.e.t. In fact the gate should control only the channel resistance between the drain and source terminals. However, in the circuit configuration of Fig. 6 there is a d.c. transfer between the gate and drain (see Fig. 2) such that the a.g.c. attack is accompanied by a voltage spike at the input to the amplifier.

The presence of these voltage spikes is most critical at the threshold of operation and they can cause undesirable noises in the audio output of the amplifier and also aperiodic oscillation due to amplified spikes causing attack of the a.g.c. system followed by normal recovery. These deleterious consequences of the circuit arrangement of Fig. 6 may be satisfactorily minimized in one of three ways.

First, controlling the rate of change of the drain voltage, this transient condition can be altered to correspond to a signal well outside the frequency pass-band of the amplifier. As the higher transient speed is limited by the charging rate of capacitor C of the driver circuit (Fig. 7), it is advisable to slow the transient to correspond to a very low frequency. For example if the a.g.c. attack time-constant is 100ms then the transient will correspond to a signal frequency of less than 10Hz.

Secondly, by restricting the value of V_{ds} to a low value, for the a.g.c. off condition, the voltage spikes at the threshold can be effectively minimized. It can be seen from the curves of Fig. 2 that the higher the value of the resistance R_1 the lower the value of V_{ds} for a given value of channel resistance. Therefore in applications where the voltage spikes are problematic, it is advisable to choose a high-valued resistor for R_1 —the upper limit being set by the extent to which the drain current may be reduced before noise problems occur. A good method of restricting V_{ds} is to shunt the drain of the f.e.t. to earth through a resistor. The value of V_d , and hence $V_{ds} (= V_d - V_{so})$ is fixed for the a.g.c. off condition by a potential divider, resistors R_1, R_{d1}, R_{d2} (not shown). The resistor R_{d1} is variable so that the required value of V_{ds} can be accurately set. The series resistance $R_{d1} + R_{d2}$ should be chosen to be very much larger than the input impedance of the amplifier to which a.g.c. is applied.

Thirdly, the drain voltage of the f.e.t. shown in the circuit of Fig. 6 can be kept constant during change of the channel resistance of the device, if the value of R_1 could be changed at exactly the same rate. This can be done by replacing R_1 with an identical f.e.t., as shown in Fig. 8. The operation of this circuit arrangement is

simply explained below, using the terminology of Fig. 8.

$$V_{d2} = \frac{V_{bb}}{1 + \frac{R_{ds1}}{R_{ds2}}} \quad (4)$$

$$V_{gs1} = V_{g1} - V_{d2}$$

$$V_{gs1} < 0$$

Consider two cases of a.g.c. operation. Suppose V_{d2} increases by a small amount due to a small increase in the value of R_{ds2} . In this case V_{gs1} becomes more negative. Therefore R_{ds1} increases, tending to restore the original value of V_{d2} . Suppose V_{d2} decreases by a small amount due to a small decrease in the value of R_{ds2} . In this case V_{gs1} becomes less negative. Therefore R_{ds1} decreases tending to restore the original value of V_{d2} .

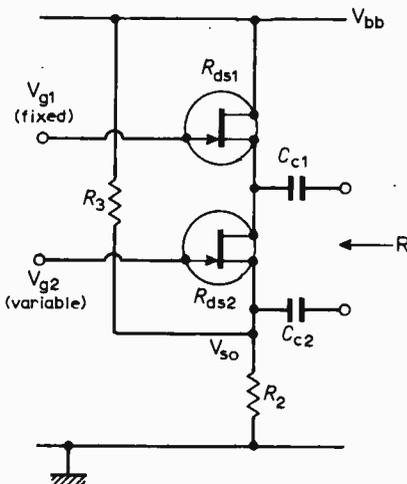


Fig. 8. Modification of Fig. 6 circuit to cancel voltage spikes that accompany a.g.c. attack.

Thus, the system is self-stabilizing, tending to reduce the change of V_{d2} to a minimum, for all values of the resistance, R . As can be seen from equation 4 this minimum depends on the degree of match of the characteristics of the two f.e.t.s. The use of identical f.e.t.s is becoming attractive for a.g.c. application with the advent of dual f.e.t. encapsulations and the facility to incorporate f.e.t. devices in bipolar integrated circuits.

Low frequency relaxation oscillation

For an amplifier with a.g.c., to maintain fidelity to incoming signals it is necessary that the a.g.c. system should have a fast attack and a very low recovery rate. Such a system is prone to low frequency relaxation oscillation for constant input signals of amplitude greater than the threshold.

This problem is obstructed by the voltage spike problem, which can also cause low frequency oscillation. However, these problems are independent, for if the d.c. transient at the drain of the f.e.t. is minimized the low frequency relaxation oscillation can still exist.

Referring to Fig. 3, it can be seen from the following argument that relaxation oscillation can occur in the a.g.c. system

with a constant input signal that is sufficiently large to cause a.g.c. operation.

In practice there is a finite time delay in the attack of increases in input signal level. During this delay the amplifier's output signal has excessive amplitude, tending to cause the system to deliver maximum attenuation of the input signal. Thus for a short time the input signal to the amplifier is over-attenuated so that the amplifier output is reduced below the threshold, V_{o1} . The system recovers with a long time constant until V_{out} again exceeds V_{o1} , when the cycle repeats itself.

A solution to this problem would be found by speeding up the attack time. However, the minimum attack time is limited by the nature of the a.g.c. system's circuitry to several milliseconds. Consequently the problem must be solved by slowing down the attack time, to correspond to a low frequency at which the loop gain is insufficient for this oscillation to occur.

This delay can be effected by increasing the forward resistance of rectification of the driver circuit, shown in Fig. 7. Thus by increasing the value of the resistor R_4' the stability of the system can be improved. In practice resistor R_4' should be set at the lowest frequency for which a.g.c. is required.

As mentioned above this low frequency relaxation oscillation is complicated by the presence of the d.c. transient at the drain of the f.e.t., the so-called voltage spike. If the dual f.e.t. approach is adopted then it is probable that faster attack times would be possible without oscillation.

Failure of devices due to transients

During normal operation of an n -channel junction f.e.t. the controlling gate-source voltage is negative. Under this condition the input impedance at the gate terminal is very high indeed and very little gate current is taken. However, if this potential difference should become positive the input impedance would be very low indeed and an appreciable gate current would flow, limited only by external resistance in the circuit.

The limiting value of gate current for the BFW10 is given in the published data ratings as 10mA.

The junction f.e.t. in the circuit of Fig. 6 is most vulnerable during the transient conditions of switch-on and switch-off of the supply voltage. To provide protection for the device it is advisable to insert a large value of resistance in the gate circuit of the device.

(To be continued)

Books Received

Cecil E. Watts — Pioneer of Direct Disc Recording by Agnes Watts. It would not be an exaggeration to describe the late C. E. Watts as a genius in the mould of Edison. This intriguing account of the domestic background to his tireless work and original thought in the cause of high-quality sound reproduction reveals the vicissitudes of his early struggles and the ultimate success of his formulation for a disc coating giving an unprecedentedly low level of background noise. The importance of this discovery and its impact on the development of wartime reporting, together with Watts' work on portable recording machines, resulted in his company being taken over by the Government. Mrs Watts' lively account reveals less well-known aspects of her husband's subsequent work in recording as an aid to scientific research, and in the care and maintenance of disc records. Pp. 116. Price £2.25. William Clowes & Sons Ltd, London.

The Origins of Maritime Radio by R. F. Pocock and G. R. M. Garratt is a story of the introduction of wireless telegraphy into the Royal Navy between 1896 and 1900. The foreword states that much of the early history of wireless telegraphy, particularly the period covered by this book, has hitherto been largely unknown to the outside world, mainly due to the fact that government departments were involved — particularly the Admiralty and the Royal Navy. Records were recently discovered by Mr. R. F. Pocock in the Public Record Office, and their examination by Mr. Pocock and Mr. G. R. M. Garratt, a Keeper of the Science Museum, has disclosed the story which is briefly told. The book is available from the Science Museum or Her Majesty's Stationery Office. Pp.60. Price £1.00. Her Majesty's Stationery Office, 49 High Holborn, London WC1V 6HB.

Introduction to Semiconductor Devices by F. J. Bailey aims to provide sufficient basic theory and practical know-how for students, technicians and engineers to be able to understand and use semiconductor devices. SI units have been used throughout. The emphasis is on a practical study of the devices, assuming very little prior knowledge of either semiconductor theory or electronics in general.

The first chapter contains an introduction to semiconductor theory, sufficient to enable the reader to understand the subsequent chapters. The bulk of the book then deals with some of the more fundamental electronic devices: diodes, bipolar transistors, junction f.e.t.s, insulated gate f.e.t.s (m.o.s.) and thyristors. In each case the devices are considered in some detail, the general pattern being: fabrication, principles of operation, characteristic curves, parameters, equivalent circuits, general remarks on merits and demerits, and, finally, applications. Particular attention is paid to explaining manufacturers' data sheets. The final chapter is devoted to monolithic i.c.s. Pp. 238. Price £3.92 (hardback), £1.95 (paperback). George Allen & Unwin Ltd, Ruskin House, Museum Street, London W.C.1.

Instruments, Electronics & Automation

Some new products seen at the recent London Shows

The trend, mentioned last month in the report on the Paris Components Show, of a general decline in interest in large conglomerate exhibitions continued with the I.E.A. Show. There was about a 30% drop in the number of visitors compared with the 1970 I.E.A. (110,000 to 77,000). The number of exhibitors fell from 800 to 704; the missing 96 included a number of big names. The exhibition catalogue sold out — sales were about 12% up on the previous year — which reflects the organizers attempt to make it into a more useful reference work. No advertising was allowed in the catalogue and exhibitors had to pay for the space used to describe their products in order to recover the cost of publication.

The organizers are to be applauded for their efforts in sectionalizing the exhibition (as was done in Paris) but by doing so admit that there is a need for specialist exhibitions even if they take the form of separate exhibitions within a much larger exhibition.

For the seventh successive year an exhibition was staged at the Kensington Close Hotel during the I.E.A. Show. The companies participating were Exel, Hinchley, Marconi Instruments, P.C.D., REL Equipment and Components, Smiths Industries, Tektronix U.K. and Telonic Industries U.K. There were also several other exhibitions staged by individual companies in West London hotels during the I.E.A. week and we have included in these pages new products from several of the "extra mural" shows as well as products seen at the I.E.A.

I.C. for television receivers

Now appearing in some of the new television sets at the recent trade shows in London is the Mullard integrated circuit type TCA270. Claimed to be one of the largest i.c.s produced for use in television receivers, it contains circuits intended to obtain the maximum benefit from integration techniques. It has a synchronous demodulator and an a.f.c. system using integrated circuit multipliers that the makers say would be impracticable with discrete components. Conventional detectors and a.f.c. systems tend to radiate harmonics, resulting in interference patterns on certain television channels. The balanced multiplier circuits in the TCA270 are said to give negligible harmonic radiation. The synchronous detector, which achieves carrier regeneration by filtering and limiting the input signal, gives, according to set designers, significant improvements in linearity and intermodulation compared to the conventional envelope detector. The beat pattern between sub-carrier and sound is not visible on some of the receivers we have seen at the recent trade shows. The a.f.c. discriminator, which requires an external single tuned

circuit to set the 90° phase shift, is followed by a d.c. amplifier providing a control output suitable for application to varicap tuners. The device also includes a video amplifier that inverts interference pulses to protect the synchronization channel and prevents noise pulses appearing in the picture. Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

WW509 for further details

Automatic circuit tester

The Testmatic TM30 will perform up to 30 d.c. measurements automatically on circuit cards. If a fault is detected the instrument stops and provides an indication of which test the circuit failed and why it failed (high or low, plus a meter indication). The test sequence and measurement limits (100k Ω /V) are 'hard-wired' on a plug-in programme card which can be changed rapidly. The instrument contains two power supplies for the circuit under test which can be set between 5 and 18V (500mA) and can be switched by the programme. The pass tolerance for each

test can be set between 2.5 and 45% of the desired value. A complete test sequence takes between 5 and 15 seconds, depending on the exact details of the programme. Connection to the board under test can be by plugs, sockets or spring loaded probes. The Testmatic is manufactured by The Wayne Kerr Company Ltd, Tolworth, Surbiton, Surrey KT6 7ER, and costs £650.

WW504 for further details

Complementary m.o.s. memories

An example in Motorola's c.m.o.s. range on display was a 64-bit, fully decoded, non-destructive read/write random access memory, type MCM14505L. It is claimed to be the first c.m.o.s. memory to include full address decoding within the monolithic structure. Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middlesex.

WW516 for further details

Pure sinewave source

An instrument which will provide either a pure sinewave (harmonic distortion typically 0.1% up to 10kHz) or a square wave, both over the frequency range 0.1Hz to 1.1MHz is the Signal Source Type 472 shown by Brookdeal Electronics. Using a linear segment method of generating the sinewave, the instrument is claimed to be completely free of "amplitude bounce", making it suitable for swept frequency applications. Frequency control, either manual or by an external voltage, is linear over the range of two decades. The sinewave output level is controllable from 5V r.m.s. to less than 500 μ V by a continuously variable potentiometer in conjunction with a push-button step attenuator (0, 20, 40 and 60dB steps). Square-wave output is fixed at 3V peak-to-peak from a 600 Ω source. This instrument is the second of a series of signal sources, the first of which was the company's Type 471. Brookdeal Electronics Ltd, Market Street, Bracknell RG12 1JU, Berks.

WW507 for further details

Noise dose meter

The damage to hearing caused by loud noises is a function not of instantaneous peak sound pressures but of sound energy, which is a product of sound power and time. The Bruel & Kjaer noise dose meter, type 4423, operates in conjunction with sound measuring equipment to assess what is termed the "equivalent continuous sound level" according to the requirements of the ISO Recommendations R1996 and R1999 and of DIN 45641 (draft April 1971). The frequency range is 20Hz to 20kHz, measuring time can range from a few seconds to several days, and three built-in counters indicate noise dose count, elapsed time and overload duration (if any).

The instrument has to be used with a special slide rule (supplied) by which the "equivalent continuous sound level" is calculated from the readings of the counters and the setting of the attenuator in the associated sound measuring equipment. The makers say that the instrument is equally applicable to the assessment of hearing impairment risk and annoyance from industrial or road and air traffic noise. Any of the B & K sound level meters, measuring amplifiers, spectrometers and analysers may be used in a dose measuring set-up. B & K Laboratories Ltd, Cross Lances Road, Hounslow, Middx.

WW510 for further details

Sine/square generator and frequency meter

Intended mainly for testing amplifiers, this new Grundig sine- or square-wave generator operates up to 1MHz. It can also be used as an accurate frequency meter using the number-tube frequency display. Out-

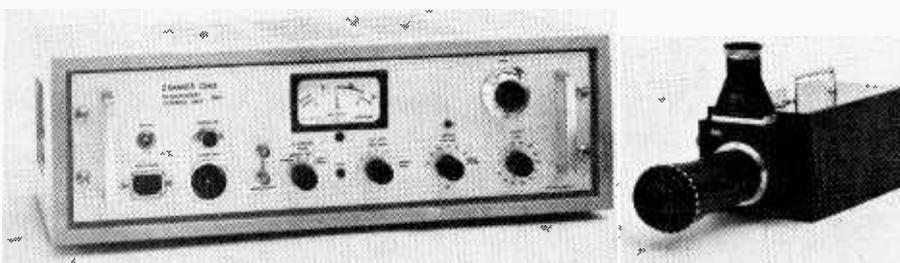
put voltage is up to 10V (asymmetrical) and for square-wave outputs rise and fall time is 20 and 30ns respectively. Available in the U.K. from Grundig (Gt. Britain) Ltd, 40 Newlands Park, London SE26. WW519 for further details



Non-contact displacement measurement

A precision non-contact displacement transducer intended for investigating mechanical resonances and vibrations is made by Zimmer Ohg, of Darmstadt. The electro-optical instrument will produce an electrical indication of displacement, velocity or acceleration of small structures up to a frequency of 150kHz. It is specially useful in electronics, for example in checking switch and relay contact bounce, development of electromechanical devices, investigating mechanical resonance of "whiskers" in diodes, in integrated circuits,

capacitors and in printed circuit structures and of course it has many uses in other branches of engineering. In operation, a closed-loop scanning system keeps the image at a central point on the aperture plate of an image converter tube, target displacement being proportional to a correcting current fed through a deflection yoke. Accurate in displacement measurement to $\pm 1\%$, its range extends from 1mm to 20m full scale with a resolution of 0.01%. Obtainable in the U.K. through Tekman Electronics Ltd, 52 St Clements, Oxford. WW517 for further details



Timebase counter-timer

The Model 104B manufactured by Monsanto is a preset/variable timebase counter-timer with all the conventional dual input counter-timer facilities and also the ability to scale the input or the internal clock by a factor n or $1/n$. Preset operation is also possible. The instrument includes b.c.d., gate and carry pulse outputs and remote control facilities. Frequency range is from 5Hz-32MHz, and it will count up to 99,999 pulses and can be preset for any number up to this figure. At the end of the preset figure, count measurement ceases and an output signal is provided. Period and multiple period measurements cover from $2\mu\text{s}$ to 100s. Provision is made for rate measurement over the frequency range of the counter. In the rate mode, the count is normalized by any timebase over a $10^8:1$ range. The internal crystal oscillator has an ageing rate of better than $5:10^8$ per day. The instrument measures $4\frac{1}{2} \times 7\frac{1}{2} \times 9$ in and weighs 5lb 10oz. Techmation Ltd, 58 Edgware Way, Edgware, Middlesex HA8 8JP.

WW515 for further details

Inexpensive resistors

Metal-film resistors made by Electronorg-technika, of U.S.S.R., are available through Z & I Aero Services. With 1-watt dissipation at 70°C they have $\pm 5\%$ tolerance in values ranging from 27Ω to $10\text{M}\Omega$, and cost £3.80 per 1000. Three other types with $\frac{1}{8}$, $\frac{1}{4}$ and $\frac{1}{2}$ watt dissipation will be available shortly. Pre-set potentiometers with a dissipation of $\frac{1}{4}$ watt cost £20 per 1000 and 100mW types costing £5 per 1000 will be in stock soon. Carbon film resistors are stocked in $\frac{1}{8}$, $\frac{1}{4}$ and $\frac{1}{2}$ watt sizes costing £3 per 1000. Z & I Aero Services Ltd, 44a Westbourne Grove, London W2 5SF.

WW526 for further details

Reed relays

A range of miniature dry-reeds relays by Willy Günther KG, of Nuremberg, is available from Engel & Gibbs. Types so far available have contact ratings from 0.15 to 2.5A, contact resistance of 80 to $300\text{m}\Omega$ and maximum switching voltages of 250V. Rated loads range from 3 to 100VA. A particular type of relay is the 3340 series intended for p.c. board mounting. Resistance of the standard rhodium contacts is $120\text{m}\Omega$ rated at 12VA with maximum current of 0.5A, but other contact materials are available. Coils are available for 5, 6, 12 and 24V with resistances of 270, 270, 1000 and 3300Ω respectively. Relays with latching and break contact arrangements will be available shortly. Engel & Gibbs Ltd, Elstree Way, Boreham Wood, Herts WD6 1SQ. WW524 for further details

continued on page 343

Batteries for stand-by operation

Sonnenschein "dryfit" ST batteries are a new range designed for float operation and offer 100% starting capacity with extended shelf life (compared with the original PC range) under constant charge conditions of between four and five years. Independence of operating position is due to the retention of the electrolyte in the battery by its gel-like substance. Each cell is closed by a one-way valve which incorporates a safety device to prevent evaporation loss of electrolyte. Capacities range from 2V to 12V with ampere-hour ratings of 0.9 to 20. F.W.O. Bauch Ltd, 49 Theobald Street, Boreham Wood, Herts WD6 4RZ.

WW520 for further details

Constant-voltage, high-energy batteries

Zinc-air cells, which have better energy density than other primary cells, are in quantity production by Energy Conversion Ltd. Known as Metair cells, they maintain constant voltage for longer than mercury-zinc cells. They have a good performance at high discharge rates and can supply up to 10 times their rated current for pulsed loads. The R1 ('N') size cell intended for low discharge rates (hearing aids, transistor radio sets) has a capacity of 1.3Ah at 15mA, and a voltage of 1.4V at 10mA. A high-rate version also has a 1.3Ah capacity but at 100mA, with a voltage of 1.1V at 150mA. A capacity of 3.25Ah at 350mA is featured in the R6 ('AA') size. For heavier-duty use including driving motors, cell size R20 or 'D' (U2 size) provides 20Ah at 1A, cell voltage being 1.1V at 2A. (Capacities given refer to 20°C and discharge down to 0.8V.) Operating temperature range is -20 to +55°C, and the cells have a shelf life such that capacity is 85% after a year's storage at +20°C. Energy Conversion Ltd, Priestley Road, Basingstoke, Hants.

WW527 for further details

Sound level meter

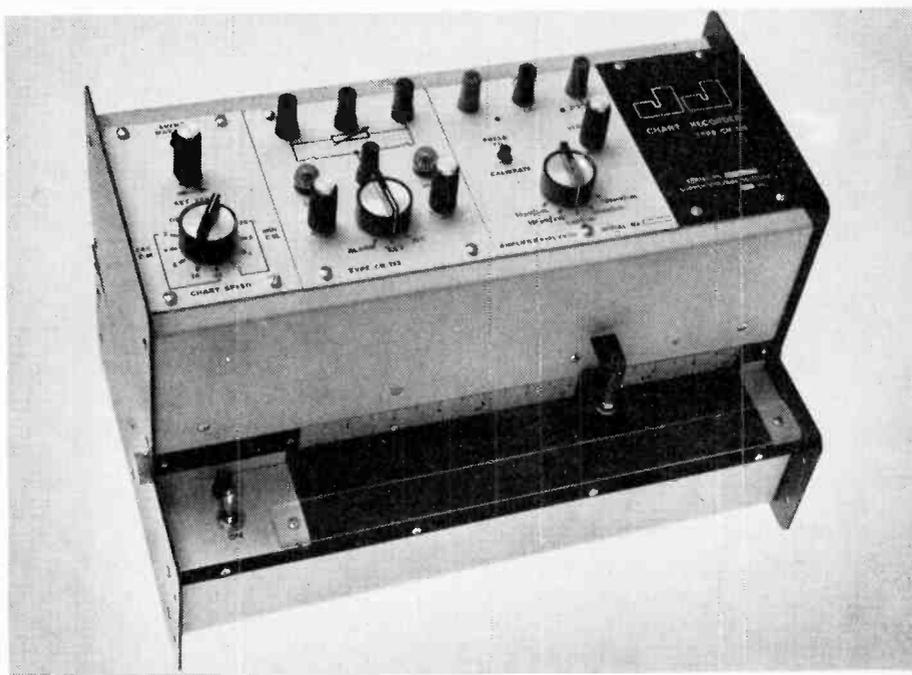
Mamash Applied Science Laboratories (Ramat Aviv, Israel) showed a portable sound level meter which could be read from a distance eliminating inaccuracies caused by reflections from the body of the operator. The instrument (type S716) covers the range from 18 to 140dB with automatic switching between two ranges widening the coverage obtained without the need for manual switching. An octave band analyser can be connected to the sound level meter for use between 31.5Hz and 8kHz. A separate unit used for initial calibration is available and has a battery life of 150 hours (continuous).

WW506 for further details

Chart recorder

The CR100 paper recorder by J. J. Lloyd Instruments incorporates many features not normally included on a £125 recorder. It has nine paper speeds, from the inverse of 1s/cm to 10min/cm. Sensitivity is 20cm/mV with a plug-in pre-amplifier. It will take a ball-point, felt-tipped or capillary stylus pen and has a writing speed of 20cm/s. The infinite-resolution servo feedback potentiometer with its linearity of

0.1% allows a basic accuracy of $\pm 0.5\% \pm 1\text{mm}$. There are many other features about the instrument, like alternative plug-in amplifiers, manual advance, remote event marker, pH facilities, transparent bed for projection, adjustable damping and voltage reference for calibration. J. J. Lloyd Instruments Ltd, Brook Avenue, Warsash, Southampton SO3 6HP.
WW523 for further details

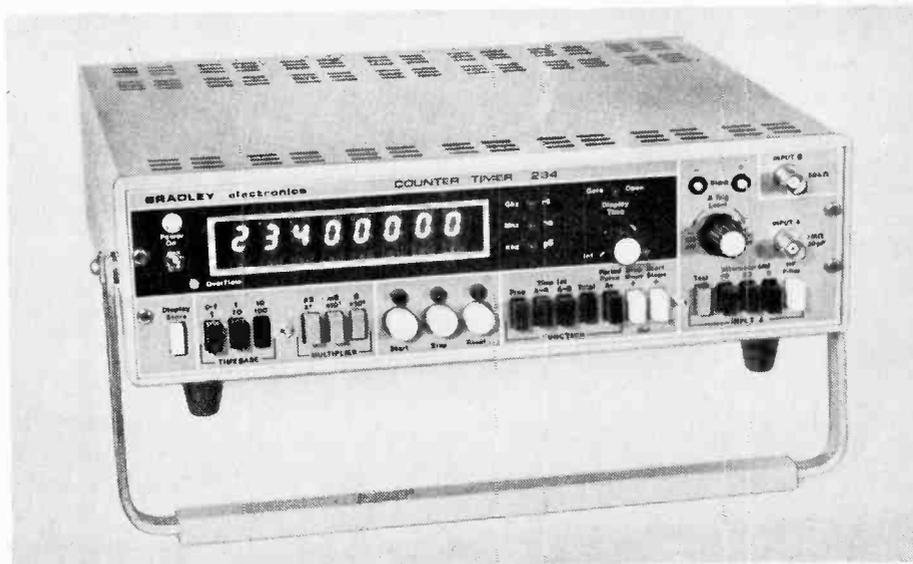


Counter timer

The Bradley type 234 two-channel counter timer will operate at frequencies up to 100MHz and has an 8-digit display. It provides the following measurement functions: frequency, totalize, period, period average, ratio, time interval, and time interval average (down to 25ns). The 10MHz clock is temperature compensated, or as an option, a stable clock in a temperature controlled oven is available. A third channel is provided for an external clock standard, or for ratio measurements.

There is push-button selection of timebase and multiplier ranges, and of functions and attenuator settings. Trigger indicator lamps, range indicator lamps, and gate and overflow indicators are provided to assist the operator. An h.f. filter enables consistent measurements at lower frequencies to be made in the presence of high-frequency interference. Price £259. G. & E. Bradley Ltd, Electral House, Neasden Lane, London N.W.10.

WW512 for further details



Stereo coder

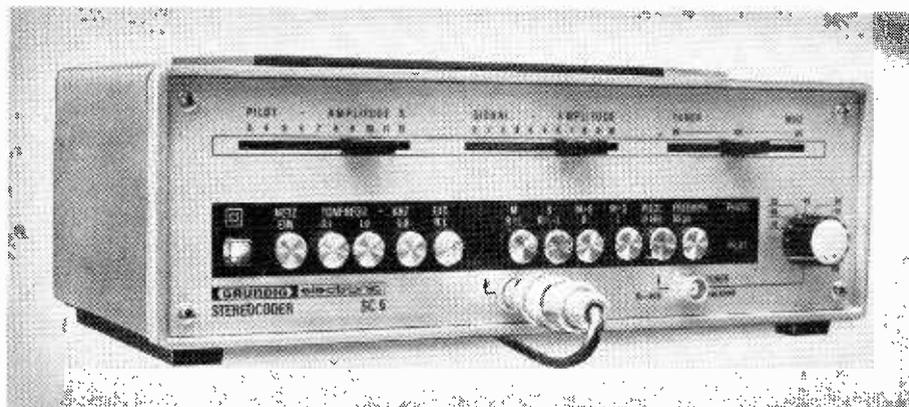
Intended for testing stereo broadcast decoders is a stereo coder SC5 made by

internal modulation	100Hz, 1 or 5kHz, distortion 0.5%
crosstalk	> 40dB, 0.5 to 6kHz
subcarrier level	< 40dB
carrier to pilot	
intermod.	< 50dB
output level	up to 6V pk-pk
input mod. level	0.5V

Grundig. The instrument can be used for testing any stereo decoder with or without v.h.f. carrier. With it, intermodulation and non-linear distortion can be measured, as well as providing the usual simple test.

Pilot tone frequency is crystal controlled to ± 1 Hz. Price is £110. Grundig (Gt. Britain) Ltd, 40 Newlands Park, London SE26.

WW518 for further details

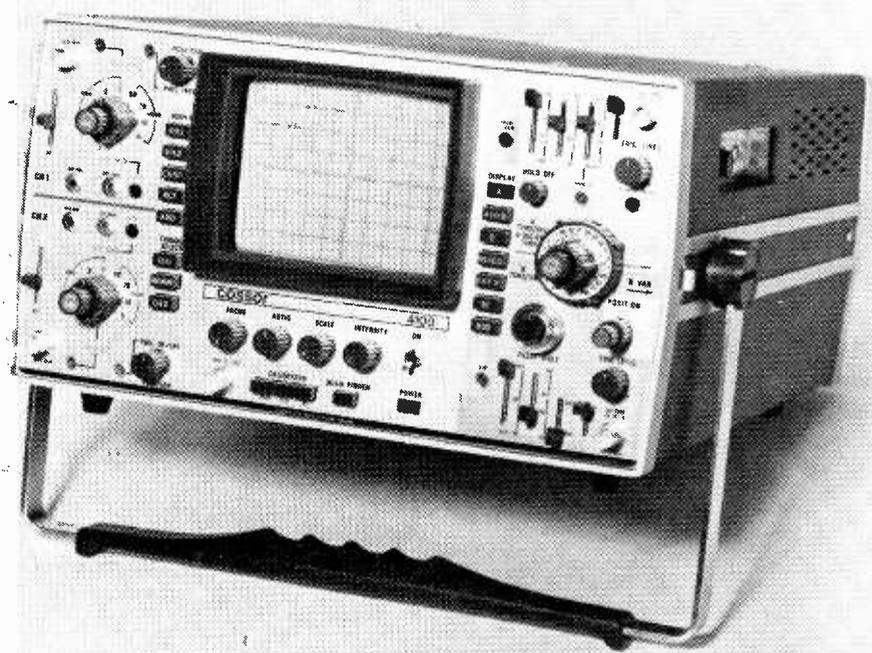


High-frequency oscilloscope

With a large bandwidth of 75MHz, the Cossor Model 4100 dual-trace oscilloscope is intended primarily for use in the computer and digital equipment field. It has a large c.r.t. screen area, 8 x 10cm, for the size and weight of the instrument, which are, respectively, 177mm (high) x 300mm (wide) x 482mm (deep) and 12kg. The c.r.t. (Telefunken) was specially chosen for its short length and Cossor claim they are the first instrument manufacturers to use it. The two traces, identified as A and B, can be used in the following modes of display: A only; A intensified by B; B delayed by A; X-Y; and mixed sweep. The Y amplifiers comprise two channels;

each giving vertical deflection sensitivities of 5mV/div to 2V/div. Dual-channel operation is possible at 1mV/div to a frequency of 25MHz. The timebase sweeps are: timebase A, 0.05 μ s/div. to 0.2s/div. in 21 calibrated steps, a variable control covering between the steps and extending the range to 0.5s/div.; timebase B, 0.05 μ s/div. to 0.1s/div. in 20 calibrated steps, the variable control covering between steps and extending the range to 0.25s/div. A warning lamp indicates uncalibrated settings. Price £750. Cossor Electronics Ltd, The Pinnacles, Elizabeth Way, Harlow, Essex.

WW513 for further details



Catching that elusive transient

Single voltage transients, repetitive waveforms or slowly changing signals can all be recorded in the transient recorder, (type DL905) shown by Datalab, and viewed later on an oscilloscope or reproduced on a chart recorder. If wished the sampled waveform can be fed to a computer in binary form or, with optional accessories, to an X Y plotter; to a tape punch; to an ASR33 electric typewriter; to a signal averager or any other signal analysis system. The advantage of this last option is that the DL905 effectively increases the frequency response of the signal analysis system. A further option provides a b.c.d. output.

The principle of this instrument was mentioned in the Physics Exhibition report last month but as many engineers may not be fully conversant with this sort of equipment a fairly complete description follows.

Basically the input signal is converted into digital form in an 8-bit analogue-to-digital converter capable of operating at up to 5MHz. The digital information describing the input signal is then stored in a 1,024 word semiconductor m.o.s. memory. Once stored the data in the memory can be read out in digital form via one of the optional interfaces mentioned above or can be continuously accessed, 'tail-chase' fashion, for display on an oscilloscope via an internal digital-to-analogue converter.

The controls provided on the instrument are very similar to those found on an oscilloscope. There is an input attenuator with 12 settings from 0.01 to 50V and an attendant offset control. Nearly all the remaining controls are associated with the two timebases (sweeps A and B) although there are some output controls which influence the readout instrument being used; punch/plot, trace invert, trace expansion (horizontal) and X and Y shift.

There are three main modes of operation:

Delayed sweep: A trigger pulse starts a recording sweep after a delay which can be set on a three-decade switch from 0 to 9.99 times the setting of the sweep A control. During the sweep the signal at the input is sampled and stored at intervals determined by the sweep A control. After 1024 samples have been taken the display mode is started and signals can be viewed on an oscilloscope.

Switched sweep: Once triggered a recording sweep is carried out at a rate determined by sweep A and the number of samples to be stored at this rate is set on the delay three-decade switch; the remainder of the memory is filled with samples taken at a rate defined by sweep B. A change of trace intensity on the oscilloscope display identifies the point of changeover between the two sweep rates.

Pre-trigger recording: This mode eliminates the use of cumbersome delay lines and tape loops for recording spontaneous waveforms or signals occurring prior to a

trigger. When the DL905 is set, continuous recording commences at a rate defined by sweep A; this ensures that data is already in the memory before the arrival of a trigger pulse. The delay control defines the number of pre-trigger samples to be kept when a trigger arrives, the remainder of the memory being filled with information taken during and after the trigger point. A change in display intensity indicates the pre-trigger portion of the record.

- input impedance — 1M Ω , 24pF
- frequency response — d.c. — 3MHz
- resolution — 8 bits, 1 part in 256
- settling time — 200ns
- external trigger — $\pm 5V$ ($\pm 50V$ ext. $\div 10$)
- sampling rate — 5MHz max.
- sweep A and B — 200 μs to 10s in 1, 2, 5 sequence (refers to time taken to obtain 1,000 equally spaced samples)
- size — 233 \times 133 \times 417mm
- weight — 9kg

Data Laboratories Ltd, 28 Wates Way, Mitcham, Surrey CR4 4HR.
WW501 for further details

Mains checker

A small instrument for detecting and recording mains voltage transients above a preset threshold and for measuring mains frequency and voltage was shown by Cole Electronics. The instrument incorporates a meter for the voltage and frequency measurements and an electromagnetic counter for recording transients. The threshold of the transient detecting circuit can be set by push buttons to one of four values above the instantaneous value of the mains. If the transient exceeds the threshold for more than 1 μs (with a rise time of faster than 25 μs) the counter will register one count, a light will flash and a +5V pulse will be available for external equipment.

- threshold level settings — 10, 50, 100, 200V
- threshold accuracy — $\pm 10\%$
- max. counter indication — 9,999
- max. count rate — 10Hz pulses/s
- voltage measurement range — 200-260V a.c., peak sensing, r.m.s. calibrated. (also available 100-130V)
- accuracy 10-40 $^{\circ}C$ — $\pm 1\%$ of reading
- frequency measurement range — 35-65Hz
- accuracy — $\pm 1\%$ of reading
- input — 200-260V a.c., 6VA, 35-65Hz
- size — 190 \times 120 \times 75mm
- weight — 1.5kg

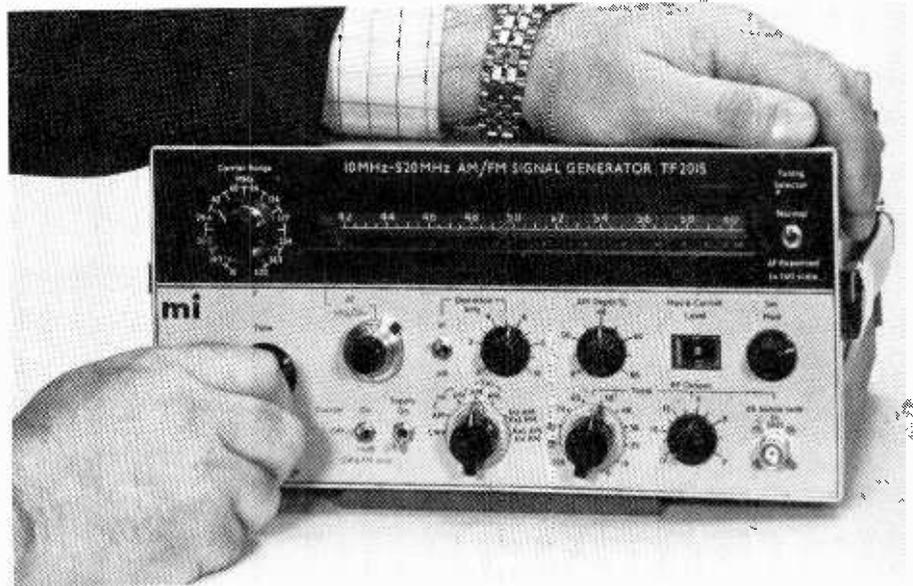
Cole Electronics Ltd., 7-15 Lansdowne Road, Croydon CR9 2HB.
WW502 for further details

A.M./F.M. signal generator

A prototype of a general purpose a.m./f.m. signal generator covering 10 to 520MHz, in 11 switched bands, was shown by Marconi Instruments. The TF2015, as it will be designated, is expected to sell for less than £600. As can be seen from the photograph it is very compact, measuring 140 \times 286 \times 311mm, and weighs 5.4kg. Fundamental frequency generation by voltage tuned oscillators is used throughout. Here are some facts from the specification:

- Amplitude modulation**
- depth — continuously variable up to 80% by directly calibrated control
- calibration accuracy — $\pm 5\%$ mod. depth at 30%, rising to $\pm 10\%$ mod. depth at 80%
- envelope distortion — less than 5% at modulation depths up to 30%

- internal frequency — 1kHz $\pm 10\%$
 - external freq. range — 20Hz to 20kHz
 - Frequency modulation**
 - deviation — continuously variable in two ranges with full scale settings of 10 and 100kHz $\pm 15\%$ of full scale at 1kHz
 - calibration accuracy — less than 2% using internal modulating oscillator
 - envelope distortion — 1kHz $\pm 10\%$
 - internal frequency — 30Hz to 100kHz
 - external freq. range — 30Hz to 100kHz, within characteristics $\pm 1dB$ of the response at 1kHz
- Marconi Instruments Ltd, St. Albans, Herts.
WW530 for further details

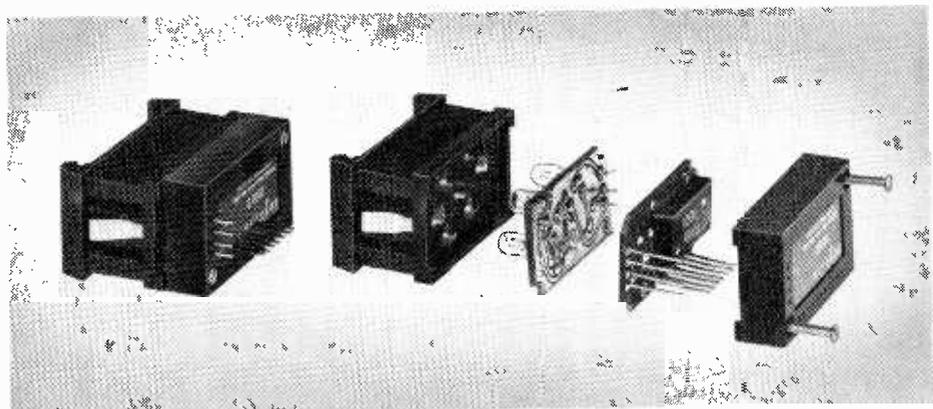


Wide-angle number displays

A feature of the high luminance displays made by Spectra-Tek is the wide viewing angle of 140 $^{\circ}$, made possible by a specially shaped diffuser. The seven-segment display has characters 25-mm high. Segment joints when illuminated are very neat and unobtrusive. Using incandescent lamps, each module includes a thick-film hybrid microcircuit as a driver/decoder, directly compatible with 5-V logic systems. Decimal points are included and polarizers reduce

glare. Life is expected to be 100,000 hours (11 years) at 11V. These type 2500 displays form the basis of type 714-B 3 $\frac{1}{2}$ digit analogue-to-digital converter (£55 one-off), general-purpose DS723 3 $\frac{1}{2}$ -digit panel meter which uses a novel auto-zero feedback network, and DS724 4 $\frac{1}{2}$ -digit panel meter. Displays are also available with 14mm-high characters. Spectra-Tek U.K. Ltd, Kirkby Moorside, York YO6 6DW.

WW522 for further details



Range of a.m./f.m. receivers

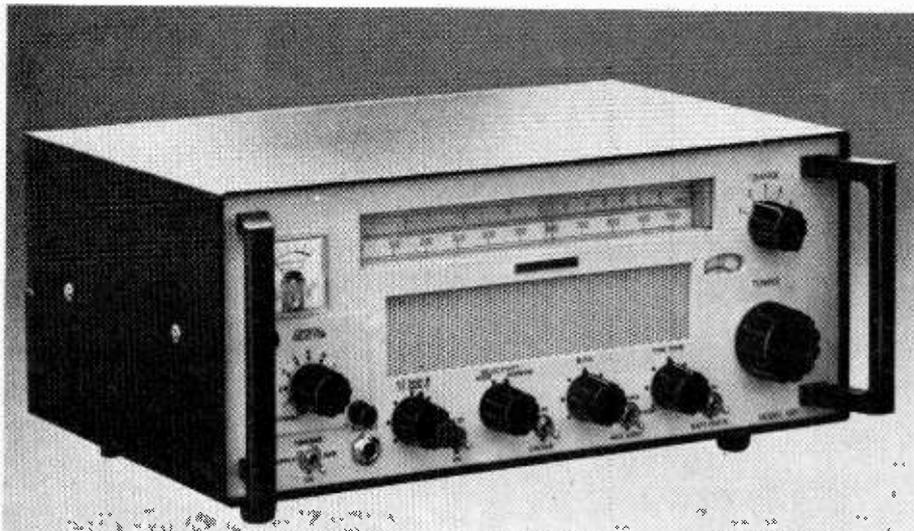
Eddystone Radio were showing a new range of receivers, the 1000 series, based on a modular form of construction which facilitates the production of any one of five models. Approximately 70% of the circuitry is common to all five models, all of which are completely solid-state. The five models, which are priced in the U.K. at between £200 and £300, are:

- 1000—basic h.f./m.f. communications receiver
- 1001—as 1000 but with up to 10 crystal-controlled spot frequencies (shown)
- 1002—a.m./f.m. broadcast receiver covering h.f., m.f. and v.h.f. bands with provision for stereo reception
- 1004—marine receiver covering 150-535kHz and 1.6-30MHz in seven ranges
- 1005—general purpose, crystal controlled, receiver with provision for facsimile reception.

Here is a summary of the data for the 1000:

frequency coverage	550kHz to 30MHz in five ranges
i.f.	455kHz
reception modes	c.w., m.c.w. and 'phone (s.s.b.)
power supplies	a.c. 100/120V or 200/250V (40-60Hz); d.c. 12V internal rechargeable NiCd battery
sensitivity	15 μ V (550-1500kHz), 5 μ V on all other frequencies for 15dB s+n/n
audio output	internal speaker, 800mW at 5% distortion
dimensions	137mm high \times 335mm wide \times 242mm deep
weight	8.2kg inc. battery

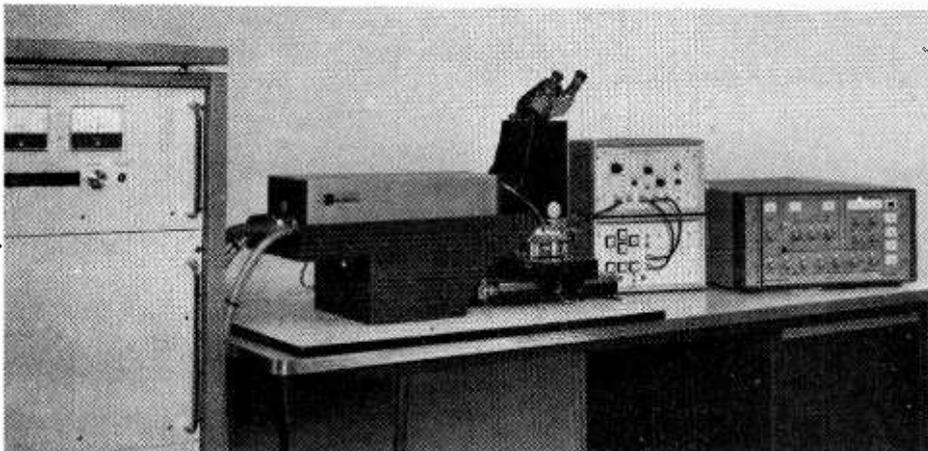
Eddystone Radio Ltd, Alvechurch Rd, Birmingham B31 3PP.
WW528 for further details



Thick-film resistor trimmer

Resistor trimmer model 454B by Laser Associates is intended for thick-film circuits and can also scribe alumina and silicon sheets. It incorporates a new acousto-optic Q-switch enabling the otherwise c.w. neodymium-doped YAG laser to be switched at high repetition frequencies to give a peak power of 10kW at 1kHz. Pulse length is about 200ns. Repetition rate is variable from 50Hz to 50kHz. The

optical system is the same as used on the earlier lower-power 454C, and consists of a beam expander, a 45° mirror and a cutting lens. The \times 50 binocular viewing head enables cutting to be viewed through this lens via a safety filter. Width of cut can be between 10 and 100 μ m. The trimmer needs a water supply. Laser Associates, Paynes Lane, Rugby, Warwickshire.
WW525 for further details



R.F. signal generator

AMF Venner introduced the 754 r.f. signal generator intended for work on m.f. and h.f. a.m. receivers. Continuous coverage from 85kHz to 100MHz is provided in six, push-button selected, ranges. An integral digital frequency meter, with its own crystal reference oscillator, gives a four digit (+1 overflow) indication of the output frequency. The carrier oscillator operates on fundamentals only and frequency multiplying methods to achieve the coverage are not used.

ranges	—1, 85-310kHz; 2, 310-1,000kHz; 3, 1-3.1MHz; 4, 3.1-10MHz; 5, 10-31MHz; 6, 31-100MHz
stability	—0.01% after 15 mins
setting accuracy	—better than 0.1%
reference oscillator	—100kHz crystal, stability 5×10^{-5}
output	—1 μ V to 1V across 50 Ω
attenuator	—pushbutton providing 1, 2, 3, 6 and five 20dB steps plus >10dB variable control
leakage radiation	—< 1 μ V
internal modulation	—a.m., 1kHz variable to 30% depth, panel meter indication
external modulation	—a.m., 100Hz to 100kHz when carrier above 10MHz; 100Hz to 10kHz when carrier below 10MHz; depth variable to 30%
meter	—switchable to indicate carrier level, modulation depth. Carrier level in volts p.d. for zero attenuation and dB relative 1V
temperature range	—-10-55°C
size	—320 \times 360 \times 105mm
weight	—6kg

AMF Venner, Kingston By-Pass, New Malden, Surrey.

WW503 for further details

Hall-effect switches

The new Sprague ULN-3000 devices include a Hall-effect switch and a Schmitt trigger on one integrated circuit chip. Switching is dependent on the proximity of an external magnet whose magnetic flux passes through the Hall cell perpendicular to the chip face. The ULN-3000 device has a maximum activation flux of 750 gauss and a minimum deactivation flux of 100 gauss. Since these new Hall-

effect switches have no moving parts, they do not suffer from contact contamination and can operate at speeds in excess of 100kHz indefinitely. The output of the ULN-3000 is a grounded emitter open-collector transistor. If the magnetic field is below the deactivation gauss level, it is "off" and goes "on" when the field increases to the activation point. The analogue voltage generated by the internal Hall cell is converted to a digital output by a trigger circuit and amplified. In the off state, the output is capable of sustaining up to +5.5V d.c. with less than 100 μ A leakage current. Devices rated for a 12V level are available on special request. In the on state, the output will sink 20mA with a voltage drop less than 400mV. Nominal supply voltage is +5V d.c. variable between +4.75V d.c. and +5.25V d.c., and the ambient operating temperature range is 0 to 70°C. Three packages are available: plastic 8-lead dual in-line "M", plastic 8-lead double-ended "R", and 4-lead single-ended plastic "S". In the "M" and "R" package, the ULN-3000 may have an optional internal 2k Ω pull-up resistor. This facilitates direct interfacing with most d.t.l./t.t.l. logic circuits. Complete data on the ULN-3000 Hall effect switch is given in Sprague Electric engineering bulletin 27 402 and supplementary applications notes. Sprague Electric (U.K.) Ltd, 159 High Street, Yiewsley, West Drayton, Middlesex.

WW514 for further details

Variable phase function generator

A unique feature of a new l.f. function generator by Prosser Scientific Instruments is that phase between any of the available functions is variable. Designed for control engineering applications, the A104 variable-phase waveform generator is useful too as a general-purpose laboratory generator. Sinusoidal, rectangular and triangular waveforms are available from almost zero frequency up to 100kHz. A frequency sweep range of 100:1 is available, and both frequency and phase can be remotely controlled. Linearity of the waveforms is $\pm 1\%$ or better, and rise and fall times of the rectangular wave are 100ns. Various other features are provided. Prosser Scientific Instruments Ltd, Lady Lane, Hadleigh, Ipswich, Suffolk IP7 6DQ.

WW521 for further details

Digital multimeter

The makers of the well-known Avometer have a new digital multimeter, the DA114, which succeeds their previous type DA112. It has a four-digit display, the decimal point being automatically positioned at the appropriate place, and the various measurement functions and ranges are selected by push-buttons. There is an internal calibration system based on a

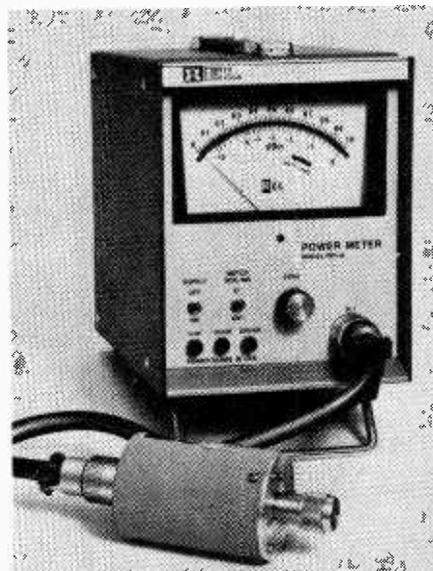


zener diode, each instrument having its own particular calibration voltage, e.g. 1.057V. Voltage, d.c. and a.c., can be measured between 100mV and 1kV full range; direct and alternating current between 100 μ A and 1A full range; and resistance between 100 Ω and 1M Ω full range. Two versions are available: the DA114-M for mains only operation (price £96.50, weight 2kg); and the DA114-B for battery/mains operation with automatic switch-over to the alternative supply (price £126.00). Avo Ltd, Avocet House, Dover, Kent.

WW511 for further details

Microwave power meter

REL Equipment and Components Ltd announced its first 'own label' product — a microwave power meter, available in either an analogue (RR1-A) or digital (RR1-D) readout version. Various power ranges are available depending on the thermo-electric power head used (0.01 — 1mW, 0.1 — 10mW and 1 — 100mW). The standard frequency range is 10MHz to 12.4GHz, but other ranges can be



provided. Power head consists of a thermo-electric calorimeter (which converts the input r.f. power to a low-level d.c. signal) and a linear i.c. amplifier which produces a d.c. current proportional to the r.f. input. The meter measures 165 \times 135 \times 254mm and weighs 3.4kg. REL Equipment and Components Ltd, Croft House, Bancroft, Hitchin, Herts. WW529 for further details

Teaching aid

Basic electricity, semiconductors and transducers are subjects covered by a teaching aid introduced by Feedback Instruments Ltd. Future additions will extend coverage to feedback and automatic control, digital switching circuits, communications principles and environmental technology. Each kit consists of three or four modules which plug in to a console containing power supply and signal busbars, and a power supply unit providing outputs from -15V to +40V d.c. and up to 20V a.c.

Two comprehensive manuals are supplied, one for the student and one for the instructor. Feedback Instruments Ltd, Park Road, Crowborough, Sussex. WW505 for further details

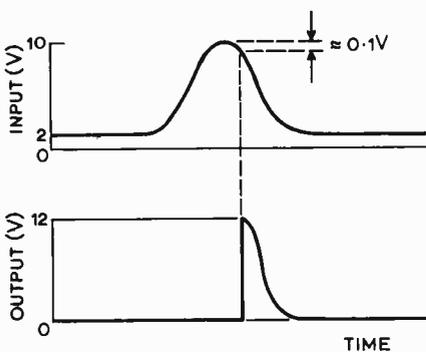
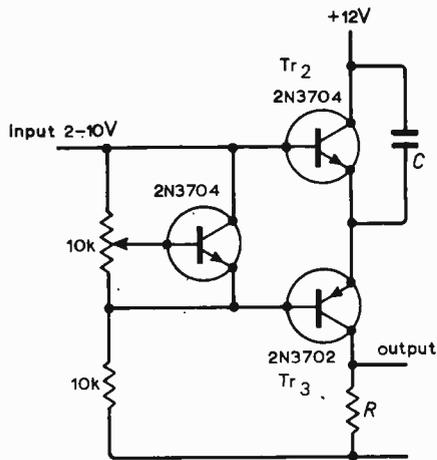
I.Cs for calculators

The development of integrated circuits with high packing densities has led to considerable reductions in the cost, size and complexity of electronic desk calculators. Examples shown by Mullard are the m.o.s./l.s.i. circuits in a series which has been developed specifically for manufacture of calculators. The FDY series contains 15 integrated circuits that are supplied in sets; each set containing all the functions needed for one type of calculator. The simplest calculator is made with set EDC101/EDC111, which contains four integrated circuits: it can perform all the basic types of calculations — add, subtract, divide and multiply — with eight- or sixteen-digit inputs and produce an eight- or sixteen-digit output, according to choice. A more complex calculator can be made with set EDC120 and was demonstrated at the exhibition. This accepts 14-digit inputs and produces a 14-digit answer; in addition to the normal calculating processes it will also compute squares and square roots. Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

WW508 for further details

Further information

Professional readers requiring further information on the products mentioned in this report are invited to use our reader enquiry service for which cards will be found elsewhere in this issue.

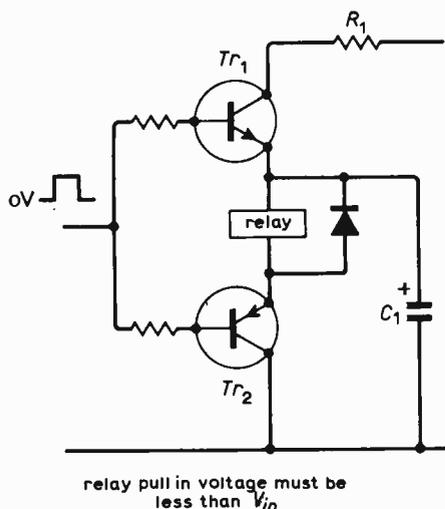


being suitable for audio frequencies, but they are not critical. Output impedance is high so a buffer stage will be needed if feeding into a low impedance.

P. E. Pinnock,
Brunel University.

Long-pulse relay trigger

When equipment needs to be relay-operated momentarily, after a preceding long monostable pulse, the circuit shown can be employed. The positive monostable pulse allows C_1 to charge via R_1 and Tr_1 . (Tr_2 is



off.) When the input returns to 0V, Tr_1 switches off and Tr_2 on, thereby discharging C_1 through the relay which energizes. As C_1 discharges, the current is reduced to below the relay holding current. The circuit is useful in battery-operated equipment as no standby current is required.

R. L. Bratby,
Oxford.

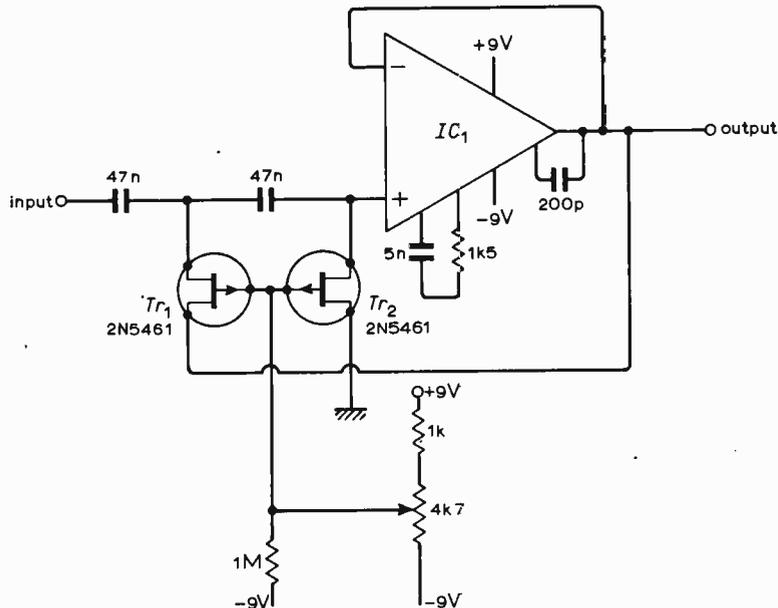
Voltage-controlled high-pass filter

The oscillator circuit shown is a version of the standard op-amp multivibrator adapted for voltage control by using a f.e.t., which gives an approximately logarithmic frequency-to-voltage characteristic. Its main disadvantages are the output shape, which is a short pulse, and the need to use low supply voltages to increase the frequency range.

The high-pass filter is also adapted for

f.e.t. voltage control. Cut-off (3dB down) is approximately from 100Hz to 10kHz at 12dB per octave. It is possible that a pair of commercially matched f.e.t.s could be used but this has not been tried. The integrated op-amp can be a $\mu A709$ (SGS), SN72709 (Texas) or SN72709DN (Fairchild).

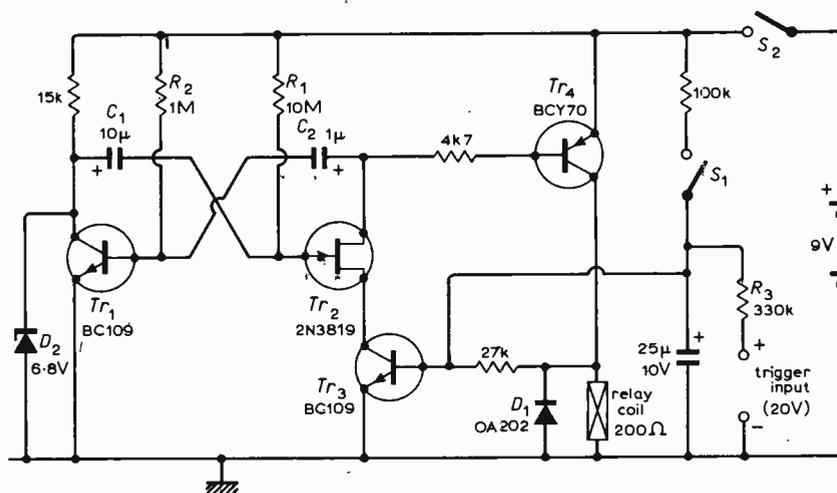
David G. Malham,
University College, Cardiff.



Monostable with long adjustable recovery

In most monostable multivibrators, the recovery time during which the circuit cannot respond to a second trigger pulse is ideally zero, but for some requirements (e.g. the operation of a relay from trigger pulses with a known maximum repetition rate and obscured by shorter pulses of a random nature the situation may demand a recovery period of many minutes. Further, if the multivibrator can be triggered like a Schmitt trigger when the input voltage exceeds a certain level, the recognition of a genuine trigger pulse can be accomplished

simply by integration of the input signal. These properties are combined in the circuit shown which uses only four transistors and has a current drain of about 250 μA . The heart of the circuit is a cross-coupled astable multivibrator in which the active devices are a bipolar (Tr_1) and n-channel f.e.t. (Tr_2). Transistor, Tr_3 is used as a gate and the f.e.t. is directly coupled to the relay-driving stage, Tr_4 . The 'dead' time can be as long as 1 hr with a good quality C_1 .
D. A. Tong,
Leeds.



Electronic Building Bricks

25. Logic gates

by James Franklin

Suppose we wish to control an electric passenger lift so that the lift motor will start running only when a button in the lift is pressed and the lift doors are closed. There is a door sensing switch which is actuated when the doors are closed. Let us first list the possible states of the button, the door sensing switch and the motor:

- Button not pressed ○
- Button pressed ●
- Door switch not actuated ◇
- Door switch actuated ◆
- Lift motor not running □
- Lift motor running ■

The symbols on the right are a 'shorthand' to represent the states expressed by the words.

To clarify exactly what our control device, Fig. 1, must do we can use the symbols to tabulate all the possible combinations of states of the button and switch, and alongside them, on the right, put the corresponding *required* states of the lift motor, thus:

a	○	◇	□
b	●	◇	□
c	○	◆	□
d	●	◆	■

With condition *a* both the button and the switch are not operated, and, of course, the lift motor must not be running. In *b* the button is pressed but the door switch is not actuated, so again the lift motor must not be running. In *c* the door switch is actuated but the button is not pressed, so once more the motor must not be running. In *d*, however, both the button and the door switch are operated, so the motor should be running.

This logical breakdown of the required behaviour of the control system, is known as a 'truth table', from its similarity to the truth tables used in the study of deductive logic. Here lies the reason for the use of the term 'logic' in electronics. The table, in fact, could be written in terms of the *T* (true) and *F* (false) letter symbols commonly used in deductive logic, since we are concerned here with control devices which can have only two states (i.e. on and off). For example, if at a given moment the statement 'the button is pressed' is true (*T*), the contradictory statement 'the button is not pressed' must be false (*F*), since the button cannot be pressed

and not-pressed at the same time. So, to express the two conditions of the button in this symbolism we need only one statement, say, 'the button is pressed', which can be made either *T* or *F*. Similarly for the door switch being actuated and the lift motor running. Using this convention the truth table would be:

	●	◆	■
a	<i>F</i>	<i>F</i>	<i>F</i>
b	<i>T</i>	<i>F</i>	<i>F</i>
c	<i>F</i>	<i>T</i>	<i>F</i>
d	<i>T</i>	<i>T</i>	<i>T</i>

where the first column refers to the button, the second to the door switch and the third to the lift motor, and the truth functions* *a*, *b*, *c*, *d* correspond to the conditions *a*, *b*, *c*, *d* of the 'requirements' table.

It happens that this second table is the general truth table for what is termed a *conjunction* in deductive logic. As the name indicates, a conjunction is an expression consisting of two statements linked by the concept 'and'. This 'and', sometimes abbreviated to '&' and sometimes to '.', is called a truth-functional* *operator*. Examples are: [the grass is green] & [the flowers are blooming]; [the button is pressed] & [the door switch is not actuated]; or, more generally, *p* & *q*. The truth table displays all combinations of true and false statements, describing states-of-affairs, that are possible with a conjunction.

We see, then, that the lift control requirements set out in the first table are satisfied by the truth table of the 'and' truth-functional operator. So we need an electronic device that operates according to the conditions in this truth table. In fact this is provided by a device acting similarly to the electronic gate described in Part 23. As will be recalled, this gate can be used as a coincidence detector—an output signal being produced when, and only when, two input signals are present simultaneously. A unit of this kind working to the rules of the 'and' truth table is called an 'and gate' and its use

*What is meant by 'truth function' or 'truth-functional' is that the 'truth' is made a function of something, e.g. a truth-functional interpretation of a compound statement (examples of which are given above) takes the truth or falsity of the compound to be a function of the truth or falsity of its component statements.

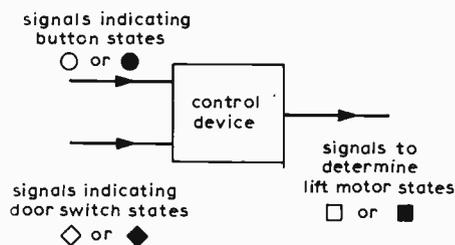


Fig. 1. Schematic of a system for controlling a lift motor from a press-button and a door sensing switch. (See text for meaning of 'state' symbols.)

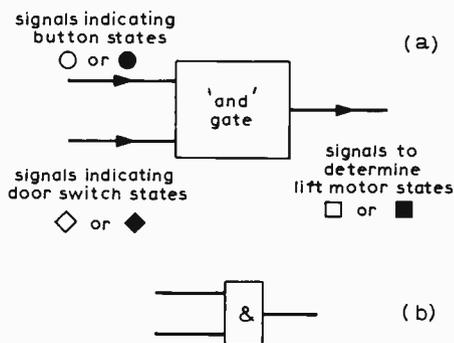


Fig. 2. The device which performs the required action is an 'and' gate (a), the logic symbol for which is shown at (b).

in the lift control problem is shown schematically in Fig. 2(a). The generalized symbol for this 'and' gate is shown at (b).

In practice logic gates can be constructed with electromechanical relays (or indeed purely mechanical devices using mechanical variables as signals), but electronic gates are widely used because of their relative cheapness, small size and high speed of operation. The operating 'signals' shown in Fig. 2(a) are usually different voltages, for example a high voltage signifying *T* in the truth table (e.g. button pressed ●) and a low voltage signifying *F* (e.g. button not pressed ○). There are, in addition, other 'logic gates', with actions corresponding to other truth-functional operators in deductive logic, but we shall deal with these in a later article.

World of Amateur Radio

V.H.F. and mobile repeaters

The decision by the M.P.T. to allow a "controlled experiment" in f.m. talk-through repeaters in the U.K. (the first will almost certainly be at Cambridge) is causing a good deal of heart-searching in v.h.f. circles. Such repeaters can clearly be of great benefit to mobile operators and those who live in bad v.h.f. sites, since they gain the advantage of being able to "use" a high aerial and the other facilities. But there is some evidence that once a repeater is in use, it removes much of the incentive to improve home equipment and tends to concentrate the local activity. Many amateurs restrict their activities to the repeaters, using rigs just good enough to actuate the repeater and to receive its signals; repeaters could also present problems in tracing local pirates.

But there is no doubt that repeaters have become popular in some countries. A recent directory of North American repeaters showed 362 at 321 sites (269 in the United States and 52 in Canada) mostly on the 144-MHz band. There must also be 20 or more in West Germany, mostly accepting calls on 144.15MHz and retransmitting them on 145.85MHz, actuated by a tone calling frequency around 1750Hz. The latest M.P.T. figures show there are over 3300 British mobile licences (2702 class A, 618 class B).

M.P.T. has recently confirmed that amateurs not holding mobile licences may operate stations from stationary vehicles provided that the equipment is rendered incapable of being used while "being transported". Disconnecting the power supply will meet this requirement.

The cancellation of the ITOS-C satellite means that the Oscar 6 amateur satellite is now postponed until about November, when it may be launched with Nimbus-E into a circular 680 statute mile orbit, with 108 minute period and 99.9° inclination.

Following consultation with the R.S.G.B. a £100 Mullard award for forward-looking technological work by radio amateurs, with emphasis on v.h.f. and u.h.f. or space communications is to be initiated.

Heath Rees, G3HWR, closely associated with the GEC club station G5FK, is the first recipient of the J. Fraser Shepherd microwave award.

The attendance at the 1972 v.h.f. convention, at Whitton, Middx, reached an all-time record of 437; next year's convention will be on April 7th. Microwave equipment shown at the convention included a crystal-locked Gunn diode transmitter and a 3-cm television link. A Scottish v.h.f. convention is being held on Sunday, October 1st at the Shawlands Hotel, Glasgow S1, organized by West of Scotland Amateur Radio Society.

Trend to transceivers

The conveniently packaged single-sideband h.f. transceiver began to make an impact on the amateur scene in 1959 when the Collins KWM-1 was marketed; but for some years the number of such rigs actually in use in Europe remained small. This was partly cost (a complete KWM-1 cost over \$1500) and partly because the lower cost units, which soon began to appear, often covered only one or a few bands, or segments of bands, and in other ways were less flexible than the traditional separate transmitter-receiver combination. But gradually even the lower cost equipments have become more versatile and more effective (though sometimes less than ideal for c.w. operation).

The KW2000B with its 1.8 MHz facility represents a popular British contribution to this highly competitive international market, but quite a few long-established U.S. firms have recently retreated or succumbed under the onslaught of the Japanese, with Yaesu, Musen and Trio offering many different models. Most home-construction is from Heath kits or from designs described in journals, such as David Bowman's currently appearing in *W.W.*

Many transceivers are now all-semiconductor apart from a driver and power amplifier valve stages — though the way still question the performance of some all-transistor receiver front-ends, despite the increasing use of dual-gate field-effect devices. For s.s.b. generation, the filter technique is generally used, though with a wide range of filter frequencies: some use mechanical filters at 455kHz, but more models now have crystal filters around 3180, 5200 or 9000

kHz. Many use TV line-output valves in the final amplifier, taking advantage of the low duty cycle of s.s.b. These need to be handled with rather more care than the models using 6146B transmitting-type valves.

Recent market surveys in the United States suggest that amateurs there are now buying transceivers in substantially larger numbers than separate receivers and transmitters — and this is probably true also in the U.K. and Europe.

Australian 70-cm moon-bounce

A new Australian v.h.f. newsletter "Six-up" reports that moonbounce signals have been exchanged on 432MHz between VK2AMW at Dapto, near Wollongong, and WA6HXW in California. The Australian equipment has 400W output at 432MHz with a stability better than one part in ten million, and uses a 30-ft diameter parabolic dish aerial (28dB gain over dipole) polar mounted and controlled from the operating shack about 100 yards away. The project, led by Lyle Patterson, VK2ALU, has been made possible by reactivating former radio telescope facilities of Wollongong University College.

In brief

Speakers at the ambitious North and West Amateur Radio Convention — planned as a weekend residential meeting but open also to one-day visitors — to be held at Lancaster University on September 23-24 will include 'Dud' Charman, G6CJ (aerials), J. M. Bryant, G8FNT of Plessey (using integrated circuits), Geoff Stone, G3FZL (v.h.f.), Dr Max Lazarus (microwaves) and R. Coates (radio teleprinting). There will be a talk on Racal receivers, an interference forum, trade exhibits, technical films and the like. Booking forms and details from: P. Jones, University of Lancaster Amateur Radio Society, Department of Environmental Sciences, The University, Lancaster . . . Barking Radio and Electronics Society (G3XBF) is operating an exhibition station, GB2DTS, at the Dagenham Town Show, Central Park, Dagenham on July 8 and 9, on all bands 1.8 — 432MHz with talk-in facilities on 1.8 and 144MHz. Visiting amateurs welcome . . . M.P.T. seem likely to introduce a three-year amateur licence to reduce administrative costs . . . In a recent 3.5-MHz low-power contest Alan Wybrow, G3JVJ, made 45 contacts using an inverted-V dipole with his TT11 transmitter drawing 6.5mA from a 70V supply . . . George Pataki, WB2AQC (former YO2BO) and his wife Eva (WA2BAV) are running a "welcoming centre for foreign amateurs visiting New York city". Their address is 34-24 76th Street, Jackson Heights, New York 11372 . . . Almost 50% of the popular summer camps for children in the United States maintain amateur radio activities.

Pat Hawker, G3VA

About People

At the Royal Television Society's annual ball in London on 12th May the Society's gold medal for outstanding contributions to television was awarded to **Georges Hansen**, director of the European Broadcasting Union's Technical Centre in Brussels. Monsieur Hansen, who graduated from Brussels University in electrical and mechanical engineering, joined the Belgian Broadcasting Service and in 1936 became head of research. He served with the Belgian Army throughout the war and spent some years in the United Kingdom as a liaison officer with the British Forces. He returned to Radio Tefusion Belge after the war and became chief engineer and deputy director general in 1946. He has been director of the E.B.U.'s Technical Centre since 1956.

C. J. Dalton, B.Sc., Ph.D., and **J. R. Sanders**, M.A., of the B.B.C. Research Department, have received the Geoffrey Parr Award of the Royal Television Society in recognition of their distinguished work in developing the system known as 'sound-in-syncs'. Designed for distributing television programmes from studio to broadcast transmitters the system accommodates the sound signal, as coded pulses in the video waveform within the line sync pulse period, instead of on a separate channel. At each transmitter the sound signal is separated from the video waveform, decoded and then broadcast in the usual way. Dr Dalton graduated at Bristol University where he obtained his doctorate in electrical engineering in 1960. In that year he joined the B.B.C.'s Research Department. He is now senior engineer in the studio group of the B.B.C.'s Design Department. Mr Sanders graduated in 1955 with an honours degree in physics at Pembroke College, Oxford. After two years' graduate apprenticeship with the B.B.C. he worked on TARIF (Technical Apparatus for the Rectification of Indifferent Film) of which he was co-inventor. After two years with the systems section he transferred to the image scanning section in 1967 and became its head at the end of last year.

Sir Ronald Melville, K.C.B., has been appointed director of the Electronic Components Board in succession to the late **Sir Alan Dudley**. Sir Ronald, who is a Permanent Secretary attached for special duties to the Civil Service Department, will retire from the Civil Service towards the end of the year. He is being seconded by the Government for part-time employment with the Electronic Components Board until the date of his retirement. Educated at Charterhouse and Magdalene College, Cambridge, he joined the Civil Service in the Air Ministry in 1934.

Bernard J. O'Kane, Ph.D., B.Eng., M.I.E.E., chairman of Marconi-Elliott Avionic Systems Ltd, has been re-elected president of the Electronic Engineering Association. A graduate of Liverpool University he joined the Research Laboratories of the General Electric Company in 1935 where he served until 1941 when he was transferred on loan to the Telecommunications Research Establishment at Malvern. He rejoined the G.E.C. Radio Department in 1945. During 1947 Dr O'Kane joined International Aeradio Ltd, as chief engineer, where he remained until his appointment in 1952 as chief air radio engineer with the Marconi Company.

The Royal Television Society's John Logie Baird Travelling Scholarship, sponsored by Radio Rentals Ltd, has been awarded to **Robert Clayton**, B.Sc. He spent a year with the Marconi Company before going to the University of Manchester Institute of Science and Technology where he graduated in 1971. He is at present engaged on a research programme in computer picture processing with an M.Sc. and Ph.D. in view.

P. A. D. Duffell has been appointed director of international product marketing, and will be responsible for the promotion, development and co-ordination of the EMI Group's worldwide interests in consumer durable and commercial electronic products.

Aged 51, Mr. Duffell joined EMI in 1965, becoming managing director of British Domestic Appliances International Ltd and subsequently a director of EMI Electronics Ltd. Immediately prior to taking up his new appointment, Mr. Duffell was managing director of EMI's Sound and Vision Equipment Division, at Hayes, where he is succeeded by **J. F. Willsher**, formerly director of EMI Sound Products, Treorchy, S. Wales.

Plessey has announced the appointments of **J. W. Mermagen** as European marketing manager for Garrard and **R. E. Poulton** as audio products manager. Mr. Mermagen was previously marketing services manager, and Mr. Poulton, U.K. sales manager. Mr. Poulton originally joined the Garrard research and development department. He is a member of Panel T of the R.E.C.M.F. on standards for tape and tape recording and, as a member of the British Standards Institution sub-committee TLE25/2, he was recently concerned with the revision of BS415 on electrical safety requirements of mains-operated domestic sound and vision equipment.

Paul Heelas is the newly appointed assistant manager of Ferrograph's professional equipment division. Mr. Heelas, who is 28, has joined Ferrograph from Ekco Instru-



Paul Heelas

ments Ltd., Southend-on-Sea, where he was assistant marketing manager for the division making computer peripherals and nuclear equipment.

Terry Mann, Grad.I.E.R.E., who joined Coutant Electronics in 1970, has been appointed sales manager. Prior to joining Coutant he was with Advance Electronics for five years. Mr. Mann, who is 32, will be responsible for the company's U.K. marketing. Overseas marketing will continue to be handled by **Derrick Taylor**, Coutant's sales director.

Bernard L. Muller is to retire from his present position as manager of the Domestic Products Division of Bosch Ltd. Mr. Muller, who will

shortly be 60, began in the radio trade with Marconiphone in 1936. He later moved to Cossor, and then in 1948 to Sobell as sales manager. In 1958 he was appointed sales director of Regentone, where he remained until taking up his present position with Bosch in 1965.

Pye TMC Ltd has appointed **David R. Hester**, M.A. (Physics), M.I.E.E., as deputy divisional manager in the New Products Division. He entered the electronics industry in 1956 and has had production and development experience with Mullard and with Bell Telephones' Northern Electric Co. in Canada. Mr Hester spent five years at Plessey's Allen Clark Research Centre, Caswell, where he was responsible for integrated circuit process development and the associated pilot plant, and had latterly been quality assurance manager at Plessey's Swindon factory.

Honorary Fellowship of the Royal Television Society has been awarded **Dr. Gerhard Lubszynski**, now a lecturer at the University of Strathclyde, Glasgow. Dr. Lubszynski was an original member of Sir Isaac Shoenberg's pioneering television team at E.M.I. Much of his original work was on the development of the Emitron tube and the Super Emitron, which he invented. He retired from E.M.I. in 1968.

John Dale-Lace has been appointed marketing director of Advance Filmcap Ltd. Prior to joining Filmcap, he spent seven years at London Electrical Manufacturing Company Ltd (LEMCO) where he was successively sales manager and, since 1967, sales director. He was at one time sales manager of the Radio and Electronic Components Division of A.E.I.

Tom Lane has been appointed U.K. sales manager of Fairchild Semiconductor Ltd which he joined four years ago. He spent 10 years as circuit design engineer with Marconi and four years as a sales engineer with Motorola Semiconductor.

Fane Acoustics Ltd, the loud-speaker manufacturing member of the Audio Fidelity Group, has appointed **David Ferguson** as works manager at its Batley factory. Mr Ferguson, who is 47, was at one time with BSR at East Kilbride, Scotland, as factory manager.

D. W. F. Tatchell has been appointed as a senior executive with L.G.T. of Paris. Mr Tatchell was in charge of de Havilland's Physics Laboratory for 10 years.

New Products

Battery soldering instrument

A lightweight thermally controlled soldering instrument has been introduced by Adcola Products to operate from a standard car battery. The complete tool weighs less than 2oz and has a replaceable plug-in element. Two models are available with soldering bit diameters of $\frac{3}{16}$ in and $\frac{1}{4}$ in, rated at 23 and 27W respectively to provide an operating bit temperature of 360°C. The Invader battery model will normally melt solder within a couple of minutes of connection and will reach full operating temperature in less than five minutes. Crocodile clips are provided at the end of 12ft of p.v.c. supply cable. The Invader $\frac{3}{16}$ in diameter bit model BL 646 retails at £2.37 and the larger model, BL 1076, for £2.47. A range of standard copper and iron plated long life bits is also available. Adcola Products Ltd, Adcola House, Gauden Road, London S.W.4.
WW307 for further details

Acoustic coupler

Up to 120 characters per second can be handled by the new Moore Reed TC-401 acoustic coupler. It is designed for use with Post Office No. 1 Modems and allows high-speed operation of video data terminals such as the Moore Reed VT-107. Receive mode speeds of 600 to 1200bauds are available and the keyboard operates at 75bauds, utilizing a supervisory channel.



An enclosed receptacle, capable of accommodating most modern telephone handsets, provides insulation from outside noise. The TC-401 measures 292mm wide \times 127mm deep \times 218mm high and weighs 3.2kg. Cost of the unit is £295. Moore Reed and Company Ltd, Walworth Industrial Estate, Andover, Hants.

WW304 for further details

Travelling-wave tubes

EMI-Varian has introduced two new travelling-wave tubes as economical output tubes for microwave relay link systems or as drivers in satellite ground station transmitters. Designated VTJ-2609A1 and VTJ-2609A11, together they cover the frequency range of 5.925 to 7.125GHz (type VTJ-2609A1 from 5.925 to 6.425GHz and type VTJ-2609A11 from 6.425 to 7.125GHz). They have a nominal gain of 38dB at an output of 10W. Saturation output is at least 20W. Main features of these t.w.t.s include: p.p.m. focusing; depressed-collector operation for higher efficiency; low a.m. to p.m. conversion factor; and fully interchangeable and pre-focused envelope for fast, simple replacement. The tubes offer exceptionally long life, with mean time between failures greater than 25,000 hours. EMI-Varian Ltd, Hayes, Middlesex.

WW315 for further details

Video signal selector

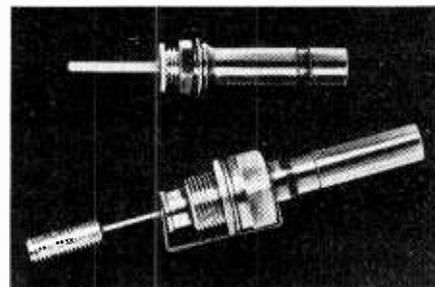
Rohde & Schwarz have introduced a video signal selector type ASF (frequency range 1Hz to 20MHz) as a remotely controllable switching unit, which stores the switching state in the event of a power failure. It has five bridging-filter inputs, which are each provided with a high-impedance, low-capacitance buffer stage, and are selected via a diode switching circuit with transient suppression. The final amplifier feeds five operational outputs and one test output. The signal switching circuitry is controlled by flip-flops and either vertical or sync pulses provide the clock pulses by which the

flip-flops are synchronized during signal switching. If no external signal is provided, then the 50kHz pulses from the readout oscillator cause the immediate transfer of the information stored. U.K. agents: Aveley Electric Ltd, Arisdale Avenue, South Ockendon, Essex.

WW305 for further details

Quartz dielectric piston trimmers

Two high-voltage piston trimmers for use up to u.h.f. are suitable for auto tune in high-power valve amplifiers at v.h.f. and u.h.f., particularly in military/airborne environments. The type AT7550 has a capacitance range of 1.5-26pF. A Q factor of 750 minimum is guaranteed at 400MHz.



The breakdown voltage is 2,200V. The quartz dielectric has a working temperature range of -55°C to $+125^{\circ}\text{C}$ and the overall temperature coefficient is guaranteed 0 ± 50 parts per million per degree C. The AT7551 is a smaller version offering a capacitance range of 0.8-10pF with a Q of 300 minimum at 500MHz. Operating voltage can be as high as 500V. Both trimmers are said to be capable of a minimum of 100,000 mechanical operating cycles. MCP Electronics Ltd, Alperston, Wembley, Middlesex, HA0 4PE.

WW311 for further details

Electrostatic printer

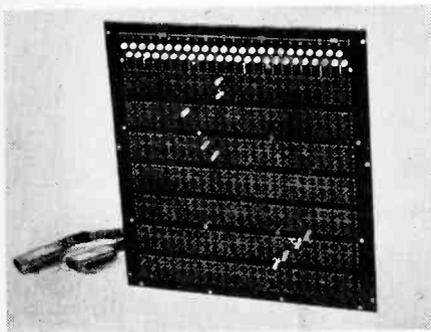
Information displayed on computer video terminals can be obtained in printed form using an electrostatic page printer introduced by SE Computer Peripherals, an EMI subsidiary. This printer can provide ten copies of a complete video screen of data — each copy taking three seconds to produce. The equipment has a maximum speed of 700 characters per second. In addition to its support role with video terminals, the SE printer can be used as a communications terminal with most types of minicomputers. The printer can be supplied with its own buffer store, or memory, if required. This facility enables much faster operation when multiple copies of data are required, by reducing the "down time" on the video terminal or communications line. As the first page copy is generated from the video terminal, the printer also stores this data in its memory to use in producing successive copies, freeing the video terminal or communications

line for further data processing. It can be utilized to serve more than one video display and by the addition of a "printer sharing module" and switching facilities, eight displays can be linked to one printer. The machine uses 6in wide insulation-coated paper. Measuring about 40cm square and 22cm high, the machine is available as a desk top model or equipped with a trolley. It is simple to operate, with only three controls: on/off switch; selector for print quantities; and remote print switch. SE Laboratories (Engineering) Ltd, Feltham, Middx.

WW309 for further details

Matrix plugboard

A plugboard matrix switching panel has been introduced by Sealectro for applications requiring a daily cycle of programmable events. Having a matrix of 48×42 the Sealectroboard can programme 42 functions at half-hourly intervals over a 24-hour period. Operation is achieved by energizing the vertical busbars in sequence



by a clock pulse every 30 minutes directed to a motorized selector switch. By means of interconnecting plugs inserted in the matrix holes at the desired cross-points, any combination of outputs is possible on the horizontal busbars. Indication of the programme status is provided by the pin situations and by indicator lamps above the vertical columns. Sealectro Ltd, Walton Road, Farlington, Portsmouth PO6 1TB. **WW301 for further details**

Solid-state switches

Crown Microwave have introduced a line of solid-state switches which are known as "linearized attenuator modulators". These switches are designed for systems that require a linear (± 1 dB) voltage versus attenuation with extremely fast speeds. As an example these switches typically achieve 10dB/V attenuation from 0 to 5V. Both 40 and 60dB maximum attenuation types are available. Both flatness (typically ± 1 dB across octave and greater bandwidths) versus frequency and the fast speed (insertion loss to maximum isolation in less than 30ns including settling time), make these devices applicable as fast attenuators or amplitude modulators. Tony Chapman Electronics, 3 Cecil Court, London Road, Enfield, Middx. **WW320 for further details**



Programmable attenuator

The Hatfield 2010 series of rack-mounting radio-frequency attenuators comprises units with attenuation values of up to 119.9dB in 0.1dB/ Ω steps at either 50 or 75 characteristic impedance. This design, in addition to 4-line compatible b.c.d. programming, provides one out of ten line code

control, permitting the unit to be remotely programmed by one single pole rotary switch per decade and local programming from thumb wheel switches on the front panel of the instrument. Hatfield Instruments Ltd, Burrington Way, Plymouth, Devon PL5 3LZ.

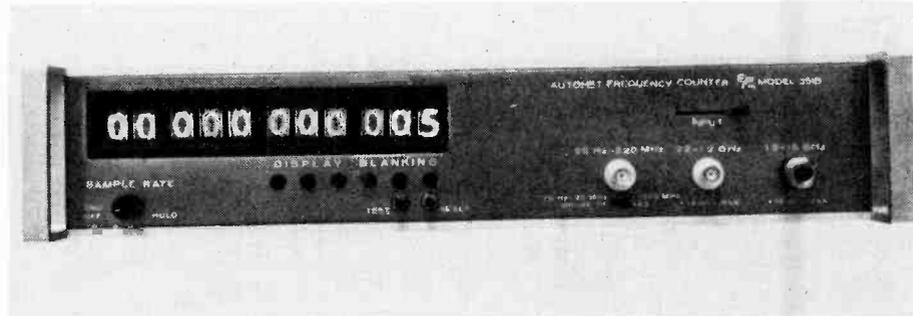
WW322 for further details

Frequency counter

Dana's new fully programmable frequency counter Model 351B gives continuous automatic frequency measurements from 20Hz to 18GHz. It has an 11-digit display, divided into GHz, MHz, kHz and Hz and has a read-out resolution of 1Hz. Accuracy is ± 1 count \pm the crystal timebase stability (5 parts in 10^9 for 24 hours). Greater accuracy can be achieved with an external reference or by using one of the two optional high-stability timebases. Resolutions of 10Hz and 1Hz are attainable at maximum rates of 10 readings per sec. and one reading per sec., respectively. Where

lower resolution will suffice, push-button display blanking simplifies the read-out by turning off unwanted digits. Available options include a high-stability timebase (ageing rate less than 1 part in 10^9 per 24 hours after 72 hours warm up); an "ultra high stability" timebase (ageing rate less than 5 parts in 10^{10} per 24 hours after 72 hours warm-up); parallel b.c.d. output; and programmable i.f. offsets. Price of the basic Model 351B counter is £2,250. Dana Electronics Ltd, Bilton Way, Dallow Road, Luton, Beds.

WW310 for further details



Static inverter

This unit, type M/1000(3)400R, is designed to provide a three-phase 400Hz supply from a single-phase a.c. mains input. The main application is the ground operation and testing of airborne electronic equipment. The output voltage is maintained within the range 108 to 118V (phase neutral) over the range from 150 to 1700W. The frequency stability is $\pm 1\%$, but a crystal-controlled master oscillator can be provided, or provision can be made for locking to an external 2400Hz source. Distortion is below 8%. Screw terminal connections are provided internally, with access through the conduit fittings. An overload trip is fired

at a loading of about 2kW. The automatic slow start-up procedure must be initiated by switching off and re-starting after a short pause. Fuses are provided on the mains input circuits, for the thyristors and on individual phase outputs. Lamps monitor internal supplies, output, phase failure and overload trip. The inverter is mounted as a set of four $19 \times 10\frac{1}{2}$ in panels, a common control and combining panel and three separate phase panels, in a mild steel cabinet 45in high. R. Giffillan & Co. Ltd, Southdownview Road, Broadwater Trading Estate, Worthing, Sussex.

WW314 for further details

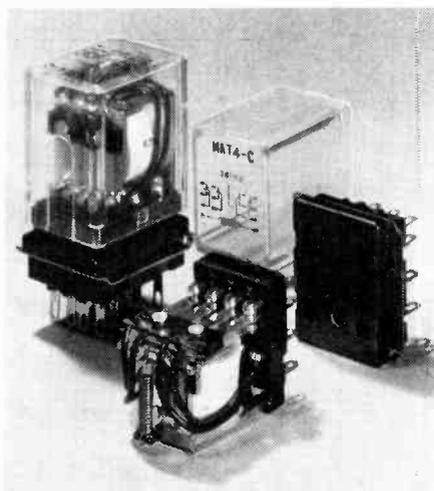
Positional transducer

A contactless positional transducer element, designed to convert angular or rotational displacement into proportional output, has been introduced by Jackson Brothers (London). The conversion is made by capacitance variation. There is no mechanical contact between the moving parts and the effective life is therefore prolonged. Housed in a Mumetal can, to nullify stray external field effects, the output can be arranged to be a.c. or d.c. The output response with input can be linear, sinusoidal, logarithmic or practically any law desired. Although it has an internal 10kHz oscillator, provision can be made for an external source. Size is $1\frac{3}{4} \times 1\frac{3}{4} \times 1\frac{1}{4}$ in and weight 4½oz. Minimum drive torque is 2g.cm. Jackson Brothers (London) Ltd, Kingsway, Waddon, Croydon, CR9 4DG, Surrey.

WW312 for further details

Miniature relay

A miniature a.c./d.c. relay with four in-line contacts and a switching capability of 3A is available from ITT Components Group Europe. The MAT4 relay has nominal operating voltages in the range 6 to 110V

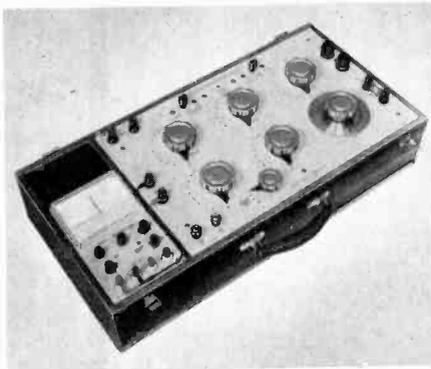


d.c. and 6 to 240V a.c., and is rated at 3A at 30V d.c./115V a.c. A mechanical life of 50 million operations is expected with an electrical life in excess of 100,000 operations at 30V, 3A. Insulation resistance is 1,000MΩ and the dielectric strength is 1,000V r.m.s. Power at 20°C is nominally 1.5W for a.c. versions and 0.9W for d.c. The MAT4 is available with terminations for plug-in or direct soldering and also for printed circuit mounting. ITT Components Group Europe, Electromechanical Product Division, West Road, Harlow, Essex.

WW302 for further details

Wheatstone bridge

Incorporating a built-in electronic d.c. null detector the Wheatstone bridge, type WS3/E, introduced by Croydon Precision Instruments, has an overall range of 0.001Ω to 10MΩ. There are five dials plus a slide-

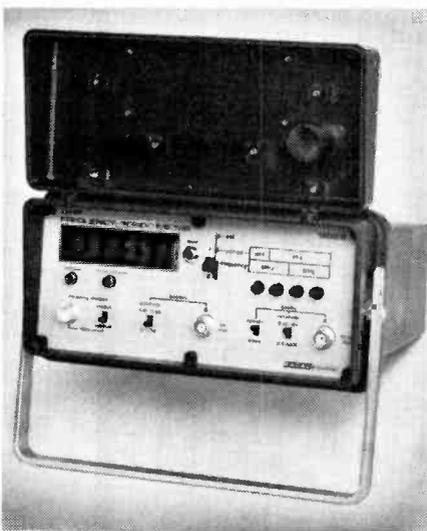


wire, operation is from internal batteries, and accuracy is $\pm 0.01\%$. Switches and coils are mounted on an insulated sub-panel. Features of the bridge include variable battery series resistor control, four pairs of plug switch selected ratios, a centre zero slidewire with adjustable zero setting and provision for short circuiting when unequal ratios are being used. There are seven ranges, from 0-11.111Ω to 0-11.111MΩ. The detector incorporated is a solid-state device which makes use of a low noise f.e.t. chopping and demodulating system where the zero is constant irrespective of load resistance. Sensitivity on the linear range is 50-0-50μV, with an input impedance of 35kΩ. In the position of minimum sensitivity the log range requires an input of 100mV to give full scale deflection on the output meter. An overload of 75V d.c. can be tolerated by the input circuits. Noise level is less than 1μV peak to peak. Two batteries of the PP9 type supply the current. Croydon Precision Instrument Company, Hampton Road, Croydon, Surrey CR9 2RU.

WW318 for further details

Frequency-period meters

Covering the h.f., v.h.f. and u.h.f. radio bands the new models 9057 and 9059 from Racal Instruments are frequency-period meters, suited for maintenance and servicing applications. Model 9059 has an input sensitivity of 10mV and covers a frequency range from 10Hz to 560MHz, while the 9057 covers from 10Hz to 80MHz. Period measurements between 10Hz and 3MHz

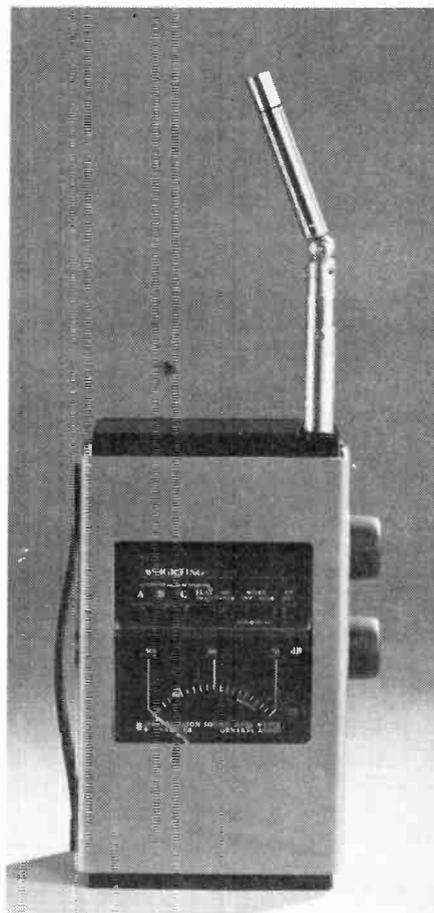


can be made with both instruments. Measurements are displayed by a six-digit in-line readout of l.e.d. indicators. The display is fully latched and an overflow indicator is included. The internal frequency standard is a temperature controlled crystal oscillator. Operating from internal batteries these instruments provide up to five hours continuous working. The battery power pack contains its own charging circuit. An a.c. power pack is available which will operate from supplies of any frequency from 45 to 440Hz. Racal Instruments Ltd, Duke Street, Windsor, Berks, SL4 1SB.

WW316 for further details

Sound level meter

General Radio model 1933 is a portable sound level meter, impulse/impact and octave-band analyser. A-, B- and C-weighted sound level measurements are pushbutton selectable and measurements are read from a linear meter scale which displays the sound level in dB over a 20dB



range. Octave band analysis requires one adjustment for correct readings since the input signal level and optimized signal-to-noise ratio are adjusted automatically. A telescoping microphone mast helps to eliminate the need for tripods and extension cables. Sound level measurements can be made from 10 to 150dB, A-, B- or C-weighted. Frequency is 5Hz to 100kHz and there are 10 octave bands with centre-frequencies from 31.5Hz to 16kHz. General Radio Company (U.K.) Ltd, Bourne End, Bucks. SL8 5AT.

WW321 for further details

Literature Received

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

ACTIVE DEVICES

The May 1972 Celdis stock catalogue includes information on Motorola semiconductors and ranges of Mullard capacitors, Celdis Ltd, 37/39 Loverock Road, Reading, Berks. WW401

Concorde Instrument Company, 28 Cricklewood Broadway, London NW2 have sent us part one of their o.e.m. price list covering 2,700 types of semiconductor WW402

We have received a catalogue describing components, connectors, power supplies, soldering irons, indicator lights etc from REL Equipment and Components, Croft House, Bancroft, Hitchin, Herts. WW403

The Erie "U.K.72" catalogue gives details of the full range of electronic components distributed by Erie Electronics Ltd, South Denes, Great Yarmouth, Norfolk WW404

Prices and brief specifications of the J1000 range of logic modules for industrial application are included in a leaflet sent by Jasmin Electronics Ltd, Station Road, Quorn, Leicestershire LE12 8BP ... WW405

Device data, design features, device selection criteria and typical circuit design and performance data for microwave devices are given in "data package". RCA Ltd Solid State-Europe, Sunbury-on-Thames, Middlesex WW406

We have received two leaflets describing products manufactured by Souriau Lectropon Ltd, Shirley Avenue, Vale Road, Windsor, Berks.

Connectors and cables WW407
Components WW408

Products described in the catalogue "Microwave Tubes and Devices" include travelling-wave tubes, backward-wave oscillators, magnetrons and microwave components. The M-O Valve Co. Ltd, Brook Green Works, London W6 WW409

Two data sheets specifying fully decoded r.a.m. i.c.s manufactured by General Instrument Microelectronics Ltd, 57/61 Mortimer Street, London W1N 7TD describe:—

Device RA-9-1103; 1024 bit dynamic memory WW410
Device RA-9-1101A; 256 bit memory WW411

"Your hundred best f.e.t.s" leaflet gives brief specifications of f.e.t.s distributed by GDS (Sales) Ltd, Michaelmas House, Salt Hill, Bath Road, Slough, Bucks. WW412

Parameters for over 100 types of epoxy-encapsulated f.e.t.s and bipolar transistors are given in a leaflet from Siliconix Ltd, Saunders Way, Sketty, Swansea, Glam. SA2 8BA WW443

PASSIVE DEVICES

Two data sheets received contain specifications of two types of resistors manufactured by Morganite Resistors Ltd, Bede Industrial Estate, Jarrow, Co. Durham, NE32 3EN.

M1-03-1 describes the FC65 metal film resistor WW413
M1-06 describes Filmet resistor modules WW414

Low torque 900 Series V4 rotary switches are the subject of data sheet 1029 from Honeywell Ltd, Charles Square, Bracknell, Berks. WW415

We have received the May 1972 edition of "Switch", which includes information on relays, timers and microswitches. Blore-Barton (Switching Division), Blore-Barton Ltd, Reedham House, Burnham, Bucks. WW416

Two advance technical data sheets, describing transistor power supply transformers, have been sent to us by Gardners Transformers Ltd, Christchurch, Hampshire.

Sheet A.T.24 describes the Solo series transformers (low aspect ratio and low external field) WW417

Sheet A.T.25 — transformers for low voltage d.c. supplies WW418

Main ratings of capacitors, potentiometers and carbon film resistors, which are distributed by LST Electronic Components Ltd, 7 Coptfold Road, Brentwood, Essex, are given in the 1972 LST industrial catalogue WW419

Two miniature carbon resistors (types RKL2 and RKL10) are described in a data sheet sent to us by Kubuk Ltd, P.O. Box 306, Shoreham-by-Sea, Sussex WW420

APPLICATION NOTES

"Theory and operational functions of silicon tunnel diodes" is the title of application note G.201. Joseph Lucas (Electrical) Ltd, Electronics Product Group, Mere Green Road, Sutton Coldfield Warwicks ... WW421

We have received two Ithaco application notes from Techmaton Ltd, 58 Edgware Way, Edgware, Middx HA8 8JP.

IAN-22: "The logarithmic lock-in amplifier" WW422

IAN-101: "Phase and amplitude response of a variable electronic filter" WW423

Three issues of "Instrumentation Talk", published by 3M United Kingdom Ltd, 3M House, Wigmore Street, London W1A 1ET, are:—

Vol. I, No. 1; "The handling and storage of instrumentation tape" WW424

Vol. I, No. 2; "The durability testing of instrumentation recording tape" ... WW425

Vol. II, No. 1; "Bulk degaussing — a.c. erasure" WW426

EQUIPMENT

A 5V, 5A power supply, type L.C.REG, for logic systems, is the subject of a data sheet from Weir Electronics Ltd, Durban Road, Bognor Regis, Sussex WW427

We have received three component and equipment catalogues from Marconi Instruments Ltd, Sanders Division, P.O. Box No. 10, Gunnels Wood Road, Stevenage, Herts. SG1 2AU.

"Flexible Waveguide & Flanges" WW428

"Microwave Component Catalogue" . WW429

"Concise Catalogue of Microwave Equipment" WW430

Sent to us by T.E.M. Sales Ltd, Gatwick Road, Crawley, Sussex, RH10 2RG — a short-form catalogue (products include chart recorders, medical instruments and analysers for air and water pollution) WW432

The Model 71A laser scanning analyser is a fault inspection system specified in a data sheet received from the Measurement and Inspection Department, Ferranti Ltd, Thornybank, Dalkeith, Midlothian WW433

Catalogue Ref. B4/72/5 describes the range of hi-fi accessories from Bib Sales, P.O. Box 78, Hemel Hempstead, Herts WW434

A leaflet describes a 6in bench type wave solder machine manufactured by Jidenco Products Ltd, Vale Road, Windsor, Berks WW435

Specifications for a respiratory flow meter, type R1, are given in a data sheet sent to us by Birchover Instruments, 84/85 Bancroft, Hitchin, Herts. SG5 1NQ WW436

"Microwave Basic Experiments II" is a booklet containing product descriptions and experiments for familiarization with microwave equipment manufactured by Sivvers Lab, Box 420 18, S-126 12 Stockholm 42, Sweden WW437

Livingstone Hire Ltd, Shirley House, 27 Camden Road, London NW1 9NR, have a new catalogue with additions including a range of battery operated bench equipment WW440

Preset and variable d.c. power supplies and hybrid d.c. regulators are described in the 1972-3 product catalogue from Lambda Electronics, Marshlands Road, Farlington, Portsmouth PO6 1ST .. WW441

We have received a leaflet describing the Series 80 Imrak equipment racking system. Imhof-Bedco Ltd, Ashley Works, Cowley Mill Road, Uxbridge, Middx. WW442

"Catalogue 99" gives details of integrated packaging, "Wire-Wrap" products and a range of i.c. accessories. Cambion Electronic Products Ltd, Castleton, Nr. Sheffield, S30 2WR WW444

A group delay measuring set M1010 is described in a leaflet we have received. Cole Electronics Ltd, Lansdowne Road, Croydon, CR9 2HB ... WW445

GENERAL INFORMATION

Five publications have been sent to us by the British Standards Institution, 2 Park Street, London W1A 2BS.

Publication PD 5686:1972 "The use of SI units" Price 85p

BS9305 N001:1972 "Silicon voltage regulator diode, 1.5W, 3.3V to 33V (5%) with hermetically sealed encapsulation" Price 85p

BS9305 N042:1972 "Silicon voltage regulator diode, 1.5W, 6.8V to 200V (5%) with hermetically sealed encapsulation" Price 85p

BS4808 Parts 1 and 2:1972 "L.F. cables and wires with p.v.c. insulation and p.v.c. sheath for telecommunication" Price 60p each

Thomas & Betts Ltd, Greenhill House, 90/93 Cowcross Street, London EC1M 6JR, have issued a publication describing a cable harnessing system, markers for wire identification, conduits and wire terminals WW438

Reference to over 300 types of products and services, from fire alarm systems to language laboratories, is made in a new publication from EMI Electronics and Industrial Operations, Blyth Road, Hayes Middx WW439

"Palmaplan: an original concept in printed circuitry" is the title of a publication describing a p.c. board design system. Palma Electronics Ltd, 9 Arkwright Road, Poyle, Colnbrook, Bucks WW440

Constructing radio receiving aerials for v.h.f. is the subject of Information Sheet 1104(2) sent to us by the Engineering Information Department, B.B.C., Broadcasting House, London W1A 4WW.